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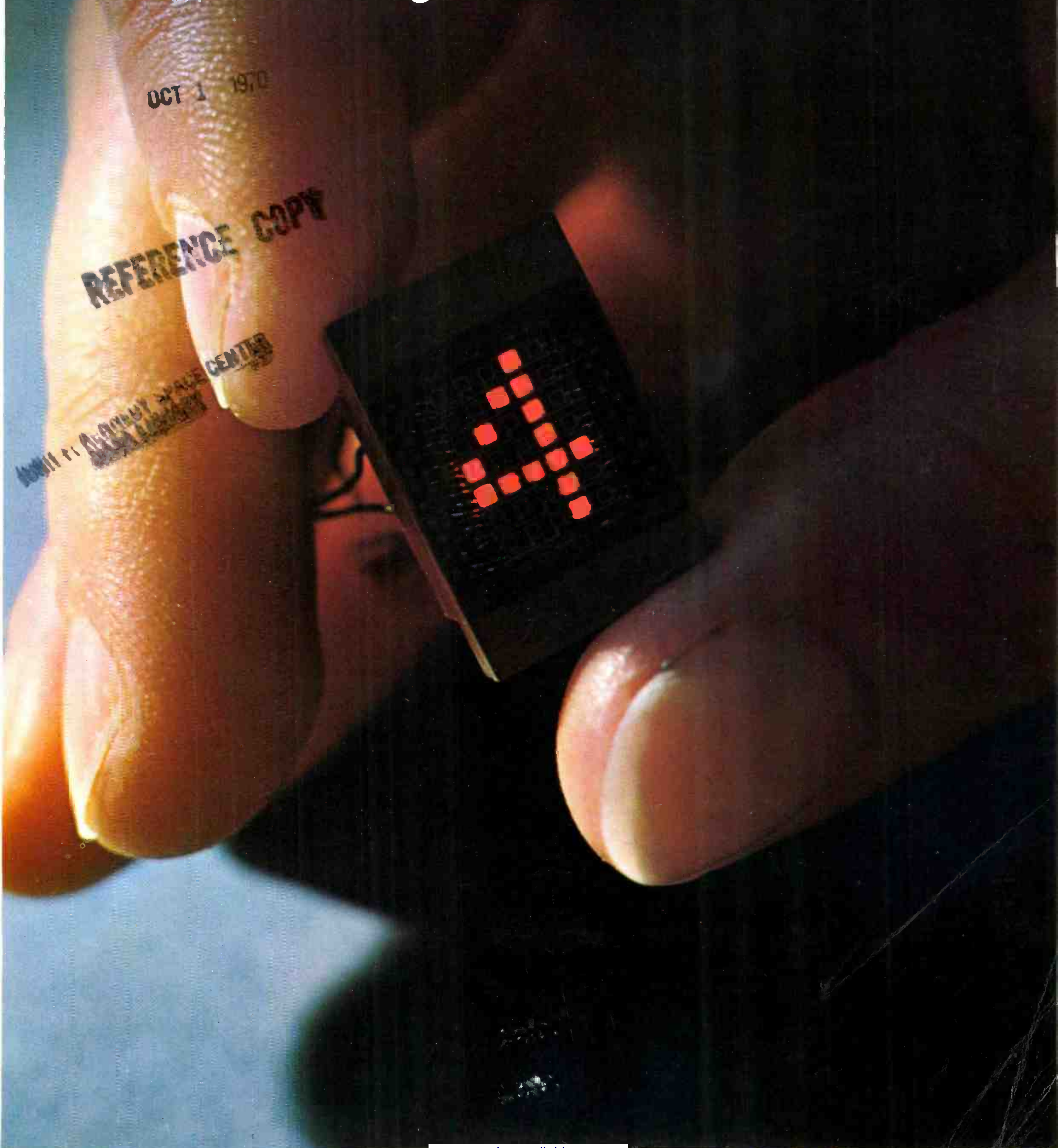

Wireless World

Electronics, Television, Radio, Audio

Sixtieth year of publication

September 1970 **Volume 76 Number 1419**

Contents

Our cover picture this month, which might be called "two kinds of digits", shows a five-by-seven array of light-emitting diodes made by Standard Telecommunications Laboratories. Display devices are discussed on page 444.

IN OUR NEXT ISSUE

The first of a series of articles on the **elements of linear i.cs**

Constructional details of a **100-watt quality amplifier**

Review of some of the latest techniques in **domestic sound and television receivers**

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Wireless World

The I.C. Industry—who will pick up the bits?

When Dr. F. E. Jones, managing director of Mullard Ltd, Britain's largest manufacturer of integrated circuits, admits that his company is losing £1M per year on i.c. production and says that he "cannot see his way through the dense undergrowth of the i.c. world" it is time for us all to sit up and take notice. In particular the Government should heed his words for if it does not Britain's integrated circuit industry will be doomed.

The problem is simply one of price. Take for example one of the simple integrated circuits, the t.t.l., quad, two-input gate. This device requires 28 separate manufacturing processes and consists of a chip of silicon about 1.5mm square containing 24 components, four of which are transistors. At the beginning of 1969 the selling price of this device was 15s 6d, by the end of the year it had fallen to 5s 6d but, at the present time, it must be marketed at between 1s 2d and 1s 8d—which is well below manufacturing costs—in order to compete with American concerns.

How can the Americans sell at this price in the U.K. when Dr. Jones has said that he cannot see any way to produce the circuit for this sort of money now, or in one year's time or even five years' time no matter what production methods are used?

The Americans, of course, have a huge home market, most of which is a closed shop to outsiders because of the "Buy American Act". During 1969 this market absorbed 413.4M integrated circuits, of which about 342M were digital. In the same period the U.K. produced only 12M integrated circuits. Also the Americans have a huge investment in the Far East where they employ 20,000 to 30,000 workers. It would appear that the silicon chips are made in America and sent by air to the Far East where they are assembled into packages by workers paid about 8d per hour.

If the finished products are taken back to the U.S. for distribution, tax is paid only on the added value. But what is the difference in value between a processed silicon chip and the same chip in a package? Because of the difficulty in assessing this added value the amount paid in tax could be negligible. When asked if the U.K. could not mount such an "offshore" operation to their advantage Dr. Jones said that the cost savings produced in this way would be nullified by our own import duty which is based on the *selling price* and not the added value. Even taking into account the obvious advantages the Americans have Dr. Jones said that he still could not see how they achieved such a low price.

It is obvious that either the Americans are selling off their surplus production in this country or they are making a deliberate attempt to undermine our own integrated circuit industry. The thought is frightening when one considers all the industries which employ integrated circuits and would therefore have to rely entirely on American sources of supply.

Before the Americans can be accused of dumping it has to be proved that they are selling components at a price which is lower than in the country of origin. But what *is* the country of origin when the various processes are carried out in different parts of the world?

If British component manufacturers wish to export to America they have to prove to the American excise authority, before the goods are shipped, that the price takes into account shipping costs and is not less than the price in this country. British manufacturers have no such protection where American goods are concerned. They cannot go on sustaining such huge losses and prompt action is needed from the Government in the form of import controls if the integrated circuit industry is to survive.

In view of the gravity of this threat to the U.K. semiconductor industry, and indeed the whole of the electronics industry, it is surprising that no reference is made to the prevailing situation in the latest "statistical survey" of the industry issued by the Electronics "Little Neddy" on the day Dr. Jones made his announcement.

Phase-locked Stereo Decoder

Improved channel separation and low distortion obtained using an inductorless circuit

by R. T. Portus, A.M.I.E.E. and A. J. Haywood*, A.M.I.E.E.*

During a stereo broadcast the f.m. carrier transmitted by the B.B.C. is composed of three parts, as shown in Fig. 1. The components are:

1. Left plus right (L + R) forming the compatible mono signal containing frequencies up to 15 kHz.
2. Left minus right (L − R) which amplitude modulates a 38 kHz carrier. (The carrier is suppressed to better than 1% in order to make full use of the maximum deviation available at the transmitter.)
3. A low level "pilot tone" at half the carrier frequency, i.e. at 19 kHz, whose zero crossing points are coincident with those of alternate cycles of the 38 kHz suppressed subcarrier. This tone is used as a reference to regenerate the suppressed subcarrier at the receiver.

Stereo decoders

There are many forms of decoders,[1] the most popular being the switching type where the 19 kHz pilot tone is filtered by a tuned circuit, frequency doubled (to 38 kHz) and used to switch the complex signal—as shown in Fig. 2. Appendix III explains why complete channel separation is not achieved by such simple switching.

Limitations of tuned-circuit decoders

Poor separation results if the derived 38 kHz switching signal is modulated by any extraneous signal. This means that all the stereo information has to be removed by the tuned circuit. Also, if the output from the 19 kHz tuned amplifier contains any frequencies which pass through subharmonics of 38 kHz, the frequency multiplication stage will produce a modulation of the recovered 38 kHz. This in turn will produce sum and difference frequencies ("birdies") when switching the incoming complex signal. From the above considerations it may be seen that a high-Q tuned circuit is required. On the other hand, if the Q is made very large the phase of the recovered subcarrier becomes extremely drift sensitive. Any change in phase will reduce the separation. The effect of uncompensated phase error on separation is as follows.

* Rolls-Royce Ltd.

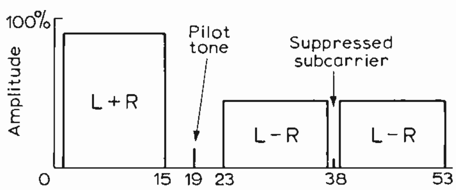


Fig. 1. Frequency spectrum of a stereo multiplex signal.

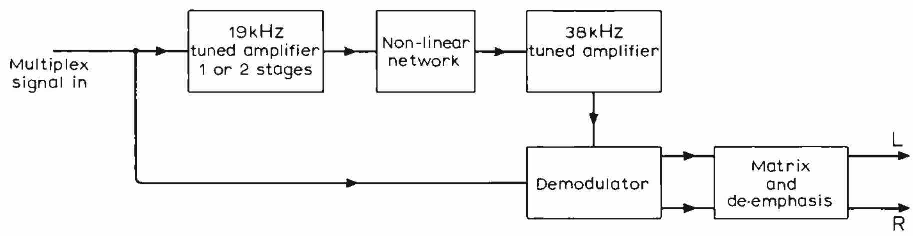


Fig. 2. Block diagram of switching stereo decoder.

Phase error at 19 kHz (deg.)	Separation (dB)
1	82·5
2½	54·5
5	42
10	30
15	23

Fig. 3 shows the change for small changes in tuned circuit L or C with various values of Q.

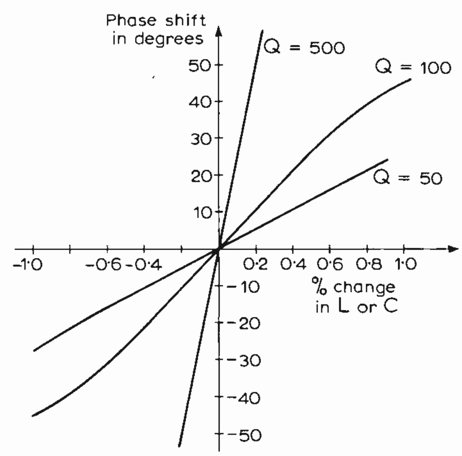


Fig. 3. Phase changes due to variation of L or C in a parallel tuned circuit.

Phase-locked decoder

Because of the shortcomings of tuned decoders the authors decided to use a phase locked loop to regenerate the 38 kHz suppressed subcarrier. A phase-locked loop is used to lock a locally generated signal in phase with an input signal whose frequency is liable to vary. Such a system may be given a very narrow bandwidth so that noise components of the input signal will not affect the loop operation.

If a locally generated 38 kHz signal is binary divided the remaining 19 kHz may be phase locked to the "pilot tone". This means that alternate zero crossings of the 38 kHz signal are coincident with those of the 19 kHz pilot tone. The locally generated 38 kHz signal is therefore in phase with the suppressed subcarrier.

The performance of such a system may be made superior to that of decoders using tuned circuits for the following reasons:

1. A phase-locked oscillator is a closed-loop system and so changes in component values, due to ageing, temperature etc., are corrected. Other systems have no comparison between input and output, and so errors may only be reduced by careful matching, the use of high quality components and accurate setting up.
2. The generation of "birdies" is very much reduced because the loop is given a low bandwidth, ensuring that phase modulation of the 38 kHz switching signal can only occur at low frequencies.
3. The setting up of a phase-locked loop is a simple matter, a d.c. voltmeter being the only instrument used. For other decoders an oscilloscope is usually required.

Performance of phase-locked decoder

The decoder (British patent application No. 35600/69) was tested using a Radiometer stereo signal generator, and the following specification obtained.

Frequency	Separation
80 Hz	28 dB
1 kHz	45 dB
10 kHz	40 dB

In a decoder fitted with a variable matrix the separation was 56 dB at 1 kHz.

Distortion at 1 kHz and with full modulation is 0·3% (predominantly 2nd harmonic), and the distortion introduced by crosstalk at 1 kHz is 67 dB down at full modulation. The mono and stereo gains are within 1 dB of each other. Cancellation is 45 dB. The input voltage range is 130 mV r.m.s. to 1·3 V r.m.s., and the output voltage 250 mV r.m.s.

The separation at 80 Hz could be improved by using matched components in the p.s.d. stage, but directional information is negligible at this frequency so that component matching is hardly worthwhile.

Operation of a phase-locked loop

The phase-locked loop (Fig. 4) can be looked upon as a servo-amplifier in which the 19 kHz pilot tone is used as an input reference phase, and the servo loop used to control a local oscillator in a fixed phase relationship. If there is a fixed phase between two signals then they must be at the same frequency. A phase-locked subcarrier may be regenerated by deriving the 19 kHz from 38 kHz by use of a binary divider. Alternate zero crossings of the 38 kHz signal are therefore held locked to the incoming 19 kHz.

The operation of the loop is as follows. The balanced phase sensitive detector (p.s.d.) produces a d.c. output proportional to the difference in phase between the input frequency (f_{in}) and a locally derived frequency (f_o), plus higher frequency components produced by the chopping of the input signal. For a balanced p.s.d. the lowest frequency present is that produced by the highest audio frequency (15 kHz) beating with the 19 kHz chopping signal. The difference signal is 4 kHz (19 kHz − 15 kHz). This error signal is then amplified and low pass filtered by the error amplifier. The filtering removes the high frequency signals produced by the balanced modulator. The error amplifier output is used to control a voltage controlled oscillator (v.c.o.).

If the p.s.d. is not perfectly balanced small low-frequency audio signals will be present at the p.s.d. output. If the frequency of these signals approaches the loop bandwidth they will not be completely filtered out by the error amplifier. The resulting "ripple" on the error amplifier output will cause "jitter" on the v.c.o. output (i.e. phase modulation).

Loop operation is such that a change in phase between (f_{in} and f_o (e.g. due to drift)

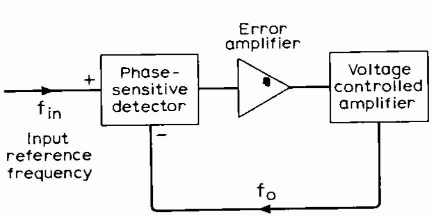


Fig. 4. Block diagram of phase-locked loop.

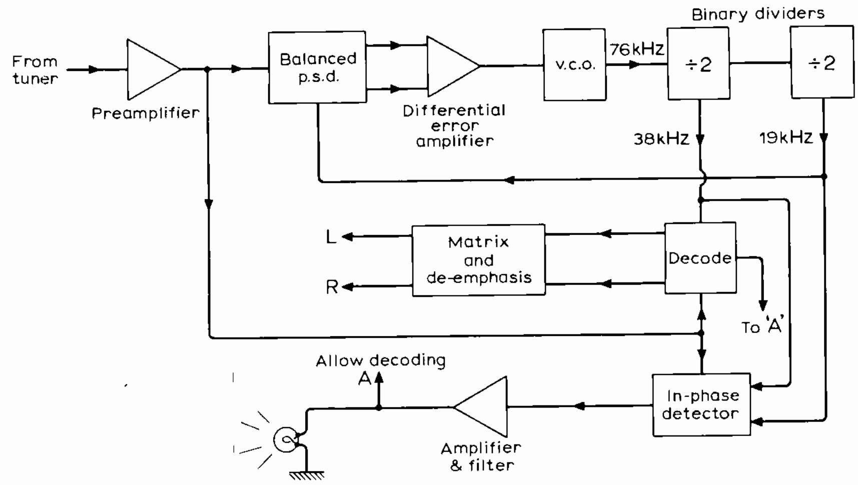


Fig. 5. Complete stereo decoder system built round the phase-locked loop.

causes the output of the p.s.d. to drive the v.c.o. in such a direction as to hold the phase of f_{in} locked to that of f_o.

The system behaves as a servo-loop with a bandwidth determined by the time constant of the filter in the error amplifier.

When deciding on loop bandwidth two points have to be considered. These are "jitter" on the v.c.o. and the time required to pull into lock. The first point is important because phase jitter will reduce channel separation. Jitter is minimized by having a low loop bandwidth.

On the other hand if loop bandwidth is made too small the loop will take a long time to pull into lock.

A compromise must therefore be made and a bandwidth of 25 Hz was found to be adequate.

Theory and design equations for the loop are given in appendices I and II.

Practical decoder system

In Fig. 5 the pre-amplifier acts as a buffer on the tuner output. Its gain is made variable so that a fixed level of pilot tone is presented to the loop independent of the tuner output level.

The phase detector is balanced, and the error amplifier is made differential in order to maintain balance. The output from the error amplifier constitutes the control voltage of the voltage controlled oscillator (v.c.o.) which is free running at approximately 76 kHz.

By the use of binary dividers square-wave outputs are obtained at 38 kHz and 19 kHz. The former output drives the decoding transistors and the latter the loop p.s.d. transistors. The decoder outputs are matrixed and de-emphasized to provide the left and right channel outputs.

On acquiring lock an output is obtained from the in-phase detector. This output is amplified and filtered, and used to allow decoding to begin, and to drive a stereo indicator lamp, if required.

Circuit description

The complete decoder circuitry is shown in Fig. 6. Tr_1 and Tr_2 comprise a pre-amplifier whose output level is adjusted by RV_1 to set the p.s.d. gain constant. The p.s.d. consists of R_5, R_7, Tr_3 and Tr_4, and these transistors are used in the inverted mode to minimize offset voltage. Since the bases of Tr_3 and Tr_4 are driven alternately at 19 kHz the p.s.d. and differential error amplifier IC_1 form a double-sideband suppressed-carrier modulator, the output of which is modified by the error amplifier response.

Audio signals close to the loop bandwidth are attenuated by C_3. The phase shift due to C_3 ($10\frac{1}{2}°$) is corrected for in the matrix which is fully described in a later section.

The error amplifier IC_1, type U6E7709393 or similar, is a low cost differential inte-integrated operational amplifier. This is chosen for its high gain and low input offset voltage. At ±6 V supply the typical open loop gain is 72 dB.

C_4 and C_6 provide an additional h.f. roll-off beyond the loop unity gain frequency so that loop stability is not affected.

The amplifier response is tailored to give an overall loop bandwidth of 25 Hz with an error response damping factor of 0·707. The d.c. gain is chosen from a consideration of the static error due to v.c.o. drift. The error is <1° at 19 kHz.

RV_2 provides an adjustment for input offset voltage, which would otherwise appear as a phase error. The low input offset voltage temperature drift of IC_1 ensures good phase stability.

The v.c.o. (Tr_6 and Tr_7) is a conventional astable multivibrator with the timing voltage of the bases controlled by IC_1. It is arranged to free run at approximately 76 kHz. RV_4 provides a fine frequency control by adjustment of the mark/space ratio.

R_{25} and D_1 ensure that the v.c.o. will not operate above 80 kHz, by clamping the error amplifier input to approximately +0·6 V.

The 76 kHz signal is then applied to IC_2 which is a dual D-type SN7474 or similar binary divider. IC_2 will then provide a 38 kHz and a 19 kHz switching signal with a 1 : 1 mark/space ratio. The 19 kHz signal is applied to the loop p.s.d. transistors Tr_3 and Tr_4, thus closing the loop. The 38 kHz signal is applied to the decoding transistors Tr_{11} and Tr_{12}.

The bases of all the switching transistors are pulled negative to remove stored charge and to hold the transistors off in the presence of audio signals.

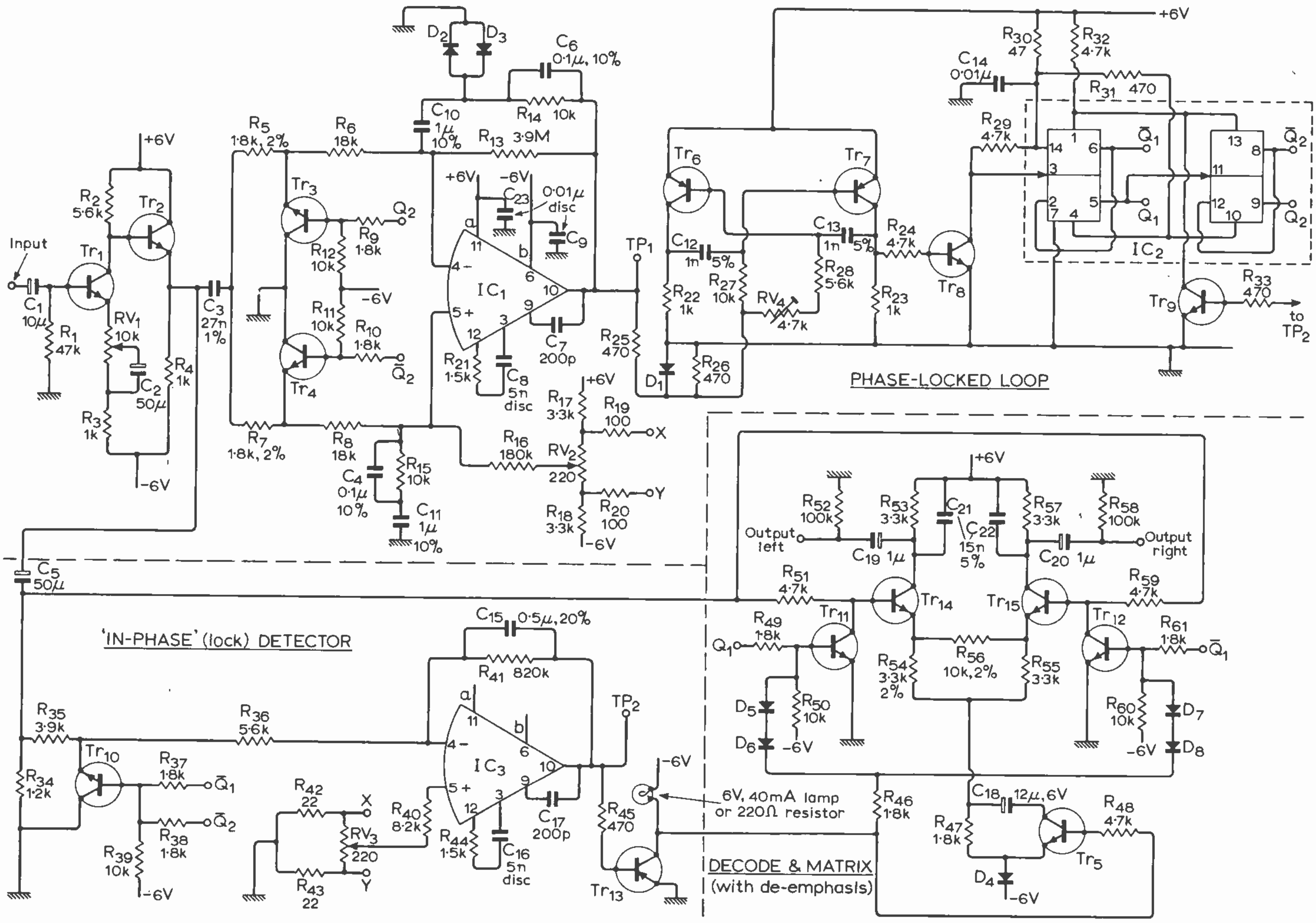
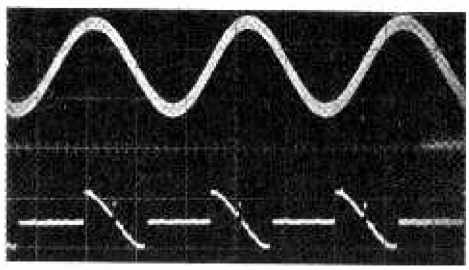


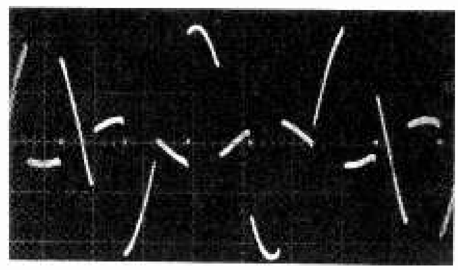
Fig. 6. Circuit diagram of decoder. Integrated circuits IC_1 and IC_3 are type U6E770 9393 or equivalent, and IC_2 type SN7474N or equivalent. Transistors can be BC108 or equivalent for n-p-n types, and ZTX500 or equivalent for p-n-p. Diodes are 1S44 or equivalent. Resistors can be ¼W 5% tolerance unless specified otherwise.

The action of D_2, D_3 is to speed up lock-in at switch-on for outputs greater than $\pm 0{\cdot}6$ V. The diodes shunt current away from C_{10} and effectively increase loop bandwidth. Between $\pm 0{\cdot}6$ V the diodes have no effect.

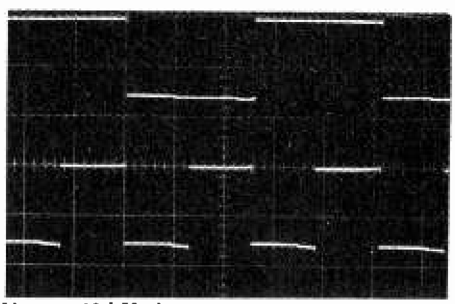
In-phase detector. Transistor Tr_{10} is driven by both binary dividers so that it is off for ¼ cycle only of the 19 kHz. Thus a pulsed d.c. component is produced when the loop is locked. This signal is filtered and amplified by IC_3 and used for the following purposes:
1. To turn on a stereo indicator lamp, if required.
2. To allow decoding to begin in Tr_{11} and Tr_{12}. If the decoder is allowed to switch during mono transmission, supersonic noise around 38 kHz will be heterodyned into the audio bandwidth causing a deterioration in s/n ratio. Also from some transmitters a 23 kHz low-level tone is present during mono transmission. This tone would produce a "beat" at 15 kHz in a free running decoder. For this reason the v.c.o. is not



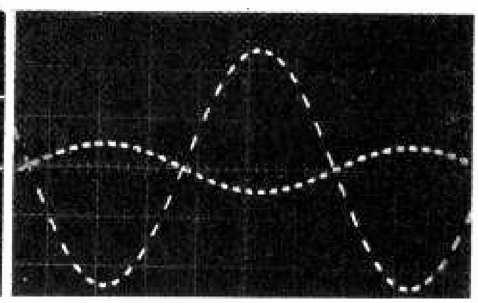
Upper: pilot tone + 150 Hz (A = B).
Lower: emitter of Tr_3.



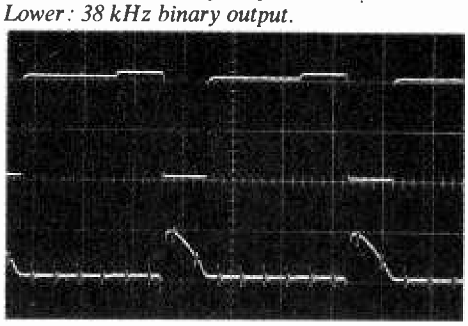
Matrixing of 15 kHz signal (A = B). No de-emphasis. Pilot omitted.



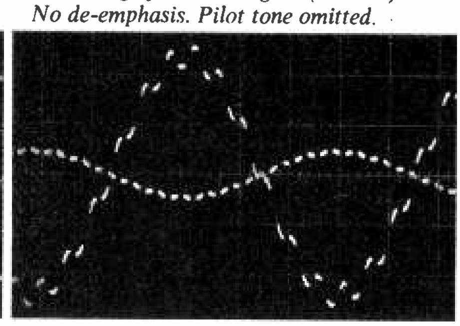
Upper: 19 kHz binary output.
Lower: 38 kHz binary output.



Matrixing of 1·5 kHz signal (A = B). No de-emphasis. Pilot tone omitted.



Upper: Tr_{10} base drive.
Lower: Tr_{10} emitter.



As above with 19 kHz pilot tone included.

Fig. 7. Oscillograms of typical waveforms generated by the locked-loop.

allowed to operate above 80 kHz, i.e. the binary output cannot exceed 20 kHz.

3. To ensure that the l.h. and r.h. channels do not reverse in the presence of noise etc. If the loop initially locks in the wrong direction i.e. l.h. and r.h. channels are reversed, a positive output will be produced by the in-phase detector, Tr_9 will be turned on thus clearing the binaries, and reversing l.h. and r.h. This will also be the case if a line transient flips one binary only. In the writers' experience the loop has never locked in the wrong direction. Typical waveforms of the loop in operation are given in Fig. 7.

4. To keep the mono and stereo gains approximately constant via Tr_5.

Matrix. Once the signal has been decoded by Tr_{11} and Tr_{12} it is matrixed and de-emphasized via Tr_{14} and Tr_{15}. Matrixing is necessary in any switching decoder as may be seen from the theory given in appendix III.

Power supply. The circuit of Fig. 6 is shown for a supply of ± 6 V. Fig. 8(a) shows a suitable low cost regulator using a miniature mains transformer.

Single supply operation is accomplished from an unregulated (15 V minimum) supply by altering the "earth" rail from the 0 V to the -6 V line as shown in Fig. 8(b).

In this case there are two minor circuit alterations: capacitor C_1 is reversed; and resistors R_{52} and R_{58} are returned to the former -6 V rail i.e. instead of the former 0 V rail. The signal "earth" line becomes, of course, the former -6 V line.

Setting up procedure

The decoder is set up using a d.c. voltmeter. The stages are as follows:

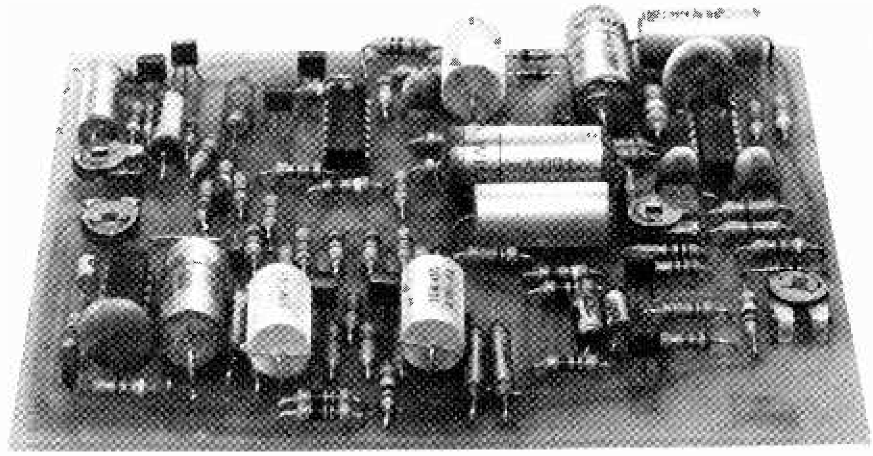
1. With no input signal adjust RV_2 and RV_3 for 0 V at test points TP_1 and TP_2 respectively.
2. Apply a stereo signal and adjust RV_1 until a negative voltage 1·5 V appear at TP_2 (Adjust RV_4 if necessary).
3. Adjust RV_4 to bring TP_1 back to zero.
4. Adjust RV_1 to bring TP_2 to -1·5 V.

Operation

Several decoders have been built and tested by the authors, using both single and dual supply operation.

Over the last year they have proved to be reliable and have required no adjustment beyond the initial setting up procedure.

The high inherent separation, in particular at high frequencies, produces a marked improvement in performance over that obtained with more conventional decoders.



Decoder components mounted on a printed circuit board approximately 100 mm × 125 mm.

Notes

This decoder design arose from work done on a phase-locked loop frequency multiplier made to improve the resolution of an engine tachometer.

The authors are grateful to J. W. Hill and D. L. Lynas for the use of equipment and for helpful criticism, and thank Rolls-Royce Ltd for permission to publish.

A kit of parts including a printed circuit board is available, mail order only from Integrex Ltd, P.O. Box 45, Derby, DE1 1TW. The price is £8 19s 6d plus 2s 6d for carriage and packing. This printed circuit board is adapted to single or dual supply options. The size of the board is 100 mm × 125 mm (approx.) and it may be seen from the photograph that a high packing density has been achieved. To simplify assembly, component positions have been marked and the track is roller tinned. A power supply kit is also available.

REFERENCE

1. "Receiving Stereo Broadcasts", *Wireless World*, September 1966.

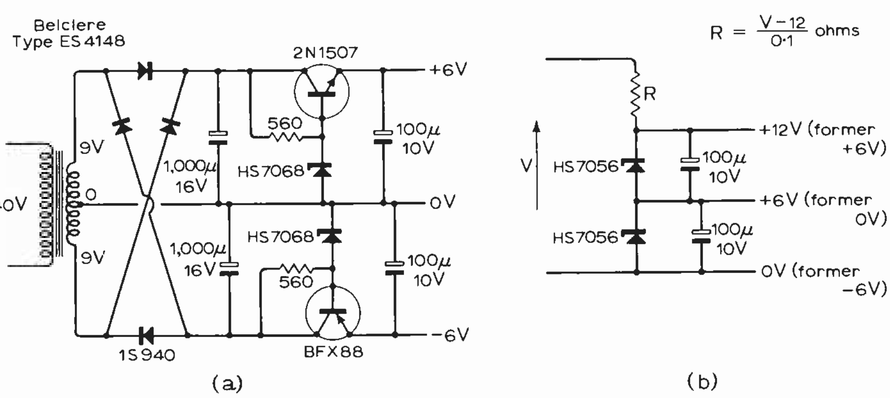


Fig. 8. Decoder power supply arrangements. (a) Stabilized supply for ± 6 V. (b) Supply obtained from an unregulated 15 V (or greater) d.c. source.

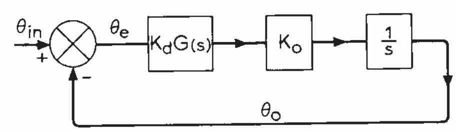
Appendix I

Loop Theory

The constants used are defined as follows:

K_d = phase detector gain constant in volts/radian.
K_o = v.c.o. gain constant in radians/s/V ($=2\times$ Hz/V).
$G(s)$ = error amplifier transfer function
θ_{in} = input reference phase angle.
θ_o = locally generated phase angle
θ_e = phase angle error $(=\theta_{in}-\theta_o)$.

The loop block diagram is redrawn below.



The integration term $(1/s)$ is introduced by the conversion of the frequency of the v.c.o. to its phase.

By inspection of the diagram

$$\frac{\theta_o}{\theta_{in}} = \frac{K_oK_dG(s)\,1/s}{1+K_oK_dG(s)\,1/s} = \frac{K_oK_dG(s)}{s+K_oK_dG(s)} \quad (1)$$

The error amplifier has a response that can be shown to be

$$G(s) = \frac{A(1+sT_2)}{1+s(T_1+T_2)} \quad (2)$$

where

$$A = \frac{R_1}{R_3};\quad T_1 = CR_1;\quad T_2 = CR_2$$

Substituting (2) in (1) we get

$$\frac{\theta_o}{\theta_{in}} = \frac{[K_oK_dA/(T_1+T_2)](1+sT_2)}{s^2+\dfrac{sK_oK_dAT_2}{T_1+T_2}+\dfrac{K_oK_dA}{T_1+T_2}} \quad (3)$$

Where it is assumed $K_oK_dAT_2 \gg 1$.

This is of the form

$$\frac{\theta_o}{\theta_{in}} = \frac{2\eta\omega_n s+\omega_n^2}{s^2+2\eta\omega_n s+\omega_n^2}$$

where using servo terminology ω_n = natural (undamped) frequency of the loop and η = loop damping factor.

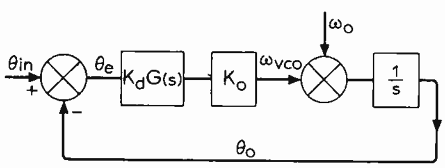
$$\text{Thus} \quad \omega_n = \left(\frac{K_o K_d A}{T_1 + T_2}\right)^{\frac{1}{2}} \qquad (4)$$

$$\eta = \frac{T_2}{2}\left(\frac{K_o K_d A}{T_1 + T_2}\right)^{\frac{1}{2}} = \frac{T_2}{2}\,\omega_n \qquad (5)$$

Static phase errors

Since the phase loop contains an integrator term then the steady state phase error will be reduced to zero if the v.c.o. free-running frequency is the same as that of the input frequency.

The response of the loop to a disturbance causing drift in the v.c.o. free running frequency can be obtained from the figure below.



where

$\omega_{v.c.o.}$ = free running v.c.o. frequency

ω_o = disturbance (or drift) frequency.

It can be seen that

$$\frac{\omega_{v.c.o.}}{\omega_o} = \frac{1/s.K_o K_d G(s)}{1 + 1/s.K_o K_d G(s)} \qquad (6)$$

By referring $\omega_{v.c.o.}$ back to an equivalent θ_e we get

$$\frac{\theta_e}{\omega_o} = \frac{1/s}{1 + 1/sK_o.K_d.G(s)} = \frac{1}{s + K_o K_d G(s)} \qquad (7)$$

For a step input of $\omega_o = \Delta\omega_o$ then

$$\theta_e.s = \frac{\Delta\omega_o}{s}.\frac{1}{s + K_o K_d G(s)} \qquad (8)$$

Applying the final value theorem we get

$$\text{steady state phase error} = \frac{\Delta\omega_o}{K_o K_d A}$$

$$(\text{since } G(0) = A) \qquad (9)$$

This phase error is required in order to pull the v.c.o. away from its free-running frequency. By making $K_o K_d A$ sufficiently large the phase error can be made as small as required.

For those who wish to pursue phase-locked loop theory a very good account is given in "Phaselock Technique" by Floyd M. Gardner, published by John Wiley and Sons Inc.

Appendix II

Loop equation constants

P.S.D. gain constant. For a phase error γ, when the loop is locked the mean output voltage from the phase detector, V_o, is given by:

$$V_o = \frac{V_{in}}{\pi}\int_{\frac{\pi}{2}+\gamma}^{\frac{3\pi}{2}+\gamma} \sin\theta \, d\theta$$

where $V_{in} \sin\theta$ is the input signal.

Thus $|V_o| \approx \frac{2}{\pi} \times V_{in} \times \gamma$ where γ is small.

Thus, if we choose an input signal of 200 mV peak to peak amplitude:

$$K_d = 63{\cdot}7 \text{ mV/radian}$$

$$\text{or} \quad \underline{K_d = 63{\cdot}7 \times 10^{-3} \text{ V/radian}}$$

Steady state error. From equation (9) it may be seen that the steady state error is given by:

$$\text{Error} = \frac{\Delta\omega_o}{K_o K_d A}$$

where $\Delta\omega_o$ is the frequency difference of the free running v.c.o. and the pilot signal, K_d is the phase detector gain constant and K_o is the v.c.o. gain constant.

Now

$K_d = 63{\cdot}7 \times 10^{-3}$ V/rad and

$K_o = 1{\cdot}2 \times 10^3 \times 2\pi$ rad/s/V (measured)

For the v.c.o., the frequency drift is approx. 0·075 %/°C. It follows that for a steady state error of 1° over 20°C temperature change.

$$\underline{A = 214}$$

From equation (4)

$$\omega_n = \frac{K_o K_d A}{T_1 + T_2} \qquad T_1 + T_2 = 4{\cdot}16 \text{ s}$$

From equation (5)

$$\eta = \frac{T_2 \omega_n}{2} \therefore \quad T_2 = 9 \text{ ms}$$

Appendix III

Matrixing in switching decoders

In an f.m. stereo broadcast the instantaneous deviation of the transmitter is given as:

$$f = 0{\cdot}9\left[\tfrac{1}{2}(A+B) + \tfrac{1}{2}(A-B)\sin\omega_s t + 0{\cdot}1 \sin\frac{\omega_s}{2}t\right]$$

where $\frac{\omega_s}{2\pi} = 38$ kHz

The pilot tone is $0{\cdot}1 \sin\frac{\omega_s}{2}t$.

In a switching decoder the input signal is treated as if it were time division multiplexed. The signal is multiplied by a square wave:

$$= \left(\frac{1}{2} + \frac{2}{\pi}\sin\omega_s t + \frac{2}{3\pi}\sin\omega_s t + \ldots\right)$$

Suppose there is phase error ϕ existing between the 38 kHz carrier used for modulation and the local 38 kHz decoding signal.

The output signal will be proportional to V_1 where:

$$V_1 = [(A+B) + (A-B)\sin\omega_s t]\left[\frac{1}{2} + \frac{2}{\pi}\sin(\omega_s t + \phi) + \frac{2}{3\pi}\sin(3\omega_s t + \phi) + \cdots\right]$$

$$\left(\text{neglecting } \sin\frac{\omega_s}{2}t\right)$$

$$= \left(\frac{1}{2} + \frac{1}{\pi}\cos\phi\right)A + \left(\frac{1}{2} - \frac{1}{\pi}\cos\phi\right)B$$

plus modulation around ω_s, $3\omega_s$, etc, which is of the form:

$$V_1 \propto (A + \Delta B)$$

where Δ is dependent on the phase shift ϕ.

If the decoding signal is shifted by 180°, the output signal may be shown to be proportional to V_2 where: $V_2 \propto (B + \Delta A)$.

The purpose of matrixing is to subtract ΔV_2 from V_1 and vice-versa.

$$\text{i.e. } V_1 - \Delta V_2 \propto A + \Delta B - \Delta B - \Delta\Delta A$$

$$\propto A - \Delta\Delta A$$

It may be seen that signals from channel A only are present in one output and from channel B in the other.

Thus, if the phase shift ϕ is known, and hence Δ, complete separation of an ideal broadcast is achieved.

Announcements

A lecture on the **British Trans-Arctic Expedition 1968/69** will be given by Sqn. Ldr. F. W. Church, D. J. Collins and R. G. Shears at a meeting of the R.S.G.B. at 18.30 on 28th September at the I.E.E., Savoy Place, London W.C.2.

A post-graduate evening course of sixteen meetings entitled **"Integrated Circuit Electronics"** will be held at North East London Polytechnic, Romford Road, London E.15. beginning 22nd October. Fee £4.

"Single Standard Colour Television Receivers" is the title of a 6-week course of evening lectures to be held at Norwood Technical College, Knight's Hill, London S.E.27, commencing 20th October. Fee £1.

Basic Electronics is the subject of two ten-evening courses comprising lectures and practical work to be held at Twickenham College of Technology, commencing 15th October and 21st January. Further details from The Principal, Twickenham College of Technology, Egerton Road, Twickenham, Middx. Fee 5gn per course.

A course in **Modern Sound Studio Techniques** is to be held through the winter at the Northern Polytechnic, Holloway Road, London N.7. Sessions will be on Thursday evenings beginning 29th October. The fee for the fifteen sessions is 6gn.

The scope of the **annual Manchester exhibition** of measuring equipment is to be extended next year to include professional-grade products in three main groups: electronic equipment, electronic components and apparatus for industrial measurement and control. The show will be held at the City Hall, Deansgate, from 5th to 8th October, 1971. Organizers: Industrial Exhibitions Ltd, 9 Argyll Street, London W1V 2HA.

Following the acquisition of **Painton & Co.**, of Northampton, by the Plessey Co., the Resistor and Connector Divisions of Plessey have combined with Painton in whose name the business will in future be conducted with R. W. Addie continuing as managing director. Painton now have four operating divisions, three commercial—connectors, resistors and exports—and one production.

Television Wobbulator

2. Circuit Details

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

In last month's article the general principles of a wobbulator specially designed for the alignment of television i.f. amplifiers were discussed. The complete circuit diagram of the instrument is shown in Figs. 1 and 2. The heart of it is the section labelled "Wobbly Oscillator" with the transistor Tr_2. This is an h.f. type operating as a Colpitt's oscillator.

The inductor is L_1 connected between the collector and the earth line, which is the *positive* of the 17-V power supply. The base is earthed for r.f. through the 1-nF capacitor C_5 and the capacitors C_7 and C_6, of 6.8pF and 82pF respectively, are between collector and base with the emitter joined to their junction. These two form the split capacitance of a Colpitt's oscillator. The tuning capacitance proper comprises D_2 in series with C_8, D_2 being the varactor diode the capacitance of which is varied by a 50-Hz voltage applied through R_9. The purpose of C_8 is merely to prevent L_1 from shorting D_2 at 50Hz.

The emitter resistor R_8 completes the d.c. path of Tr_2 but allows the emitter to be free at radio frequency. It is effectively in shunt with the base-emitter path of the transistor. The other resistor R_7 and the zener diode Z_1 give a stabilized supply of some 12V for the oscillator.

Wound in bifilar fashion with L_1 is L_2 which feeds the diode D_1 with load R_3 returned to a potential divider R_1,R_2 across the 17-V supply. The diode load capacitance is C_2 with a filter R_4, C_3. The rectified output of the diode is positive-going and is applied in series with the small bias voltage from the potential divider to the base of Tr_1 which acts as a d.c. amplifier with emitter resistance R_6. Its collector is joined to the base of Tr_2 and fed through R_5. In Fig. 1, this stage is labelled a.a.c, for automatic amplitude control, and it forms a kind of a.g.c. system whose purpose is to minimize variations of the amplitude of oscillation of Tr_2 as its frequency is varied.

If, for any reason, the amplitude increases

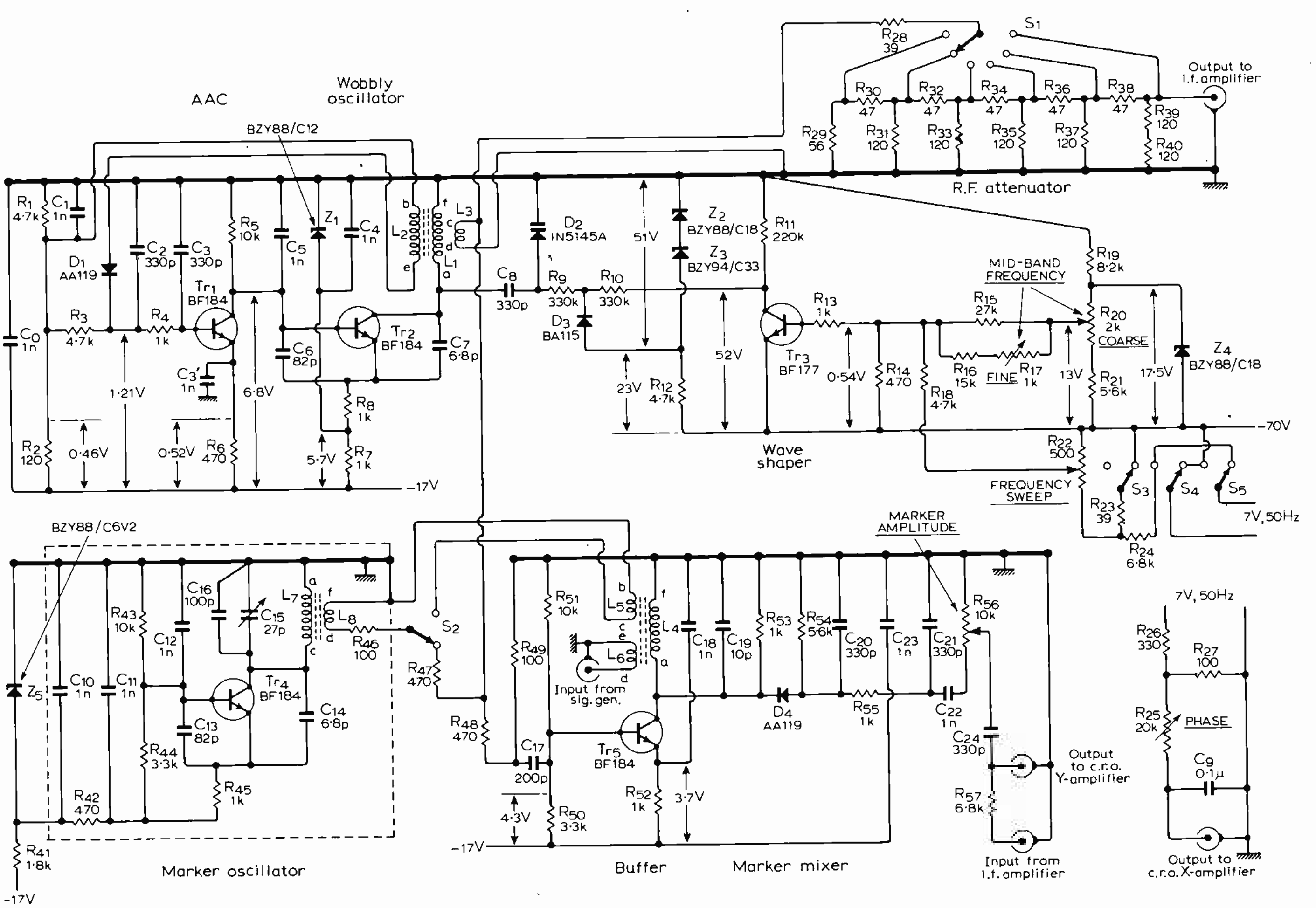


Fig. 1. Complete circuit diagram of the wobbulator, apart from the power supply. Tr_2 generates a signal varying in frequency between 30.5MHz and 42.5MHz which is fed to the amplifier under test through the attenuator controlled by S_1. Tr_3 is the wave shaper which enables a linear relation between frequency and displacement on the c.r.o. to be obtained.

the rectified output of D_1 increases and Tr_1 passes more current. Consequently the voltage drop across R_5 increases and the base-emitter voltage of Tr_1 is reduced and this reduces the amplitude of oscillation. The net result is to reduce the magnitude of any changes in the amplitude of oscillation.

The output of the oscillator is taken from the single-turn coil L_3 coupled to L_1 and connected to the r.f. attenuator. This comprises a ladder network of resistors with a single-pole 6-way switch S_1. When terminated at the output socket by 75Ω this network has a constant impedance of 37.5Ω at the switch arm for any position of the switch and, because of R_{28}, a constant impedance of 75Ω is presented to the coupling coil L_3. The actual impedance is slightly different because R_{28} actually has the standard value of 39Ω. Further, resistor tolerances in the network affect the impedance relations slightly.

Viewed from the output socket the impedance varies with the switch position, but the variation is appreciable only between the full output and the next-to-full output positions. This is inevitable with a simple ladder attenuator. The attenuator is intended merely to adjust the output to the required level in steps of about 5dB. It is not intended to provide precisely equal and known steps; that would entail a more elaborate construction and the use of precision resistors.

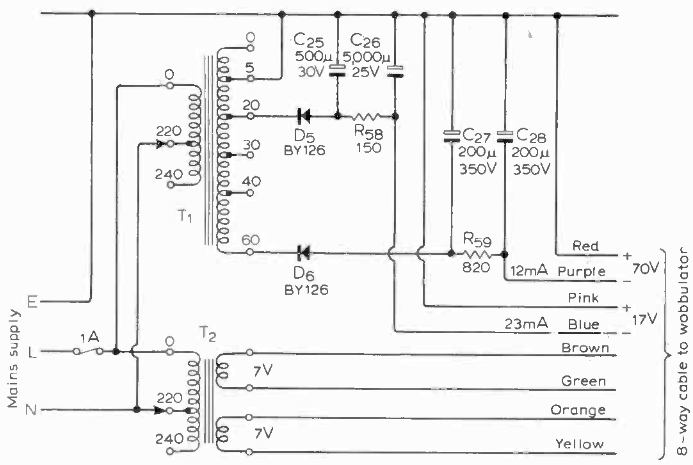
The total attenuation available is not more than some 30dB. More may be needed with some i.f. amplifiers and it is recommended that any such extra attenuation be provided externally. For this purpose, the Belling-Lee coaxial attenuators type L729 are particularly convenient and two 6-dB types (L729/6) and one 12-dB (L729/12) in conjunction with the internal attenuator should be sufficient for almost any amplifier.

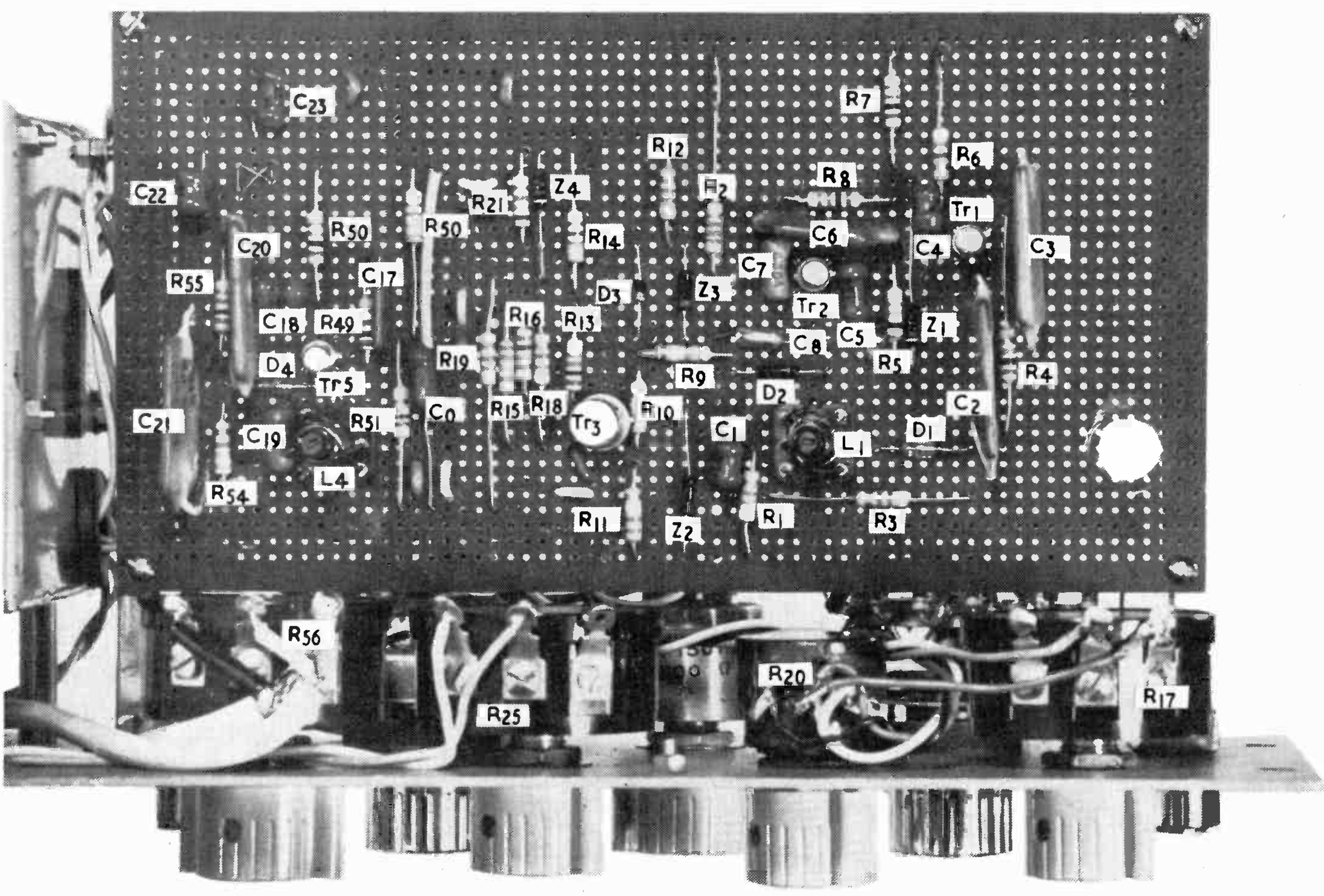
We now come to the wave shaper, Tr_3. It was explained in Part 1 that the relation between the frequency generated by the oscillator and the voltage applied to the varactor diode D_2 is logarithmic and that to obtain a linear relation between frequency and the X-deflection voltage of the oscilloscope it is necessary for the voltage applied to D_2 to have an exponential relation to the deflecting voltage. This exponential relation is provided by Tr_3.

A BF177 transistor is used. This is primarily intended for use as a video output stage and is rated for operation at 100V. It is used here on a 70-V supply, its collector being taken to earth (positive) through R_{11} of 220kΩ and its collector voltage being fed to D_2 through R_9 and R_{10} of 330kΩ each. The emitter is returned to the negative of the 70-V supply. The collector potential with respect to earth can thus vary from zero (collector current cut-off) to almost −70V (bottoming).

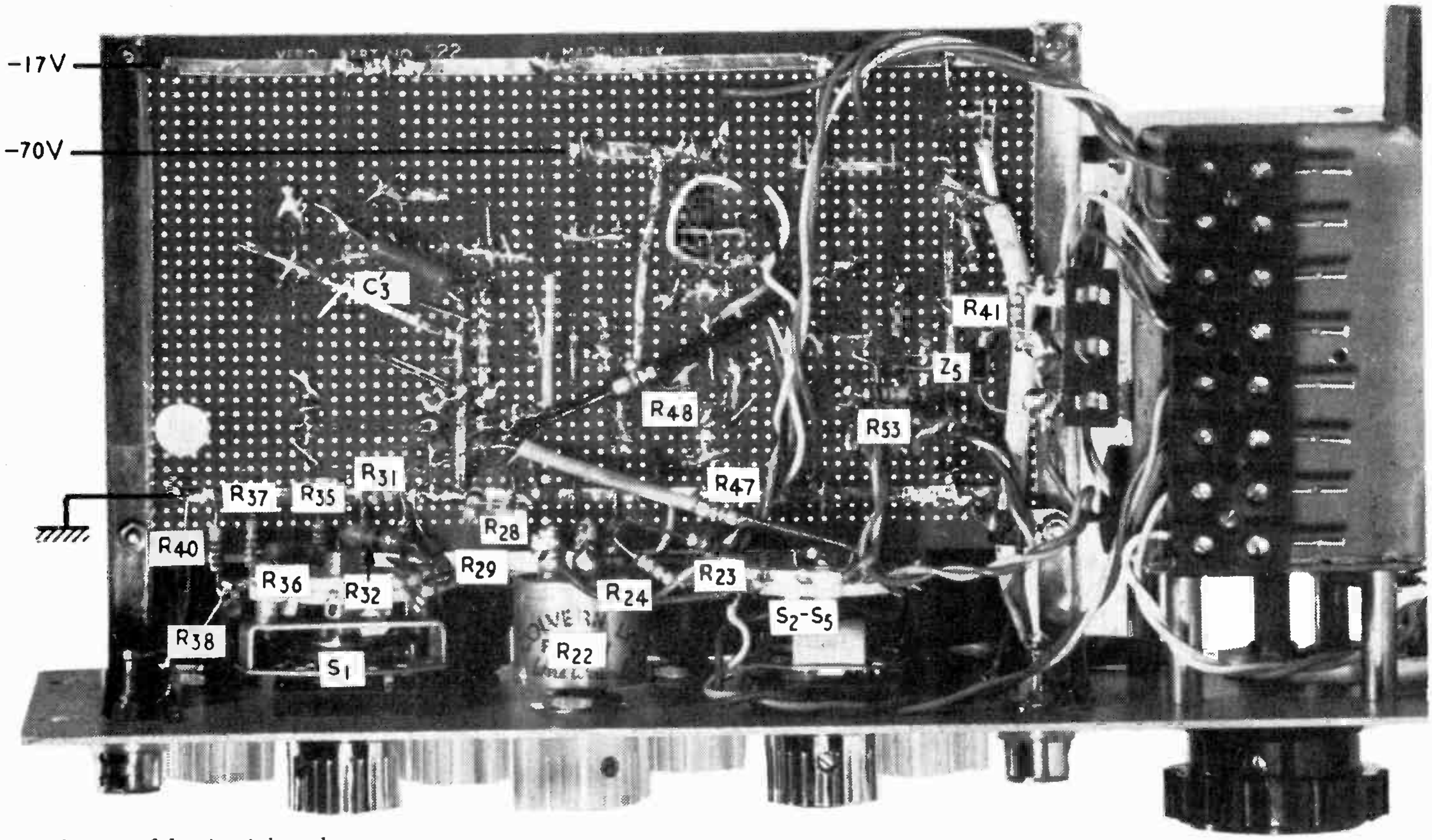
The varactor diode D_2 is rated for a maximum of 60V. The application to it of a higher voltage is thus dangerous to its life and as it is an expensive component it is

Fig. 2. Circuit diagram of the power supply. Two transformers are used since the requirements are then met by stock-item components.





Top of the circuit board showing the placement of components.



Underview of the circuit board.

important to protect it from the accidental application of anything more than 60V. The protective device comprises the diode D_3 with R_{12} and the two zener diodes Z_2 and Z_3. The zener diodes keep the anode of D_3 at 51V ($\pm 5\%$) negative to the earth line irrespective of variations of the supply voltage or the tolerance of R_{12}.

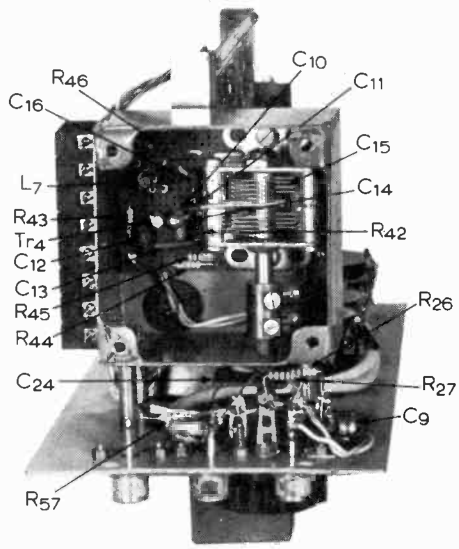
As long as the junction of R_9 and R_{10} is less negative to earth than this D_2 is non-conductive and has no effect. When the junction becomes slightly more negative than the zener voltage D_2 conducts and clamps the junction of R_9 and R_{10} at about 0.7V more than the voltage across Z_2 and Z_3. Even if Tr_3 is bottomed, therefore, the voltage applied to D_2 is limited to 54.35V, allowing for 0.8V drop across D_3 and for 5% high tolerance on the zener diodes. Two zener diodes in series are used instead of a single one to reduce the dissipation per diode. It is thought that this should reduce the risk of a diode going open-circuit, for that would put the protective circuit out of action.

The d.c. bias for Tr_3 comes from the 70-V supply. A stabilized supply of 18V is first derived from R_{19} and Z_4. A variable voltage of 13.3V to 18V is available from the slider of the "Coarse mid-band frequency control" R_{20}. Roughly 1/20 of this voltage is applied to the base of Tr_3 by a potential divider comprising R_{14} and the network R_{15}, R_{16} and R_{17}, the last of which is variable to form a "Fine mid-band frequency control". This is essential when the equipment is used to provide a narrow-band sweep for sound-channel alignment. It is not essential with the wide-band sweep for vision channel use, but even then it is convenient.

A 7-V, 50-Hz, supply from a mains transformer is applied through a phase-reversing switch S_4, S_5 to a potential divider R_{22}, R_{24}. The maximum voltage available from R_{22} is 0.48V r.m.s. and roughly 1/10 of this is applied to the base of Tr_3 through the potential divider R_{14}, R_{18}. The voltage actually applied to the transistor is thus variable from zero to a maximum of about 130mV peak-to-peak. For the narrow sweep for the sound channel S_3 connects 39Ω in shunt with R_{22}, and this reduces all voltages to about 1/12.

Another 7-V, 50-Hz, supply provides the X-deflection voltage for the oscilloscope. It is reduced to about 1.63V r.m.s. by R_{26}, R_{27}; this is about 4.7V peak-to-peak. It is sufficient with the oscilloscope used to give an X-trace of about two-thirds of the screen width and the X-expansion control on the oscilloscope permits the length of trace to be increased beyond this as desired. If the oscilloscope used does not have such a control, then R_{26}, R_{27} should be replaced by a variable potentiometer. Of course, if the oscilloscope used needs more than some 20V p-p for full X-deflection, a higher supply than 7V r.m.s. must be provided.

The voltage across R_{27} is applied through a simple variable integrator R_{25} and C_9 to provide an X-deflection voltage which is lagging in phase on the input, the phase being adjustable by R_{25}. This is necessary to correct for phase shift in the Y-deflection circuits. This occurs mainly in the wave shaper and is caused principally by C_8 with R_9 and R_{10}. Some also may occur in the receiver under test, especially if the output from it is taken from a video stage. The capacitance of D_3 also has an effect and, indeed, the addition of only 10pF across D_3 produces a noticeable phase shift. This may seem surprising in view of the frequency of 50Hz, for few realize how sensitive circuits are to phase shift.



Interior of the marker oscillator.

It is important to keep stray capacitance low in the collector circuit of Tr_3, for the waveform is not sinusoidal and so a phase shift introduced here cannot be completely corrected. It is found in practice that the residual error can be no more than the thickness of the trace on the c.r.t., which is quite adequate for all normal requirements.

This is the basic wobbulator. On one half-cycle of the sine-wave deflection the oscillator frequency varies from 30.5MHz to 42.5MHz, or such smaller range as may be set. On the next half-cycle it varies from 42.5MHz to 30.5MHz and, of course, the X-deflection varies in the opposite direction.

The traces for increasing frequency and for decreasing superpose to give a single visible trace only if the X and Y channels have the same phase shift and only if the i.f. amplifier gives the same response to increasing and decreasing frequencies. The latter does not necessarily occur, and it will not if the sweep repetition rate is too great in relation to the bandwidth of the amplifier. With wideband amplifiers, such as television amplifiers, the effect is unlikely to be observed with a 50-Hz sweep except, possibly, at the troughs introduced outside the passband by high-Q trap circuits. Even there, however, they have not so far been observed.

It is essential to have at least one marker on the oscilloscope trace to indicate frequency and it is a convenience to have two markers. An internal marker oscillator is provided and is Tr_4. Its circuit is substantially the same as that of the wobbly oscillator, but it is tuned by an ordinary variable capacitor C_{15} of 27pF. It operates at 6.2V from a supply stabilized by Z_5. It is built into a screening box, the zener diode and R_{41} being outside it to reduce the total dissipation within the box and so the temperature rise within it. The output is taken from a 2-turn coil L_8 through a 100-Ω resistor R_{46} and thence by a twisted pair of wires to S_2 and to L_5 or R_{47}. The earthy lead from L_8 is earthed at one point only, close to where it emerges from the screening box of the marker oscillator. This is important.

A socket for an input from a signal generator is provided on the front panel and is connected by a pair of twisted wires to L_6. The only earth point is that of the coaxial socket on the front panel. Again, this is important.

With the marker system used here it is desirable that r.f. from the signal generator or from the internal marker oscillator should not reach the input of the i.f. amplifier. Unless each coupling circuit is earthed at one point only, it has been found impossible to prevent greatly excessive leakage. Even with the single-point earthing, there is still some leakage and with the marker amplitude control turned right down traces of the markers are visible on the screen. They are not at all troublesome in normal usage, however.

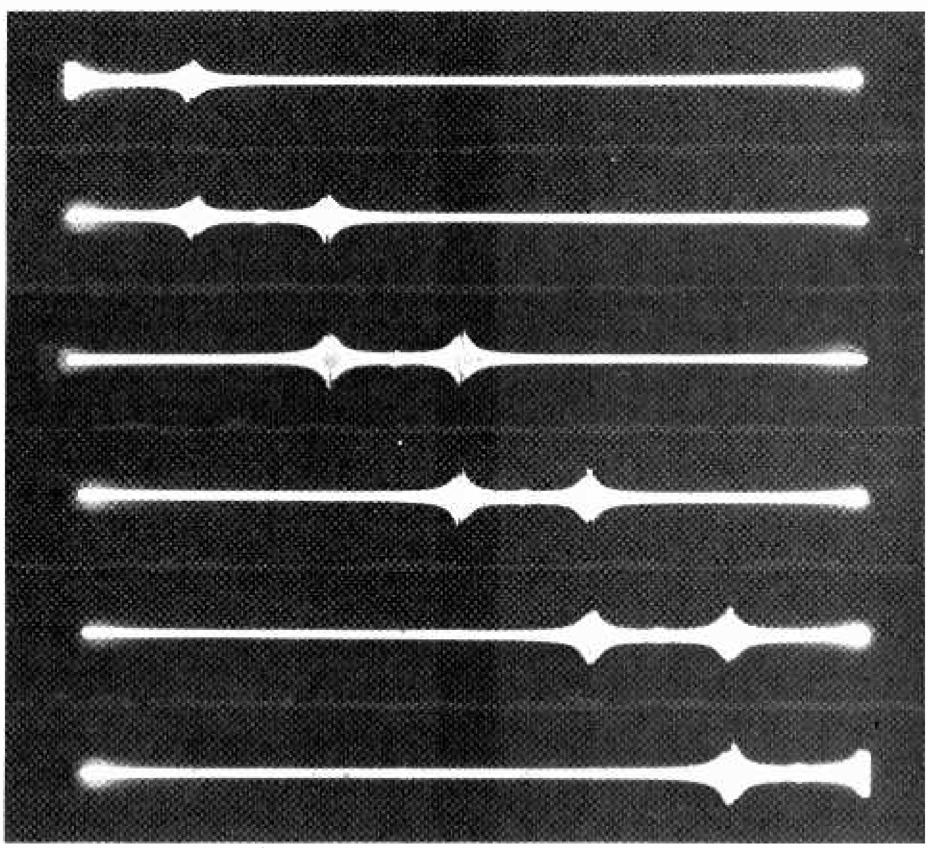
The internal marker oscillator has a second use. When S_2 connects R_{46} to R_{47} the output is connected to L_3 and so to the input of the attenuator. It is thus mixed directly with the output of the wobbly oscillator. The switches S_2 to S_5 are ganged together; they are actually a rotary switch wafer having 4-poles, each 2-way. When S_2 connects the marker output to the attenuator it also brings R_{23} into circuit to reduce the sweep and by S_4 and S_5 it reverses the phase of the 50-Hz voltage applied to Tr_3.

This is the condition for aligning an inter-carrier sound channel. The marker oscillator is set at 39.5MHz to simulate the vision carrier, and the wobbly oscillator is set by the mid-band frequency controls to sweep about a centre frequency of 33.5MHz. The 6-MHz beat between the two is developed in the receiver under test just as it is when a television signal is being received.

Reversal of the phase of the 50-Hz drive to Tr_3 is by no means essential, but is desirable in order that the frequency sweep shall be in the same direction for both channels. It is desirable to have frequency increasing to the right on the display and the connections are made to provide this for the vision channel. Because the sound-channel signal is generated as a difference frequency with the fixed frequency higher than the variable, the same connections would give a sound-channel display with frequency increasing to the left. This is remedied by reversing the phase of one 50-Hz channel.

We now come to the buffer and marker mixer, which is operative only for vision-channel alignment. L_3 is connected through a potential divider R_{48}, R_{49} to the input of Tr_5. This is a normal r.f. amplifier stage with a tuned collector circuit comprising L_4, C_{19} damped by R_{53}. It is tuned to mid-band and has a 3dB bandwidth of 12MHz. Marker inputs from the internal marker oscillator and from a signal generator are coupled into L_4 and there mixed with the signal from the wobbly oscillator. The whole is applied to the diode detector D_4 and the beats between the signals are produced at its output. The marker amplitude control R_{56} is fed through quite a small capacitor C_{22}. This has been found desirable to remove some residual hum which otherwise produced a slight unwanted vertical deflection on the c.r.t. The output from R_{56} is mixed with the output of the receiver under test, which necessitates bringing this into the wobbulator instead of taking it directly to the c.r.o.

This output is brought in at the socket "Input from i.f. amp" and is taken through R_{57} of 6.8kΩ to the socket marked "Output to c.r.o. Y-amp" and the marker signal is also fed to this through C_{24}.



Sequences of traces with the markers moved by 2MHz in each photo. The slight variation of trace length is caused by mains voltage fluctuation.

The maximum amplitude of the marker signals is fixed by the design and can only be reduced by R_{56}. The amplitude required depends on the peak-to-peak amplitude of the output of the i.f. amplifier. The maximum marker amplitude has been made about right for an i.f. output of about 2V p-p, and it is usable, although a little small, for outputs up to some 4V. Naturally, for small outputs of under 1V it is too great, and that is why the amplitude control is provided.

The newcomer to a wobbulator will undoubtedly think the amplitude available is too small, but he will soon find that it is adequate. It is essential that the marker amplitude should not be too great, since if it is it can greatly distort the trace.

The wobbulator is designed for use with an oscilloscope which has provision for an external X-deflection voltage and has some kind of amplitude adjustment for it. It must also have a Y-amplifier which is capable of giving undistorted reproduction of pulses of some 5ms duration which are repetitive at 100Hz, for that is what the output signal from the amplifier under test approximates to. An amplifier with a 3dB response at 50-Hz would be useless. the gain should be such that an input of 1V p-p will give full Y-deflection, and there must be adequate gain control.

The power supply is shown in Fig. 2 and is self-explanatory. The two 7-V windings of T_2 could quite well be on the core of T_1, but this would have necessitated a specially-wound transformer. By using separate transformers two standard components could be employed.

Half-wave rectification is used for both supplies with the diodes D_5 and D_6.

News of the Month

Television camera design for constructors

The latest and most ambitious design to be issued by the Mullard Educational Service is for a closed-circuit television camera using a 1-inch vidicon pick-up tube. According to the designers it can be built for about £45, which is less than half the price of the cheapest professional camera. A model we saw made up and working produced pictures which, although not of professional c.c.t.v. quality, were certainly good and would be acceptable to most schools, technical colleges and individual constructors.

Cost is saved by using a sub-specification vidicon of one of the simpler types (e.g. Mullard XQ1030, Philips 55850), by winding your own deflection and focus coils, by circuitry designed for cheap and readily available transistors (e.g. BCY70), by the use of a photographic camera lens rather than a television camera lens, and by an elementary type of housing (e.g. a piece of plastic drain pipe and a tobacco tin lid). In addition the camera circuitry is simplified by the adoption of sequential (non-interlaced) scanning: the model we saw demonstrated was working on 312½ lines per field, 50 fields per second.

The electronic circuits in the camera are: a video amplifier with a bandwidth of 4.5MHz and an output signal of 1V p-p into 75Ω; horizontal and vertical timebase generators; a synchronization mixer (feeding sync signals into the video amplifier); a blanking mixer (providing blanking pulses at the vidicon cathode); and a power unit (−15V for the transistor circuitry and −120V and +330V for the vidicon). The circuitry uses 23 transistors of six well-known types and 10 diodes of five types. Several monitors can be operated from the one camera.

To build the camera for £45 the designers admit that the constructor will have to "shop around" to some extent, but for those who are willing to pay for more to save this trouble, a complete kit of parts is available from Linstead Electronics, Roslyn Works, Roslyn Road, London N.15, at £70 (or £45 without the tube and lens). This company will also supply individual parts from the kit, and are offering a ready built camera with a professional looking chassis and case for £99 17s 6d. The Mullard Educational Service (Mullard House, Torrington Place, London W.C.1), of course, only supplies the design information. Their literature is not yet ready but will be available later.

The camera in use with its picture displayed on a monitor.

R.S.G.B. Exhibition

The Radio Society of Great Britain's International Radio, Engineering and Communications Exhibition will be held at the Royal Horticultural New Hall, Westminster, London S.W.1, from the 19th to 22nd of August. The exhibition will be open during this period from 10.00 to 21.00 and the admission charge is 4s.

At 14.30 on the second and third days there will be lectures on mobile equipment and s.s.b. receivers.

Wireless World will have a stand at the exhibition (No. 11) and we plan to demonstrate the phase-locked loop stereo decoder and the television wobbulator described in this issue. We also hope to demonstrate the surface temperature thermometer described in April 1969, the logic display aid described in the May to December 1969 issues and the electronic dice which was described in the April 1970 issue. Also on the stand we shall be selling a selection of books and reprints of some of the articles which have appeared in *Wireless World*.

Other exhibitors include: Adcola, Amateur Radio Shop, Angus McKenzie Hi-fi, Baginton Electronics, British Amateur Radio Teleprinter Group, British Amateur Television Club, City and Guilds of London Institute, E.M.S.A.C., J. Michael Gale, Garex Electronics, J. Beam Engineering Ltd, K. W. Electronics Ltd, Lowe Electronics, Microwave Modules Ltd, Ministry of Posts & Telecommunications, Nombrex (1969) Ltd, *Practical Wireless*, Professional & Academic Book Exhibition, Radio Amateur Invalid and Bedfast Club, Radio Shack Ltd, Radio and Space Research, R.A.F., Royal Navy A.R.S., Royal Signals A.R.S., Telecomms, Weller Electric Ltd, Western Electronics, World Association of Methodist Radio Amateurs and Clubs.

CAT-70

The British Amateur Television Club celebrated its 21st anniversary during the weekend 25th-26th July with a Conference on Amateur Television ("CAT-70") held at Churchill College, Cambridge.

Over 100 people registered, including delegates from Belgium, France, Germany, Switzerland and the U.S.A.

The major event of the weekend took place on the Saturday afternoon. Signals from several amateur television stations were received and displayed on monitors in the Wolfson Hall, from 14.30 until 16.00. Stations contributing included G6ADM/T from Haddenham, Cambs. (11 miles); G6NOX/T from Duddenhoe End, Essex (14 miles); G6REH/T from the top of his 110ft tower at Sutton St. James, Lincs. (36 miles); G6AEV/T/A from a caravan in Cambridge (2 miles); and G6WJ/T in Great Canfield, Essex, relayed via

G6NOX/T giving a total path length of 27 miles. A wide range of equipment was shown by amateurs, ranging from vidicon cameras to integrated circuit colour pattern generators.

The main social event was the Convention Dinner, attended by over 80 people on the Saturday evening.

On the Sunday morning, lectures were presented by C. Grant Dixon on "Slow Scan Television", by Arthur Critchley on "Integrated Circuits for the Amateur", and by M. P. Davies, of the Ministry of Posts & Telecommunications, on "The Amateur Licence".

Starting your own business

A conference to be held at Fulmer Grange, near Stoke Poges, from the 27th to 29th of November will appeal to all those with ideas and a hankering to start their own business. For this type of person, with the determination to make a go of it, the conference should be a weekend well spent.

Ten successful entrepreneurs will talk about their experiences, the difficulties they faced and how they overcame them. Other speakers will explain how to raise the necessary finance and how to assess the market. For a period the conference will break up into discussion groups and each one will deal with a particular new business. Finally the type of person most likely to succeed will be discussed.

The fee for the conference is £6 for a single man but the organizers realize that once the plunge is taken your wife is going to have to share your affections with the business and they feel that she should be in at the start. So for an extra £2 you can take her along as well! Application forms are available from The Meetings Officer, The Institute of Physics and the Physical Society, 47 Belgrave Square, London S.W.1.

Hong Kong radio sales halved in Britain

In contrast to the trend in the rest of the world sales of radio receivers manufactured in Hong Kong have fallen in this country from 34 to 17% of the total market. This means that Hong Kong has slipped to second position behind Japan as far as total U.K. sales is concerned. Hong Kong attributes this decline to their concentration on the six-transistor single waveband receiver for the American market which is not popular here.

These figures are taken from a survey conducted by the Hong Kong Development Council which says that the concentration on the American market results from the heavy U.S. investment in the industry. Eight American firms employ more than a third of Hong Kong's radio industry's 18,000 work force. Britain's only representative in the industry is Pye who has a 51% interest in Coronet Industries Ltd, a firm in Kowloon which employs about 600 workers.

Venus-Mercury spacecraft

The National Aeronautics and Space Administration have selected seven scientific investigations for the Mariner-Venus-Mercury 1973 (MVM-73) spacecraft to photograph the two planets, measure the particles and fields surrounding them and study their atmospheres and ionospheres. The 900-pound spacecraft is planned to be launched in the autumn of 1973 and should pass within 3,300 miles of Venus in February 1974 and 625 miles of Mercury in March 1974.

Teams of scientists have been formed to conduct the seven investigations listed below.

Imaging science: Weighing 61 pounds the experiment will employ two television cameras fitted with 1,500-mm Cassegrain telescopes. These will produce pictures of Mercury with a resolution similar to that of pictures of the Moon taken through Earth-based telescopes.

Because three 210-foot diameter aerials will be in operation in 1974 (two are now under construction at Tidbinbilla, Australia, and near Madrid, Spain), many of the pictures will be transmitted directly to Earth instead of being recorded for later transmission. Some 5,700 frames of Venus and about 2,740 frames of Mercury will be taken.

Radio science: Using radio signals from MVM-73's two transmitters (20-W S-band and 200-mV X-band) the spacecraft's trajectory will provide dual occultation of the radio signals at Mercury and a single occultation of the radio signals at Mercury and a single occultation at Venus. This will provide the investigators with information on the interplanetary phenomena during flight and, at planetary encounters, information on the atmospheres ionospheres, mass, radius and surface characteristics of both planets.

Plasma science: Will use an instrument called a scanning electronic analyser which is a set of hemispherical analyser plates and an electron multiplier mounted on a scan platform. It will measure ions from 80 to 8,000 eV and electrons from 40 to 400 eV. During the flight, the instrument will study the structure of the solar wind between the orbits of Earth and Mercury while the Pioneer-F and -G spacecraft, to be in flight during the same period, will be measuring the solar wind between the orbits of Earth and Jupiter. This will make possible a unique comparison of measurements at wide distances across the solar systems for the first time.

Magnetometer: An experiment to measure the magnetic field, it weighs 11 pounds and uses two triaxial, fluxgate magnetometers mounted at different distances from the spacecraft on one boom. The magnetometer team will use measurements near Venus and Mercury to make the first experimental study of the solar wind interaction with Mercury and to determine whether or not a magnetic field exists at the planet.

Ultraviolet spectrometer: Consisting of two ultraviolet grating spectrometers, direct studies of airglow and occultation techniques will be used to gather data. The spectrometer team will use the data obtained by solar occultation to search for the presence of an atmosphere on Mercury and to determine its structure and composition. Additional information on the character of the atmosphere of Venus will also be obtained.

Infrared radiometer: Consists of two one-inch-diameter telescopes each calibrated for a broad spectral band to measure temperature emissions from the two planets.

Charged particles: A six-pound instrument will measure electrons above 2keV and protons above 6keV.

3D colour TV

An experimental broadcast is to be made in Holland on November 9th using a principal which was applied long before the war in cinemas. The viewers will wear spectacles with one red and one green eyepiece so that they will see the programme in three dimensions and in colour.

Professional engineers, the choice

Members of the fourteen institutions which constitute the Council of Engineering Institutions can opt to join two other organizations which will handle such matters as welfare, pay, conditions and the like. An institution, being a learned society with the object of furthering science, cannot concern itself with the affairs of private individuals.

The engineer who wishes to have the full backing of a trade union, with all that this implies, can join the United Kingdom Association of Professional Engineers (UKAPE). This organization is intent on building a reputation for ethical and responsible conduct but it will also use its powers to protect the individual from exploitation. It aims to improve his conditions of employment and to regulate the relations between employer and employee.

Some engineers may not wish to join a trade union, maybe because they are employers themselves or for some other reason, but they would still like to belong to an organization with their welfare at heart. Such an organization is the newly formed Professional Engineers Association Ltd (PEAL). Members of the institutions within C.E.I. are being invited to join this organization and providing 15,000 do so PEAL say that they will be able to provide an efficient service. If PEAL is successfully launched the assets of the Engineers Guild, which was formed in 1938, will be made available to PEAL.

Tubes of Babel

If the current demand continues the number of public telephone circuits on Britain's trunk system will double in the next four years. In order to cope with this increase the Post Office have developed a 60-MHz cable which will provide 97,200 circuits. The first example of this cable will be laid in 1973 between London and Birmingham and will be followed shortly after by a similar cable in the reverse direction.

The cable will consist of nine coaxial pairs, or tubes, operating from 4 to 60MHz. Each pair will carry twelve 900-circuit broadbands using a frequency division multiplex system. Repeaters, energized over the cable from power feeding stations, will be placed every 1,500 metres to compensate for the high attenuation in the cable at 60MHz. The cable will be laid in a specially constructed concrete duct deeper than is normal for telephone cables and access will be through manholes.

C.E.I. changes its exam

Part 1 Of the Council of Engineering Institutions' examination—describable as "what every engineer should know"—has not so far been very successful in the radio-electronics sphere in the sense that few aspiring engineers have chosen to take it and even fewer have passed. To make it more attractive the C.E.I. has made some changes, both in subject matter and standard, which will come into force in 1972.

Originally Part 1 consisted of six compulsory subjects, of which only one, applied electricity, had to do with electro-technology, and the standard was equivalent to about half way through a three-year degree course. At present the electronics man can, if he wishes, replace three of the six (properties of materials, applied thermodynamics, fluid mechanics) with three subjects more closely related to electronics. The changes for 1972 will be the introduction of certain optional subjects and a reduction of the standard to a point about 12 months through a three-year degree course. Passes will be required in four compulsory subjects, mathematics, mechanics, properties of materials, and presentation of engineering information, and in two of the following optional subjects: electrotechnics, electronics, chemistry, thermodynamics.

Details of the changes, syllabuses and specimen papers are given in C.E.I. Statement No. 8 which may be obtained from the C.E.I. (2 Little Smith Street, London S.W.1) or any of its constituent institutions, price 4s 0d post free. Also under review is Part 2 of the exam, but any changes which may result will not come into effect until 1974 at the earliest.

Colour receiver sales

During June 38,000 colour television receivers were delivered to the trade. This is 2,000 more than in May and brings the total number of deliveries since the inauguration of the colour service in 1967 to half a million. The upward trend is illustrated by the fact that half of this total (0.25M receivers) have been delivered during the last nine months. The increasing sales of colour receivers is reflected in a slight slackening in sales of monochrome sets [1970 (1969), May 133,000 (149,000), June 124,000 (135,000)]. Single standard monochrome receivers accounted for 40% of the June total.

Experimental communications link

A 30km experimental communications link which uses a 50mm circular waveguide to transmit up to 300,000 simultaneous two-way telephone conversations or 200 colour television channels is to be installed by the Post Office between the research station at Martlesham Heath and Mendlesham in Suffolk. The Marconi Company Research Division is to develop and supply the prototype terminal and repeater equipment under a Post Office contract worth £180,000.

For the experimental link the frequency range will be divided into two main bands, 32 to 50GHz and 50 to 90GHz, although ultimately, this will be extended to 100GHz.

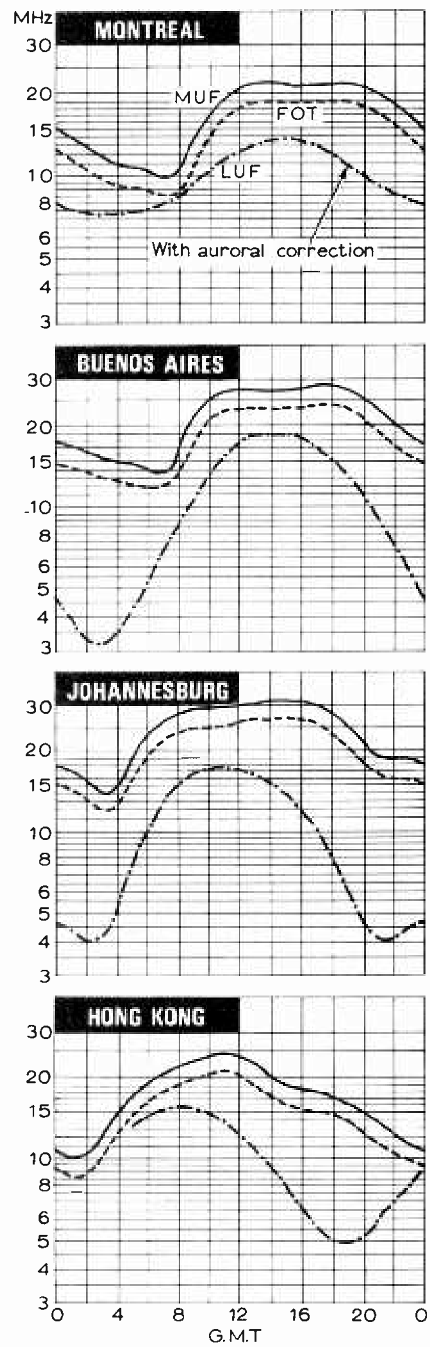
Correction

Please make the following amendments to Fig. 9 of the article "Electronic Morse Keyer" by C. I. B. Trusson and M. R. Gleason published last month. On the upper view of the circuit board the lead from the tune/operate switch should be connected to pin 3 of the MP104 (not pin 2) and the pin numbering of the MP102 should be altered to read 5 to 9 from top to bottom on the left-hand side and 4, 3, 2, 1, 10 on the right-hand side. On the underside view of the board the pin numbering of the MP102 should be amended as described and the position of the conductor breaks at the following points should be lowered by one position: 29/22, 29/28, 25/22, 26/33, 28/37, 27/37. Finally the break at 11/41 should be moved one place to the right.

H.F. Predictions—September

Seasonal change is evident as a slight increase in peak median standard usable frequencies (MUFs). This is sustained during daylight on all routes illustrated except Hong Kong which, due to the large time difference between control points, develops a continuously varying MUF.

MUFs shown are predicted medians of day-to-day values for the month. Distribution of daily values about the median varies with geographic location, season and time of day. Long-term observations have shown that the optimum traffic frequency (FOT)—i.e. MUF exceeded on 90% of days—is between 72% and 90% of MUF. A constant 85% is used for these charts, prepared by Cable & Wireless.



PUT

The programmable unijunction

by O. Greiter

The unijunction transistor and some of its applications have been described in articles in *Wireless World*, last month and the month before. To remind readers who have never met the unijunction we may note that it consists essentially of a silicon bar with an ohmic, non-rectifying, contact at each end and a tapping point near the middle. It is a resistor with a tap, except that the tap is a rectifying junction: one junction, hence the name. With a voltage applied to the bar it behaves just as a fixed-tap potentiometer with a diode in series with the tap. Unless the bias on the diode exceeds the sum of the diode starting voltage and the voltage at the tap, the diode is cut off. The tap is about two-thirds of the way along the bar, and this fraction is called the stand-off ratio. Problem: how accurately can you attach a tap to a sliver of silicon? The geometry actually used makes the problem even more difficult, so that the production line stand-off ratio has a wide tolerance and it costs extra money to get even a moderate tolerance.

When the diode is brought into conduction carriers are injected into the silicon bar, and as they are swept into the bottom part of the potentiometer they alter its resistance, drop the tap ratio, let more carriers in through the diode and so on. Rather like a one-man band, each part of the unijunction plays many parts.

The disadvantage of this device is that the tolerances must be dealt with in the external circuit. Oscillators and timers need to have a wide range of adjustment to take up the tolerance in the stand-off ratio, and this is particularly objectionable if it is required to choose components with the right temperature coefficients to compensate for the temperature changes in the unijunction itself.

The PUT

The programmable unijunction, so-called because the parameters are fixed by external elements and because it has three junctions, not one, gets rid of many of the disadvantages of the standard unijunction, even though it is just about the same price. It is officially described as a complementary thyristor, although by thyristor standards it is a very low level device indeed. We shall see that it is related to another, older, device.

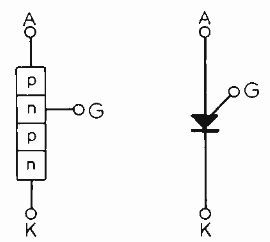


Fig. 1. The structure and symbol for a programmable unijunction transistor, showing anode, gate (anode gate) and cathode connections.

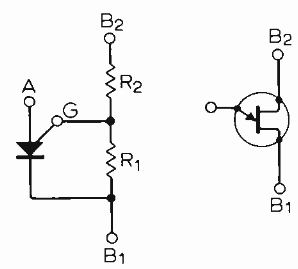


Fig. 2. Adding resistors R_1 and R_2 to the PUT gives the equivalent of the ordinary unijunction.

The basic structure of the PUT and the symbol used are shown in Fig. 1. In normal use the basic connection, the starting point, is to set up the simple circuit of Fig. 2. The gate is connected to the junction of R_1 and R_2, so that the upper junction is held at ηV_s, where $\eta = R_1/(R_1+R_2)$. This is just what we have in the unijunction, except that now R_1 and R_2 are ordinary resistors, which we can choose of any value, and any tolerance, we like. We can fix the stand-off ratio with as much precision as our money will buy, and certainly 1% will not cost very much.

So far we have added two resistors, and the cost of connecting them. However, in an ordinary unijunction circuit we put a resistor in the B2 lead, both to limit the current and, by choosing the right value for a particular specimen, to get the right temperature coefficient. We also need, in many circuits, a resistor in the B1 lead so that the current through the untriggered B2–B1 path will not flow through the thyristor gate, or transistor base, which is to be pulsed. In the quiescent state there is no current out at the cathode of the PUT and unless a resistor is needed for other reasons, to help with the voltage rating of the load device, for example, we need not use one. Thyristor trigger circuits thus come out, as we see in Fig. 3, with exactly the same number of components.

Performance analysis

The standard description of the behaviour of thyristors is based on the operation of cutting the device diagonally. This (shown in Fig. 4) allows the three junction device to be split into two transistors, forming an interconnected p-n-p, n-p-n, pair. The equations are

$$I_{c1} = h_{fe1}(I_{c2})+(h_{fe1}+1)I_{co1}$$
$$I_{c2} = h_{fe2}I_{c1}+(h_{fe2}+1)I_{co2}$$

The input current at the anode is

$$I_A = I_{c1}+I_{c2}$$

Solving these equations gives

$$I_A = \frac{(1+h_{fe1})(1+h_{fe2})(I_{co1}+I_{co2})}{(1-h_{fe1}.h_{fe2})}$$

At very low currents, h_{fe1} and h_{fe2} are very small and $I_A \simeq I_{co1}+I_{co2}$. Thus if the currents are small, the gain is low and thus the currents are small. As soon as we apply forward bias to either base we raise the value of h_{fe} for that transistor and in consequence let more current flow into the other; raising its h_{fe}, thus, and so on. When the

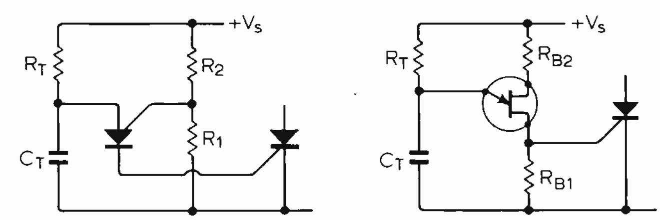


Fig. 3. In a practical scheme the number of components is the same with unijunction and PUT.

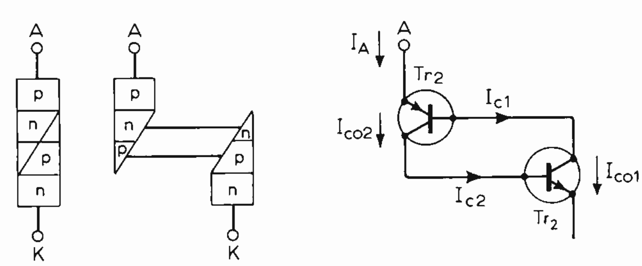


Fig. 4. The classic thyristor analysis.

product $h_{fe1}.h_{fe2}$ reaches unity, $I_A \approx \infty$.

A more elegant analysis can be carried out using the circuit of Fig. 5. If we use the current gain, emitter to collector, notation,

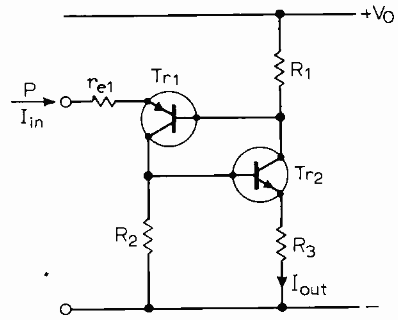


Fig. 5. Resistors added for more detailed analysis.

the overall current gain, $I_{out}/I_{in} = \alpha$, is given approximately by

$$\alpha = \frac{\alpha_1 R_2}{(1-\alpha_2)R_2 + R_3}$$

and the input impedance at P is

$$R_{in} = \kappa - R_1[\alpha\alpha_2 - (1-\alpha_1)].$$

Here κ is a positive term which contains device impedances. Very simply this boils down to a condition that the input resistance is negative if $R_1[\alpha\alpha_2 + \alpha_1 - 1] > r_{e1}$, the base-emitter input resistance of the p-n-p device.

The addition of these resistance elements, and the inclusion in a full analysis of the device resistances, leads to a situation in which the negative region of R_{in} has a reasonably constant value. The infinities do not appear. The overall characteristic becomes something like the one shown in

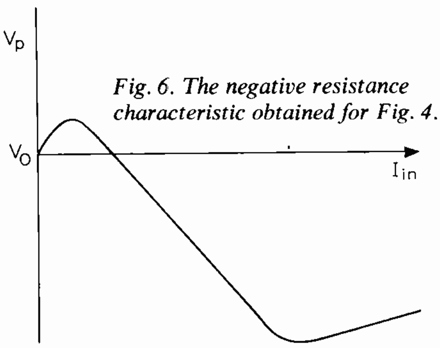


Fig. 6. The negative resistance characteristic obtained for Fig. 4.

Fig. 6. This, with its peak point and valley point, where both transistors are bottomed, is just the shape we have already seen for the unijunction. Some of us, indeed, have already seen this a good many years ago. Fig. 7 shows a common base amplifier, without the supply connections. The input impedance at the emitter is

$$R_{in} = r_e + (1-\alpha)R_B.$$

A negative resistance will be obtained if

$$(\alpha - 1)R_B > r_e$$

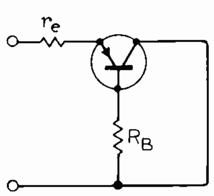


Fig. 7. In this circuit R_B includes the internal r_b.

or $\qquad \alpha > (r_e + R_B)/R_B$

The PUT as a point-contact transistor

Those readers who cut their teeth on junction transistors will regard this as a rather pointless exercise. Those who spent confused years with point-contact transistors will know that α can be greater than unity, and will remember that a stable amplifier needed a low impedance in the base line. The point contact transistor was, in general, a p-n-p-n device, but this was not an essential piece of knowledge. The essential detail was that you could measure α and it was greater than unity.

By accepting the fact that, current ratings apart, the PUT can be considered as a point contact transistor, we are free to look at various pulse circuits which we used twenty years ago. We can also get rather more "feel" to the conventional use of the PUT. The circuits of Fig. 2 can be rearranged in

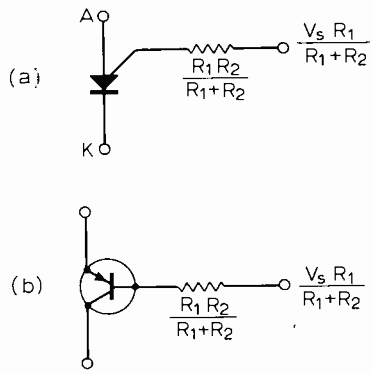


Fig. 8. The basic gate supply circuit, for PUT, and base bias for point-contact transistor.

the form of Fig. 8. We see that R_B is, in fact, the parallel combination, $R_1R_2/(R_1+R_2)$. The peak point, at which $R_{in} = 0$, will be when

$$\alpha - 1 = \frac{r_e}{R_B}$$

Memory, always a bad guide, suggests that α was not very dependent on current at low levels. The current dependent term is r_e. This would give us a peak point which was dependent in just the same way on R_B and we would expect to be able to choose the peak point current by choice of R_B. It turns out that this is indeed so, and the curves for a cheap and a more expensive

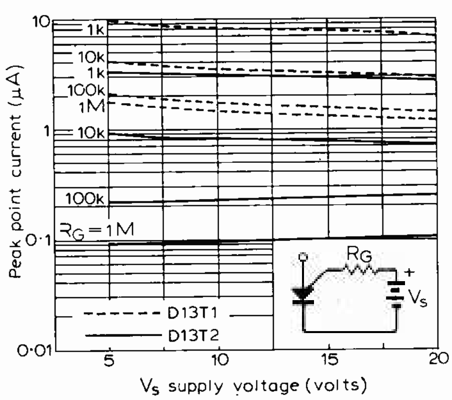


Fig. 9. From G.E. data. $I_p - V_S.R_G$ parameter.

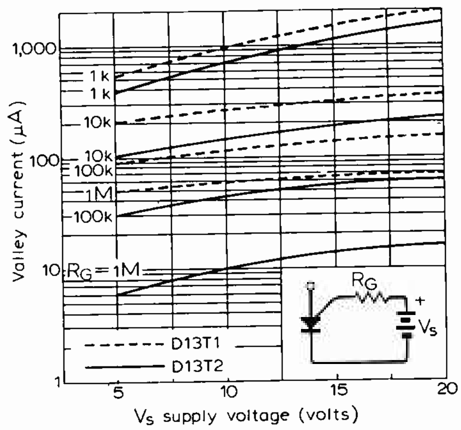


Fig. 10. From G.E. data. $I_v - V_S.R_G$ parameter.

type of PUT shown in Fig. 9 shows this quite clearly. For high values of $R_G = R_B$, the peak point current is very small.

It is not surprising that the valley point current is also very dependent on the gate supply impedance. The characteristics are shown in Fig. 10. Very roughly we see that $I_v = 100I_p$. It is a feature of the PUT, however, that these characteristics are determined by an external element. An approximate expression for the impedance looking in at the base is $Z_B = r_b + \left(\frac{1}{1-\alpha}\right)R_E$.

Forgetting the internal r_b, and looking for the turning point at which $R_B + Z_B = 0$

$$(\alpha - 1)R_B = R_E.$$

We can get results rather nicely if R_B is very large indeed, and since the peak point current is small the base current will be small. A non-linear resistance element will give just the right effect of high resistance round the peak point but, when the current increases, a low resistance to allow a high valley current to be set up. Two ways of

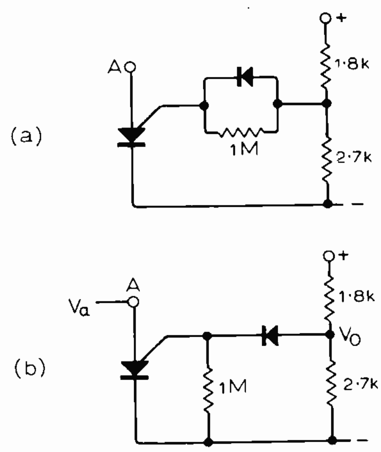


Fig. 11. Two ways of getting a low value of I_p combined with high I_v.

connecting this are shown in Fig. 11. The off-state gate current is very small, less than 0·1 μA, so that it is desirable to stabilize the peak point working condition by means of a 1 MΩ resistance. This is the form shown in Fig. 11(a). In Fig. 11(b) the gate supply impedance is not so high, as the arrangement leads to a small drain current down through the diode and 1 MΩ. The equivalent gate supply impedance is then between five hundred thousand and seven hundred thousand ohms. However, if the diode is reasonably similar to the upper p-n junction of the PUT and is operated at a current equal to the peak points current—say 1·5 A for the D13T1—the two temperature coefficients will stay in balance and the device will always trigger at $V_A = V_o$.

Both these circuits give a ratio I_v/I_p of above 500, compared with the ratio of about 100 if the diode is not used. It is probably desirable to look at the detail of this diode temperature compensation in terms of the point contact transistor. In Fig. 12 the transistor is shown with a leakage resistance R_t from base to collector. This represents the current path at the peak point. The circuit reduces then to the bridge form on the right, in which the "detector zero" is

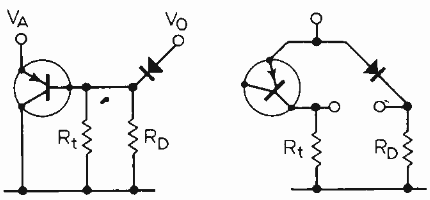


Fig. 12. Temperature compensation in terms of the point-contact transistor, and the bridge equivalent.

obviously maintained so long as the two diodes stay in step. One might expect that without going to an individually matched solution the improvement would be a factor of about 10. Just what the original temperature dependence would be is a matter of the form of the circuit. The data sheets give temperature curves, but they do not line up with the other characteristics. Nor do the characteristics shown in Figs. 9 and 10 line up with the tables. The data sheet information is added as an appendix.

Operational limits

Once we know what a device will do, in the sense of knowing that *nature* of its performance, we need to know the *quantity* of its performance, and the sort of circuit in which that performance can be exploited. The PUT is limited to 40 V working and to a steady state current of 150 mA—say 100 mA for a reasonable temperature range. It is imprudent to ask for pulses of more than 0·5 A. The maximum discharge, in an *RC* oscillator, with no extra resistance to limit the current, is 250 μJ. This means that if the circuit uses a 4 μF capacitor you should see that it triggers at below 10 V, so that $(\frac{1}{2})CV^2 = (\frac{1}{2}).4.10^{-6}.10^2 = 200\,\mu J$. In such a circuit, the supply resistance will probably need to be more than 50 kΩ and less than 10 MΩ with the cheap device. The time range thus extends up to about half a minute.

Relaxation oscillator

The most common circuit application is in the simple relaxation oscillator shown triggering a thyristor in Fig. 3. The capaci-

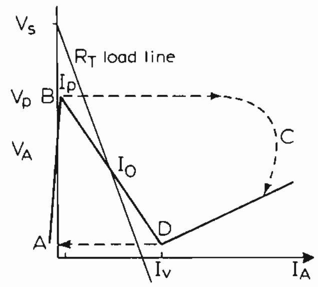


Fig. 13. Relaxation oscillator behaviour.

tance C_T is charged through R_T until the peak point is reached. The PUT is triggered, to discharge C_T into the cathode load. As this is much the same as the unijunction we can use Fig. 13 to remind us that the anode voltage moves up on the first stroke from zero towards B, along the path AB. At B the voltage can no longer be sustained across the PUT, and since the voltage across the capacitor will not change instantaneously the system switches to the only stable point, out beyond C. During the flight, current is flowing, and the voltage drops to give an intersection in the region C. The only current which can be supplied through R_T is much less than this, and the capacitor discharges, carrying the working point down to D. For the working point to approach I_o it would be necessary to follow the path DI_o. C_T is not capable of changing its voltage quickly upwards, because it can only charge through R_T. The working point jumps from D to A. There are some much more elegant ways of treating this circuit. We could use the negative resistance concept and look for the way the root dodges about between the left-hand side of the plane and the right-hand side. No doubt some of the brighter contributors to *Wireless World* will explain all this to us one day. The simple approach tells us how it behaves, if it works: the theory tells us if it works.

Notice that the R_T load line must cut the device characteristic once only, in the negative impedance region. This means, very roughly

$$\frac{\eta V_s}{I_p} > R_T > \frac{V_s}{I_v}$$

To get a particular frequency we choose R_T within this range, and then put

$$C_T = \frac{1}{fR_T \log\left(\frac{1}{1-\eta}\right)}$$

$$= \frac{1}{fR_T \log(1 + R_1/R_2)}$$

If we make $R_1/R_2 = 1{\cdot}7$ this reduces, since $\log_e(e) = 1$, to a simple equation: $fC_TR_T = 1$.

A disadvantage of the PUT is that in general the value of R_T needed for operation is a good deal higher than the value used in the corresponding unijunction circuit. Fortunately the fact that it is operating by transistor action rather than by resistance modulation means that it comes on quick and it comes on hard. The saturation voltage at 50 mA is 1·5 V, giving a saturation resistance of 30 Ω maximum, and they say that this drops to 3 Ω at higher currents. The switching time is fast, 60–80 ns. A quick calculation shows that if we take 100 mA from a 1000 pF capacitor it will discharge at the rate of 7 V in 70 ns. Thus we shall not swing down to an unreasonable trajectory until we go below the 0·001 μF value of C_T. Rather roughly, we lose half our pulse. With the ordinary unijunction the limit is in the region of three times this value.

Other PUT applications

There is a useful G.E. application report (90·70:11/67). Some of the following circuits are taken from this report, which also contains some thyristor trigger circuits. The first is a particular relaxation oscillator, which operates at only 3 V. The circuit is shown in Fig. 14. In this low voltage circuit

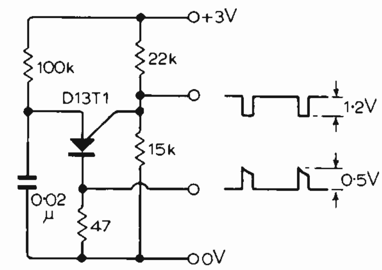


Fig. 14. 1 kHz oscillator for 2 V working.

there is the problem that the voltage needed to get the first junction into conduction, the emitter-base voltage of the equivalent point contact transistor, is an important fraction of 3 V. The anode gate is held at about 1·2 V, so that the anode must be brought up to about 1·7–2·0 V for triggering. If the more conventional stand off ratio of around $\frac{2}{3}$ had been used, the capacitance would have had to charge to above 2·5 V on its exponential run up to 3 V. This would mean a very flat approach to triggering and a great deal of uncertainty about the frequency, stroke by stroke. This circuit also gives very low drain on the battery. The total current is less than 100 μA. A conventional low voltage unijunction would demand about 1 mA.

Circuit action in which advantage is taken of the low operating current levels is exemplified by the circuit of Fig. 15. The

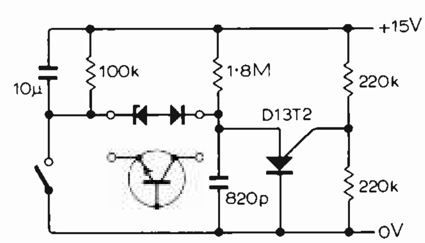


Fig. 15. Tone-pulse oscillator.

zener diode—diode circuit can be replaced by the emitter-collector path of a small transistor. As it stands, the 100 μA available at the anode through the 100 kΩ resistance will hold the PUT, a low-current type, beyond the valley point. If the switch is closed for a short time the 10 μF capacitor is charged, biasing the diode off and leaving only the 1·8 MΩ resistor to supply the anode. This is an oscillatory situation, and the circuit runs at about 1000 Hz until the capacitor has discharged through the 100 kΩ to the point where the zener diode will conduct. With the values shown the tone should be on for about 1 second.

Timers. These are basically oscillators of which the frequency is so low that all the attention is concentrated on a single cycle. The important advantage of the PUT is that because the stand-off ratio is determined by two external resistances almost all the parameters of the timer can be selected in advance. The circuit shown in Fig. 16 uses

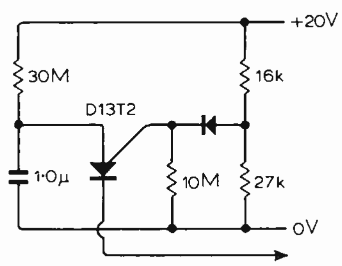


Fig. 16. 30-second timer using expensive PUT.

1% tolerance components. The stand-off ratio is 0·61, or $R_1/R_2 = 1{\cdot}69$, which agrees with our earlier finding that this is the value to make $fC_TR_T = 1$, or

$$\tau = C_TR_T$$

The effective value of R_G is not too certain, but the circuit does not make sense if it is above 1 MΩ: it is the resistance of the diode when passing about 1·5 μA. The peak-point current will be in the region of 0·1 μA. This can be supplied by the 30 MΩ resistor. It would appear that trouble might occur at very low temperatures. At -25°C the peak point current is 1 μA, and there is only 0·3 μA available. The circuit could stick just below peak point.

To operate a circuit of this kind at low temperatures, or to use the cheaper PUT, the technique is to pulse the anode gate. This technique was described in connection with the conventional unijunction. The

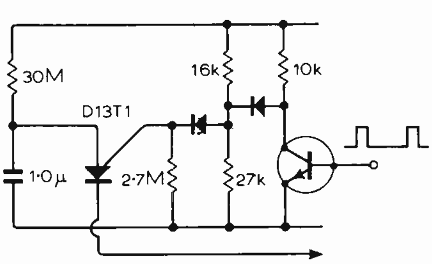


Fig. 17. Modified circuit of 30-sec timer to use cheaper PUT.

modified circuit is shown as Fig. 17. The application report quotes a 10 μs pulse at 1000 Hz, but as the circuit is a 30-second timer in the 1% class it would be reasonable to settle for a much slower sampling rate.

Time or frequency trimming is claimed to be much simpler with the PUT. The comparison made is with trimming provided

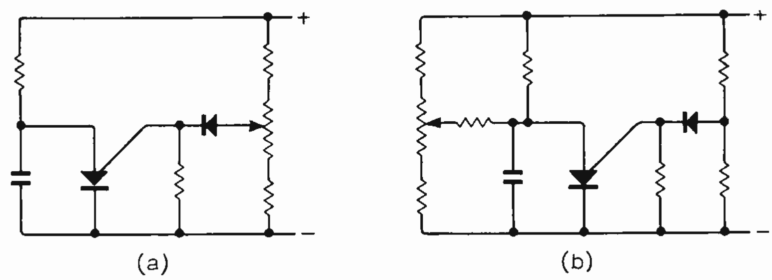


Fig. 18. Two alternative ways of frequency trimming.

by means of a variable resistance in R_T itself. It is, however, possible to trim the frequency of the unijunction and PUT circuits in the timing path in a way which still leaves a lot of flexibility in the choice of values. The two alternative ways of adjusting the frequency over a small range are shown in Fig. 18.

A number of other applications are to be found in the GE application report, which is essentially thyristor-minded in the view of this writer. The circuits for ring counters and thyristor triggering are interesting and useful, but there are some speculations to which the remaining space in this article may well be devoted.

We have seen that the PUT is very much like a point contact transistor. One difference, of course, is that it is a silicon device, whereas all, or at any rate almost all, the point contact transistors were germanium devices. The other feature is that junction technology has given us very low saturation resistance, thus widening the available working band inside the supply voltage. For the D13T1 we find that at 10 V V_s we have a valley current of 1 mA with an R_G = 1 kΩ. This valley current gives an average value of

$$R_{in} = 10\ \text{k}\Omega$$

so that, roughly

$$(1-\alpha) = R_{in}/R_B = R_{in}/R_G = 10$$

Thus $\alpha = 11$

At $R_G = R_B = 10$ kΩ

$$R_{in,e} = -10/0{\cdot}25 = -40{,}000$$

Thus $\alpha = 5$.

It is apparent that the equivalent point contact transistor is a rather non-linear device of rather uncertain characteristics. Even so, there is no reason why we should not attempt to use the PUT in some of the circuits which were evolved in the days of the point contact transistor. Fig. 19 shows a greatly simplified equivalent common-base amplifier. The input impedance of this arrangement has already been given, on the assumption that R_c is small. The input impedance of the common emitter circuit

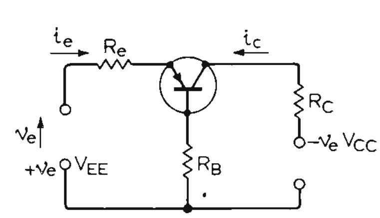


Fig. 19. Common-base point-contact circuit.

is, in simplified form

$$R_{in,b} = R_B + \left(\frac{1}{1-\alpha}\right)R_c.$$

This gives a negative resistance for the same condition as for the common base circuit. The difference is that with the common base circuit the impedance at the emitter line input is open-circuit stable, while with the common emitter circuit the impedance at the base input is short-circuit stable. We have already seen how, by putting a device which is a short-circuit at very high frequencies across the emitter input we achieve instability which drives the system from one limit condition to the other. This is the capacitance or controlled relaxation oscillator. The same effect can be produced by the use of inductance in the base circuit. A relaxation oscillator can be constructed using the arrangement of Fig. 20. The setting-up condition is, as in the

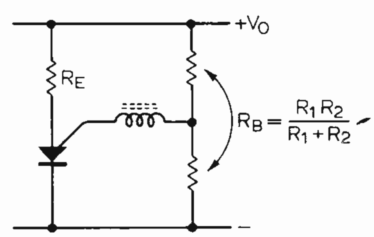


Fig. 20. Inductance controlled relaxation oscillator.

RC circuit, the choice of R_E to give a static current which is between the peak point and the valley point. If the effective value of R_B is 10 kΩ and, to make life simple we take $R_1 = R_2 = 20$ kΩ, and $V_o = 20$ V, the valley current will be rather over 200 μA. A good mid-point will be at $I_o = 100$ μA, giving $R_E = 100$ kΩ. At very high frequencies the inductance will hold the base current constant, and if instantaneously this base current is insufficient to allow 100 μA to flow the circuit will progressively cut itself off. Base current will disappear, because of the movement of the emitter. There is a *voltage* spike at the base, just as in the *RC* circuit there is a current spike when the circuit triggers. The whole action can be described on a negative resistance S-curve. It is not profitable to go into detail, because the actual device has been designed to handle high peak currents, but not to handle high peak voltages.

More interesting, from a practical point of view, are the circuits shown in Fig. 21. These two circuits provide the required high or low impedance required for oscillation at a single finite frequency. These oscillators

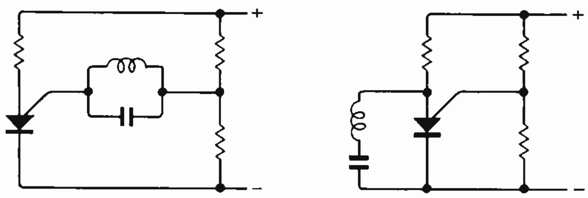


Fig. 21. Negative resistance LC oscillators.

are simple but are not likely to be very high class circuits, relying as they must on a very non-linear negative resistance to maintain oscillation, and a sharp limiting action to determine the amplitude. If we continue on the lines suggested by point contact transistor experience we arrive at a circuit of the type shown in Fig. 22, in which a low

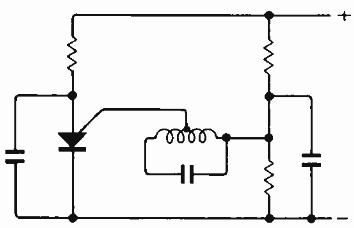


Fig. 22. High stability should be obtained by decoupling and a tapped tank circuit.

inductance, high-Q, circuit is used together with a good deal of decoupling. It would seem to be reasonable to add a thermistor in parallel with the LC circuit to set the amplitude of oscillation so that it remains inside the cut-off and saturation regions of the transistor.

One area in which the point contact transistor is missed is in the design of simple pulse regenerators. It is thus interesting to compare two circuits, one for a point contact device and one for a PUT (Fig. 23). The PUT circuit produces the output pulse when the transistor is switched off. If we consider the essential circuit as a monostable we see that the only real difference is in the choice of trigger points.

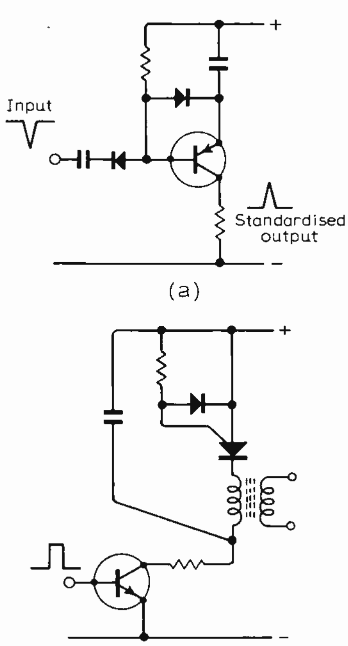


Fig. 23. Two ways of producing standard sharp pulses.

The purpose of this article, and of the two of the conventional unijunction which preceded it, has been to draw the attention of readers who have not met these devices to the wide range of circuits in which they can be applied. The PUT, in particular, would seem to be of much wider application as a transistor than is implicit in its official designation of a general purpose low power thyristor. There is enough circuit information on point contact circuits which may be applicable to keep a lot of people happy for a long time.

Conferences and Exhibitions

Further details are obtainable from the addresses in parentheses

LONDON
Sept. 7-9 — Imperial College
Computational Physics
(I.P.P.S., 47 Belgrave Square, London S.W.1)
Sept. 7-11 — Grosvenor House
International Broadcasting Convention
(I.B.C., c/o I.E.E., Savoy Place, London WC2R OBL)
Sept. 15-18 — Savoy Place
Gas Discharges
(I.E.E., Savoy Place, London WC2R OBL)
Sept. 15-18 — Olympia
Bio-medical Engineering Exhibition
(U.T.P. Exhibitions Ltd, 36-37 Furnival Street, London E.C.4)
Sept. 22 — William Beveridge Hall, University of London
Computer Output on Microfilm
(Microfilm Association of Great Britain, 109 Kingsway, London WC2B 6PU)
Sept. 28-Oct. 1 — Savoy Place
Centralised Control Systems
(I.E.E., Savoy Place, London WC2R OBL)

BIRMINGHAM
Sept. 18-20 — The University
The Nuffield Advanced Physics Course
(I.P.P.S., 47 Belgrave Square, London S.W.1)

BRIGHTON
Sept. 15-17 — Hotel Metropole
Power Sources Symposium
(International Power Sources Symposium Committee, P.O. Box 136, 26 Wellesley Road, Croydon, CR9 2EG, Surrey)

DURHAM
Sept. 2-9
British Association Meeting
(British Association for the Advancement of Science, 3 Sanctuary Bldgs, 20 Gt. Smith Street, London S.W.1)

EXETER
Sept. 15-18 — The University
Solid State Devices
(I.P.P.S., 47 Belgrave Square, London S.W.1)

HARROGATE
Sept. 18-20 — Cairn Hall
Audio 70
(Exhibition & Conference Services Ltd, Victoria House, Claremont Avenue, Harrogate)

MANCHESTER
Sept. 8-11 — Hotel Piccadilly
Electronic Instruments Exhibition
(Industrial Exhibitions, 9 Argyll Street, London W1V 2HA)

OXFORD
Sept. 14-16 — The University
Photo-electron Spectroscopy
(I.P.P.S., 47 Belgrave Square, London S.W.1)
Sept. 28-Oct. 1 — New College
Quality Assurance in Action
(Inst. of Engineering Inspection, 616 Grand Bldgs, Trafalgar Square, London W.C.2)

READING
Sept. 7-9 — The University
Ion Implantation
(I.P.P.S., 47 Belgrave Square, London S.W.1)

SHEFFIELD
Sept. 22-24 — The University
Microwave and Laser Instrumentation
(R. A. Ganderton, *Design Electronics*, Dorset House, Stamford St., London S.E.1)

SOUTHAMPTON
Sept. 7-10 — The University
Measurement Conference
(British Society for Strain Measurement, 281 Heaton Road, Newcastle upon Tyne NE6 5QB)
Sept. 22-24 — Skyway Hotel
Electronic Instruments Exhibition
(Industrial Exhibitions, 9 Argyll Street, London W1V 2HA)

SWANSEA
Sept. 21-24 — University College
Electronic Engineering in Ocean Technology
(I.E.R.E., 8-9 Bedford Square, London W.C.1)

TEDDINGTON
Sept. 2-4 — National Physical Laboratory
Man-Computer Interaction Conference
(I.E.E. Savoy Place, London WC2R OBL)

WARWICK
Sept. 21-23 — The University
Temperature Measurement
(I.P.P.S., 47 Belgrave Square, London S.W.1)

OVERSEAS
Aug. 31-Sept. 4 — Munich
Hybrid Computation
(Prof. Dr. J. Heinhold, Kongressburo AICA-IFIP 1970, Institut fur Angewandte Mathematik, Arcisstr. 21, D-8 Munchen 2)
Sept. 8-12 — Budapest
Magnetic Recording Conference
(Hungarian Optical, Acoustical and Film-technical Society, Keszült: 200-pld-ban 70/2808-MTESX HNY.Bp.)
Sept. 18-27 — Berlin
German Industries Exhibition
(Berliner Ausstellungen, 1 Berlin 19, Charlottenburg, Messedamm 22)
Sept. 21-24 — Panama City
Engineering in the Ocean Environment
(I.E.E.E., 345 East 47th Street, New York, N.Y. 10017)
Sept. 23 & 24 — New York
Electron Device Techniques
(I.E.E.E., 345 East 47th Street, New York, N.Y. 10017)

Vehicle Location Systems

Electronic techniques to improve road traffic control

by R. A. Tyler, M.I.E.R.E.*

It has been standard practice in the past to apply static means of control to traffic movements, i.e. road signs, markings in the road, fixed 'one-way' systems, etc, but as traffic flow is a dynamic function, adequate control has demanded dynamic methods, such as traffic lights and policemen on point duty. By applying more modern control techniques we should be able to progress a particular vehicle through a traffic system from starting point to destination in the most economical way. Although cost effectiveness at present precludes the use of such techniques for the benefit of the ordinary motorist, the advantages afforded a vehicle fleet operator in being able to control the dispositions of his vehicles in a given area make such a system attractive.

A promising scheme is to provide a supervisory control facility which, by continuous situation reporting, enables route plans to be revised and corrective measures to be reported to the vehicles involved (Fig. 1). Radio links which have voice, data, or voice and data channels can carry the necessary information between vehicle and base, and control room devices exist for displaying this information. However, in order to provide an overall control system there remains the need for a method of determining the position of the vehicle to a sufficiently high degree of accuracy within the traffic system.

The ease with which a vehicle position may be determined depends largely on the vehicle operating mode and the degree of co-operation which may be expected from it. Land transportation vehicles may operate in any of the following three modes:

(a) Free range—where vehicles are allowed to move about without restriction over a specified two-dimensional space, as when operating in a desert environment.

(b) Road restricted—in which vehicles, usually wheeled, are free to move anywhere within the confines of a predetermined road network. This category includes private cars, police and emergency vehicles, etc.

(c) Route plying—vehicles restricted to particular routes within a specified road or rail network. All rail vehicles are included in this group. Although fundamentally of

* The Marconi Company Ltd.

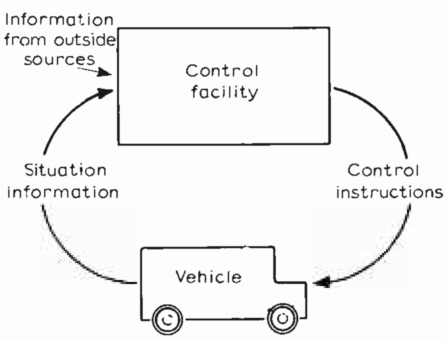


Fig. 1. Vehicle control loop.

type (b) buses may be considered as a sub-group in that they are usually constrained to the route-plying mode.

Clearly, the greater the number of constraints on a vehicle the easier it is to track. A mode (c) vehicle may be quite satisfactorily tracked along the route simply by a knowledge of start time and distance travelled.

Taking full advantage of up-to-date technology it is possible to rule out the use of radio telephone links, which demand a high degree of crew participation, and confine discussion to consideration of location systems which can automatically provide a fixed base station with regularly updated vehicle position information.

Location methods fall into two main categories. The first requires means of discovering the whereabouts of the vehicles directly from the base station. The second needs each vehicle to be provided with equipment to determine its own position and for passing this information to the base station. By using radio, both methods become practicable.

Location methods, category 1

Radio location systems falling into this category include those employing direction finding, navigation and radar. Direction finding does not easily lend itself to automatic detection of large-vehicle fleets, and ground clutter problems present considerable difficulties when using conventional radar systems.

The most promising approach would appear to come from the use of navigational aid methods which apply triangulation techniques to a signal radiated from the vehicle. Location accuracies of better than half-a-mile within an area of open terrain of 20-30 miles radius have been achieved using phase comparison, and there is no reason to suppose that this represents the limits of either range or accuracy. A major difficulty encountered with all methods where location information is to be extracted by reference to the physical nature of the radiated signal is one of multi-path reflection.

Although this difficulty is considerably reduced over areas of open countryside, it remains perhaps the most serious problem in urban environments. Early expectations of using first returned pulse (FRP) delay measurements have been shown experimentally to be disappointing. In this method the vehicle carries a pulsed transmitter which is triggered from the master base station. The difference in delay of the reply pulse from the vehicle arriving at two separated slave receiving stations is related to the receiving station base line, and the vehicle position may then be fixed using standard hyperbolic navigation techniques.

When multi-path propagation conditions exist the vehicle reply appears as a string of pulses at the receiving stations and it is difficult to determine the vehicle position (Fig. 2). In FRP systems, delay comparison is made only on the first pulse to return in each case as this must have arrived over the direct path and later pulses are neglected. Provided that the first pulse to return can be positively identified, this somewhat complex method can be used quite effectively. But in heavily built-up areas the welter of closely spaced reflected signals makes definition of the first returned pulse difficult and the accuracy of such a system thereby suffers. Also the FRP system is in considerable difficulty where more than one vehicle is involved unless selective calling is used. However, this principle may be applied to vehicles operating in any of the three modes referred to previously.

Operation in urban environments, where road junctions frequently occur at less than one hundred-yard intervals, imposes stringent accuracy requirements on any method of vehicle location. Accuracies of one hundred feet or less are simply not feasible in such an environment when using any of the methods in this category. Thus, we have to look elsewhere for a workable solution to the location problem, and from what follows it will be seen that methods falling into the second category show more promise.

Location methods, category 2

This method is basically the same as the simple one referred to in Category 1, but the vehicle is expected to know where it is at all times. It is, in effect, questioned on its position whenever the base station requires to update its record of vehicle dispositions.

The problem therefore may be split into two requirements, (a) some form of location equipment within the vehicle which allows it to determine its own position, and (b) a means of transferring this positional information back to base on demand.

There are a number of ways of achieving the requirement outlined in (a), two of which will be dealt with in greater detail later. But in order to use this method requirement (b) demands some form of communication channel between vehicle and base and some important experiments in the development of one such link will now be described.

Data link 1

In the planning stage a number of constraints were placed on the information channel such that the following initial specification was considered desirable.

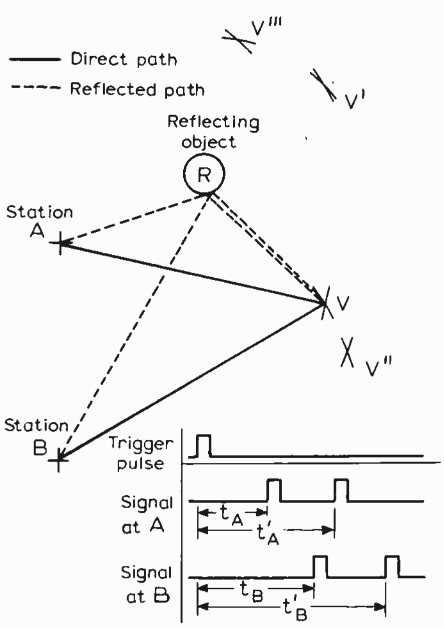


Fig. 2. Anomalies which can occur in a pulsed system.
t_A represents path A-V taken with t_B giving vehicle position V
t'_A represents path A-R-V taken with t'_B giving vehicle position V^I
t_B represents path B-V taken with t'_4 giving vehicle position V^{II}
t'_B represents path B-R-V taken with t_A giving vehicle position V^{III}
By using the first returned pulse method only delays t_A and t_B would be used to fix the position of the vehicle at V.

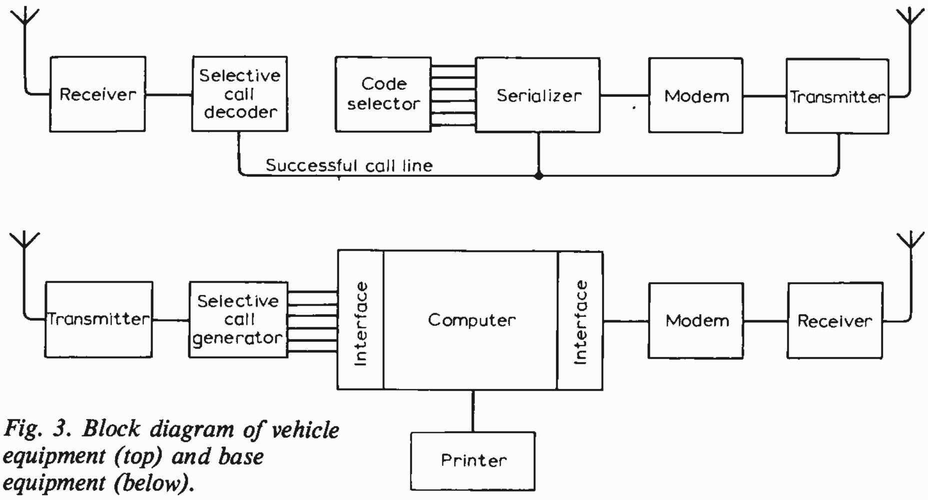


Fig. 3. Block diagram of vehicle equipment (top) and base equipment (below).

(1) The radio part of the link should consist of commercially available two-way v.h.f. equipment, so the information must be fitted into a baseband a few kHz wide.
(2) Selective calling of vehicles was essential for the interrogation methods used. As the overall system was to be automatic, a digital computer would be used to provide correctly timed interrogation signals and to interpret returned information.
(3) As returning information was to be fed into the on-line computer, it was desirable to make the return path a digital link.

It was discovered in the early stages that very little was known about the behaviour of v.h.f. data links applied to vehicles moving in and around an urban environment. Therefore it became imperative to obtain statistical results using practical equipment, and a series of experiments were set up using the computer to gather and sort the necessary data.

Early experiments were conducted using audio-tone selective calling on the base-to-vehicle link. When a particular address was recognized the vehicle would reply to base with a 40-baud message. Upon reception at the base station, this message would be de-serialized in the computer and the contents compared with a replica of the original message. If incorrect, details of this comparison would then be printed out and at the end of each run statistical data would be computed and printed (Fig. 3).

The experiments were organized in the following manner. Two complete sets of vehicle equipment were made available, each set comprising a two-way selective call f.m. radio and a specially developed digital generator which provided, on command, a pre-set message 16 bits in length. Each vehicle had a particular call sequence allocated to it, which the computer could select on a rotational basis, together with four sequences representing imaginary vehicles.

In the event of receiving no reply from a call to a genuine vehicle, the computer would break the rotational calling routine and recall the offending vehicle. If again no reply was received, a third call would be made. If still no reply was forthcoming the legend NO REPLY would be printed out and the routine calling sequence resumed. Further, if after an interval of five minutes no reply at all had been received from a particular vehicle, it was assumed that a major failure had taken place and a visual indication would be given at base (Fig. 4).

After a great number of runs had been made with the vehicles operating over diverse routes covering both urban areas and open countryside, the considerable amount of data acquired was analysed and compared with similar results obtained using the radio equipment in voice communication surveys over the same routes.

The field pattern associated with v.h.f. radio installed in a vehicle operating within an urban environment is extremely complex and difficult to analyse. It can be

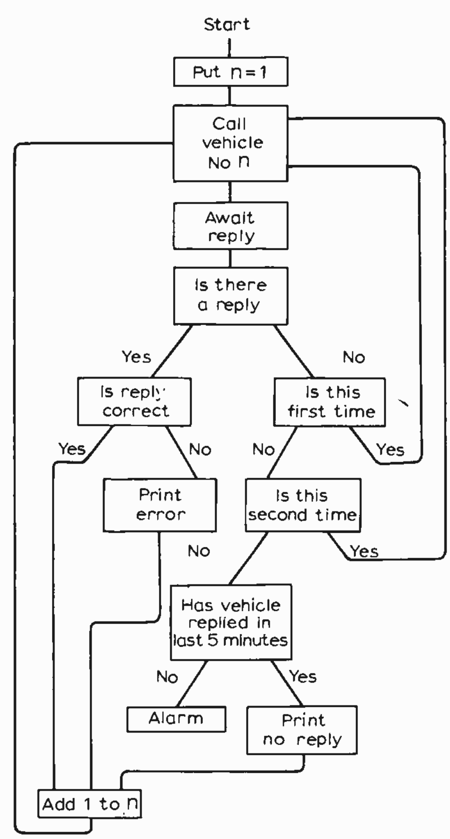


Fig. 4. Flow diagram for data link experiment.

visualized as a fixed pattern due to multiple reflections from stationary objects, buildings, street furniture, other vehicles, etc. On this is superimposed a continuously varying micro-structure due to traffic moving in the vicinity of the vehicle aerial. The situation is further complicated by the fact that the aerial is being caused to move in a somewhat erratic manner, due to traffic conditions, through this complex and changing pattern. The net result is that between vehicle and base there is an indeterminate transmission path, the characteristics of which change drastically from moment to moment. It has been shown that the field strength at the base receiver can vary as much as 20-30dB at a rate dependent upon the speed of the vehicle and further modified by the relative presence and speed of surrounding traffic.

The transmission of voice over such a path generally causes little concern as the intelligibility of speech is not greatly affected by such rapid fading. But the very nature of digital data usually demands a near perfect transmission path. Additionally, the signal may fade during the 0.4s time slot used for transmitting the 16-bit message, with consequent corruption of data. However, in spite of this gloomy picture, the experiments showed that over a known good speech path in an urban area, data error rates of only a few per cent were likely. But even with error rates of this magnitude, data, to be usable, must be subjected to some form of error detection and subsequent correction.

Obtaining navigational information

Having established provisional parameters of a mobile data link (b) the question remained, how could the necessary information be obtained and passed successfully over the link.

To solve the problem the vehicle requires some means which will enable it to determine its own position as outlined in (a). This is similar to the conditions imposed in conventional navigation and it would not be unreasonable to expect a solution to be obtained using the same approach. Indeed this approach is possible within the restraints of road traffic movements.

The classical method of navigation between two places requires several points, or 'fixes', to be taken along the route, where an accurate knowledge of actual position may be obtained. The number of fixes required depends on the accuracies of the results, distance between start point and destination, etc. Location between one fix and the next is determined by a procedure known as 'dead reckoning', which is based on a knowledge of the direction and distance travelled from the last fix; consequently provision must be made for measuring this direction and the distance travelled.

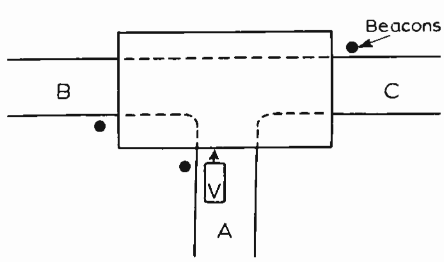


Fig. 5. T-junction as a three-port system.

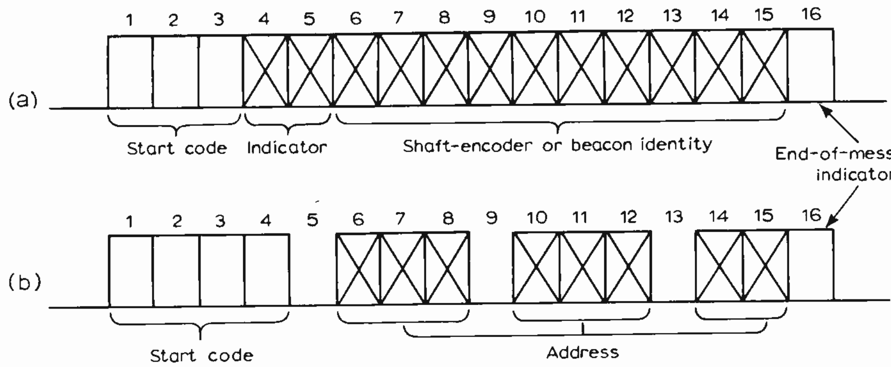


Fig. 6. (a) Vehicle to base—location message, (b) base to vehicle—vehicle address message.

It is of course possible to navigate entirely by dead reckoning. To be able to do so over large distances requires extremely accurate vector measurements with such expensive items as inertial platforms and on-board computers. However, the land vehicle operating in modes (b) or (c) is confined to a specific network of routes over which it is constrained to travel, and location systems can make use of these restraints.

By way of illustrating how such restrictions may be used, consider a *T* road junction as a three port sub-system (Fig. 5). A vehicle *V* entering at port *A* must emerge at Port *B* or *C*, or reverse out of port *A*, otherwise it must have stopped within the confines of the junction. In this example three detectors would be needed to cover the conditions of vehicle movement in areas between junctions.

Returning to the conventional method of dead reckoning with fixes, it is easily seen that such an approach can be made in a simpler manner in the road transport situation. All that is required to provide a system of vehicle location (i) to arrange within the vehicle a means of measuring distance travelled, (ii) to establish along each route a number of fixed points which, as each one is passed, may be identified by apparatus installed in the vehicle, and (iii) to determine whether or not the v.h.f. data link is adequate to pass this information to the central base station. It is, of course, assumed that the base station has data handling equipment available. This should be able to extract the necessary information from the link and display it in a form suitable for human operators to comprehend, thereby enabling control strategies to be applied.

Wheel revolution counting

In a wheeled land vehicle a simple, if somewhat crude, method of measuring distance travelled is to count the number of revolutions made by the wheels while traversing a given route. Because this can be easily arranged in a modern motor car, experiments were planned in order to gain experience of this form of measurement and to determine just how seriously data link 1 could be used in a practical system. To this end a suitably geared mechanical pick-off, operating a 10-bit digital shaft-encoder, was fitted into the speedometer drive cable of one of the vehicles used in the previous experiments.

The equipment was so arranged that a call from the base station would initiate a reply from the vehicle containing an up-to-date indication of shaft-encoder position. Tone selective calling was used to address the vehicle but in this case each of the 16 bits of the reply message was given a different significance. The first three became a fixed start code, the next two were used as an indicator, followed by 10 bits of Gray code from the shaft-encoder, and the final bit was used as an end of message indicator (Fig. 6a).

Two routes were selected, both of which started and ended at the base station site. Route *A* was about $3\frac{1}{4}$ miles long mostly through lightly built-up areas. The longer route *B*, about $8\frac{1}{2}$ miles, included a 4-mile figure-of-eight loop in a heavily built-up town centre, situated about 3 miles from the base transmitter. A number of marker points, spaced one quarter to half a mile apart were chosen on each route and the vehicle crew was asked to throw a switch when passing each marker. The operation of the switch caused the two indicator bits in the message to change, signifying to the computer that such a point had been reached.

Several calibration runs were made and the information was stored in the computer. Operational runs were then performed, the results of which were printed out in the following form. A first column identified the marker points on the route. The second and third columns showed the times at which the vehicle reached the positions indicated in the first column. These were determined both by comparing the shaft-encoder information with that obtained during the calibration run, and by a change in indicator bit condition. The fourth column indicated distance error and a fifth column provided some indication of average vehicle speed between markers. Statistical data on radio-link conditions during the run were collated at the end of each print-out.

Results showed that under good radio conditions adequate accuracies could be achieved even with a simple set-up, position indications within 50ft being frequently obtained, to some extent vindicating the wheel count method Tracking became difficult, if not impossible, with the simple system under bad radio conditions, i.e, heavy interference, shadowing, etc, indicating that, as had been expected, some form of error detection would be essential under such conditions.

As the system contained a computer it seemed natural to attempt a software solution to this problem. A vehicle tracking programme was written which enabled a number of credibility checks on the incoming data to be made and only those able to pass this screen were considered reliable. Tests with this programme gave very good tracking, even under quite heavy interference conditions, and it was felt that the feasibility of such a system for tracking route-plying vehicles had been demonstrated.

The next major step in this particular line of development was to improve the method of presenting results to a controller. It was therefore arranged to present the vehicle situation as a visual picture using a graphical display unit. It was logical to make use of this approach, as a computer was included in the system and flexibility of display afforded by this means made the case overwhelming. The opportuntity to be able to switch from a map of the area showing the immediate disposition of all his vehicles (Fig. 7) to another presentation showing the up-to-the-minute operational situation on any particular route of his choosing gives the vehicle fleet operator a unique degree of operational control at the press of a button.

Clearly, the information that might flow over a vehicle-to-base digital link need not be restricted to positional data only. In fact it was arranged on certain occasions to make a change in message indicator bits to represent a 'driver alarm' signal. By this means the vehicle driver could alert a controller simply by throwing a switch in the vehicle, which caused a flashing alarm indication to appear on the controller's display screen. In a practical system a great variety of data could be sent back to base over the link. For instance, bus operation controllers might wish to know the passenger loading at certain times, and passengers could be counted automatically when boarding or alighting at bus stages, the information being relayed back to base.

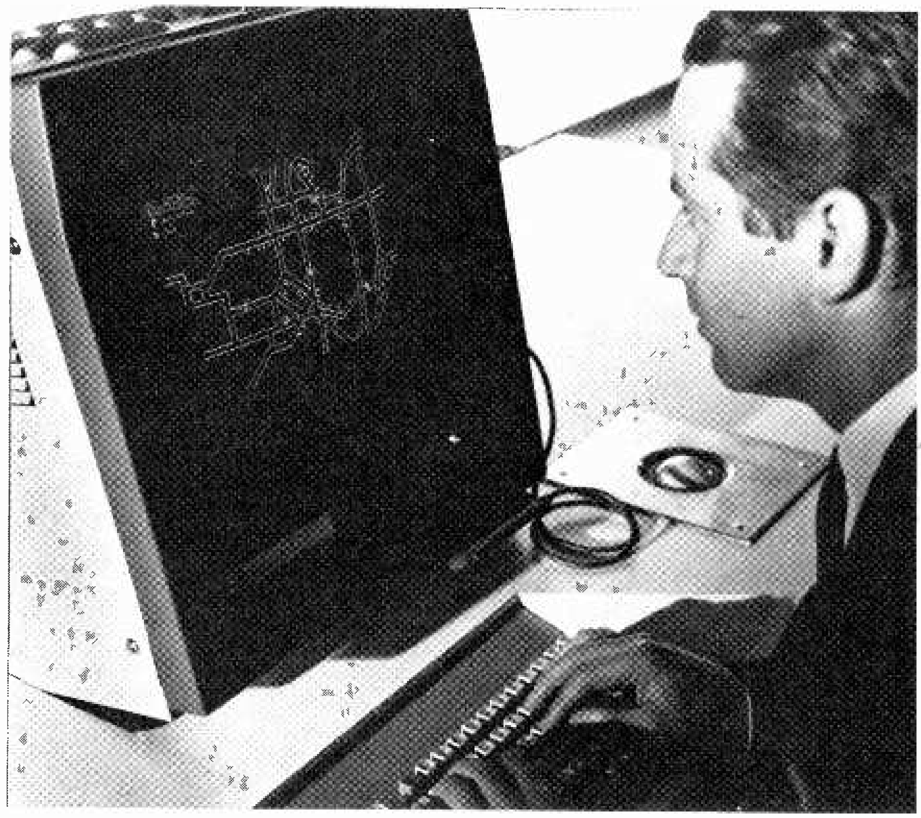


Fig. 7. Map display of the area showing the position of all vehicles.

Data link II

The work so far had been done with only one, or sometimes two, vehicles. Because of the capital outlay required for such a system, it would only become really economic when applied to fleets containing probably a few hundred vehicles. This being so, some attempt had to be made to speed-up the calling rate of data link I which had been held to about one per second, for three reasons. These were, first, the use of tone-selective calling required 500ms to achieve a successful interrogation, second, upon being called some time was needed for the vehicle transmitter to reach full power, and third, even a 16-bit message took 400ms to send at the 40-baud rate. So a faster link—data link II—was considered.

The philosophy behind any system which might be produced was that of providing add-on digital equipment to any good commercially available mobile v.h.f. radio. In consequence it was felt that only the first or third items above could be tackled at this point in time. Indeed, by increasing the signalling rate to 1,000 bauds and providing a similar outgoing digital link for pulse code addressing, the interrogation and reply time would be substantially reduced. The next step therefore was to determine the feasibility of signalling over the link at this higher rate.

It was felt that the emphasis on the two main types of error would change when using 1,000-baud rate, as opposed to 40-baud signalling, due to the nature of the radio signal. The time required to transmit a complete message at the faster rate was less than that for just one bit at the slower, and this would manifest itself in a greater number of 'no replies' than 'corrupt messages'. Experiment largely confirmed this expectation, although there was a marginal increase in overall errors. However, it was established that there was no great penalty to be paid for using the increased data rate and work could now go ahead on providing a similar digital link from base to vehicle.

This entailed major changes in the hardware used so far at both ends of the link. For instance, the serializer in the vehicle had to be modified in order to de-serialize the incoming message, and similar arrangements were required at the base end. Further, results had shown the need for some form of error detection for the outgoing link and this would now need to be accomplished by hardware methods in the vehicle. Some form of protection could be achieved by building a fair degree of redundancy into the interrogation message sent to the vehicle. For convenience, a 16-bit message was used but only 8 bits were made available for address purposes. A start code of four 'ONES' followed by a 'ZERO' preceded each address, which in turn was split into two blocks of 3 bits followed by a block of 2 bits, each block being punctuated by a 'ZERO' bit. A final 'ONE' acted as an end of message indicator as before (Fig. 6b).

A check was applied to incoming data at the vehicle end of the link to ensure that each message fitted into this format; if it did not the message was disregarded. A vehicle recognizing an address as its own would formulate a reply from the data contained at that moment in its data register and then transmit this back to base. Thus, as before, all that is required to complete a location system is to set suitably coded up-to-date vehicle position information in this register.

Further navigational methods

Having already established that a simple wheel revolution counting system could be adequate for vehicles operating in mode (c), and that using a faster data link large fleets could be controlled, attention was turned to the more difficult problem of vehicles of the mode (b) type.

A clue to the solution to this problem might be found by returning to the subject of conventional navigation. So far we have explored the dead-reckoning approach and have found comfort in this for vehicles which are restricted to given routes. A little thought will show that this, combined with fixes arranged at strategic points within a given urban network, could provide a suitable location system.

Consider a vehicle which is operating in mode (b) travelling along a straight road in an ordinary town. Before very long it is likely to reach a crossroads and is presented with an element of choice. It may continue on past the junction or it may turn left or right; somehow the location system must determine which of these possibilities has been chosen. The system previously described, within limits, measures distance travelled, and it can be assumed that the base station has good knowledge of the town geography. Such a system can determine when a vehicle arrives at a particular junction but is unable to decide by which port it leaves. Some means must therefore be provided to overcome this dilemma.

As a first approach, a measurement of the degree of turn could be considered and, given sufficient accuracy, this method has a lot to recommend it. However, most practical methods of achieving this to the required degree of accuracy entail somewhat complex modifications to the mechanical parts of the vehicle. It is interesting to note that the principle of the ancient Chinese "southpointing chariot", which was chronicled some 5,000 years ago, provides one form of solution to this problem[1,2]. Further methods rely upon magnetic compass indication, but the presence of a large and variable number of oddly assorted metallic vehicles in the immediate vicinity rather precludes such an approach in this type of environment.

Location systems using beacons

Another way of achieving the desired result is to provide a beacon at the side of each road leading away from the junction. If each beacon is uniquely coded and can be interrogated from the vehicle then, knowing the sites of each beacon, the base station is able to follow the progress of the vehicle. Again, if enough beacons are erected, one on each lamp-post perhaps, quite a good degree of positional accuracy may be obtained; sufficient for most practical purposes without the use of other methods such as wheel-revolution counting. This type of system bears a strong resemblance to railway 'block signalling', the position of each vehicle being determined by which block it is in at any time. The blocks in this case need be only a few tens of feet long.

Until recently the use of beacons was not considered to be particularly viable for vehicle location, the economics of such a system being most unfavourable. Some advantage had been gained elsewhere by using beacons in conjunction with wheel revolution counting for largely route-plying purposes. But the sheer number of beacons required for mode (b) vehicles (several at each junction), even when combined with wheel revolution counting, demands a beacon which is cheap to produce, install, maintain and run. Although in the practical case a high initial cost may be defrayed to a considerable extent by a leasing agreement with other operators. For example, once installed, beacons could be used by police, bus and tram operators, taxis, delivery van fleets, etc. As such a system can easily be made mutually exclusive, the interference problem need not arise.

Beacons, and the way they are interrogated, may take various forms. The requirement is basically one of short-range communication and any of the standard methods from optics to ultrasonics can be used. The beacons themselves may be either active or passive. Active beacons are those which continuously emit pre-coded information and the marine lighthouse is a common example of an active beacon. Passive beacons transmit their address signal only when triggered by some external source and may be further classified into two sub-groups, passive-active and passive-passive, depending upon the method by which they are powered.

Passive-active types need some form of power to drive them even in the quiescent state. A radio beacon is usually of this form in that the receiver must remain on at all times, other than when the transmitter operates, in order to amplify and detect the triggering signal. For this purpose a source of external power is required. The wartime I.F.F. transponder is a practical example of this type of beacon.

Passive-passive beacons are completely inert, requiring no external power at all while waiting to be triggered. They derive the power which enables them to reply, solely from the external energy required to trigger them. For example, the rear reflector on a motor car needs no external power supply but emits a red-coded 'address' when triggered from a white-light source. Clearly, the advantages gained from using passive-passive beacons in a vehicle location system are considerable. No batteries or mains supply are needed so that installation and maintenance effort required are minimal and they cost nothing to run. At the present time the use of optics allows the most practical method of applying a passive-passive beacon solution to the vehicle location problem.

In one such system beacons consisting of horizontal reflective strips are illuminated by a vertical scanning beam emanating from the vehicle. The reflected reply is de-scanned, detected by a photo-cell, and after amplification provides an electrical analogue of the beacon code. This information is stored in a register until a demand is received over the v.h.f. data link, whereupon the beacon address is transmitted back to base. Knowing that the code is that of a particular beacon, and the beacon is situated in a certain place, the base station now has the exact position of the vehicle and can display this to the operator in some pre-arranged form.

Using optical passive-passive methods the beacons become purely reflective, and a first requirement is to devise some way of coding each beacon in order to give a unique reply. In the present system this is achieved by using strips of retro-reflective material. Retro-reflection is the property of a material to reflect a beam of light back along the axis of the incident beam. A ray of light incident on a plane reflecting surface is reflected at the same angle as that made by the incident beam to the normal, but on the opposite side of the normal. A plane surface is only retro-reflective to light reaching it along the normal.

Beacons could be coded digitally by using the presence of a strip to represent 'ONE' and no strip to represent 'ZERO', but this simple method is unsatisfactory for the following reasons. In practice a vehicle may pass a beacon at any distance across the carriageway, it may be travelling in the inside lane when interrogating one beacon but in the outside lane when passing the next. Therefore the beacon reading equipment must be able to read and store the beacon information at distances ranging from a few feet to a few tens of feet. This means that the frequency and pulse width of a reply can vary considerably in the practical case, because the angle subtended by the beacon alters with distance from the reading head.

Further difficulties occur due to a difference in the installed heights of various reading heads. Clearly, for a system to be practicable it must accommodate reading equipment which can be fitted on the roof of a Mini or on top of a double-deck bus. In this respect too it is most desirable to allow a wide tolerance on the installation height of the beacons themselves. For these reasons the beacons must be capable of being read at a variety of vertical angles and the reading equipment must be able to extract the beacon code under these conditions. This possibility of a slant presentation results in a further change in pulse width and frequency during a single scan, but both these difficulties can be overcome by devising a method of extracting both timing and code information from the reply signal.

Modulation of the light beam by frequency shift (two colour), polarization, or pulse-width techniques afford ways of deriving the required information, and both methods have been used in the system under consideration. But before discussing these methods in greater depth some thought must be given to how the beacon is to be read.

As previously mentioned, the beacon is scanned vertically. The light source needs to supply an extremely well collimated beam, one which can project a spot of less than about $\frac{1}{4}$-in diameter at a distance of about 40 feet. This is a very exacting requirement but one which can be readily met by using a low-power laser. The monochromatic property of the laser, used together with suitable filtering, also provides a considerable signal-to-noise advantage. The beam is deflected on to an eight-sided mirror-drum which rotates at about 2,500 r.p.m., thereby scanning any beacon in its path. The need for retro-reflection now becomes clear, the return signal so formed is de-scanned by the mirror-drum and deflected on to the sensitive area of a photo-diode (Fig. 8). This is all the optics required for the pulse-width coded beacon as the necessary information can be extracted from the photo-diode output electronically, and we are now able to consider the operation of this system in greater detail.

Width-coded beacons (Fig. 9a) are divided into $1\frac{1}{2}$-in horizontal zones, the first inch of a 'ONE' zone and the first half-inch of a 'ZERO' zone consist of reflective material (Fig. 10a). The signal received from such a beacon is of the form shown in Fig. 10b, and this is used to gate an up-and-down counter which normally counts a faster-running pulse signal. A positive-going edge causes the counter to count up until a negative-going edge reverses the count. The state of the counter is read out at each positive edge and if it contains a positive number the appropriate register position is written to 'ONE' if negative a 'ZERO' is written. Immediately after reading the counter is reset ready for the next signal bit. This process is shown diagrammatically in Fig. 10c.

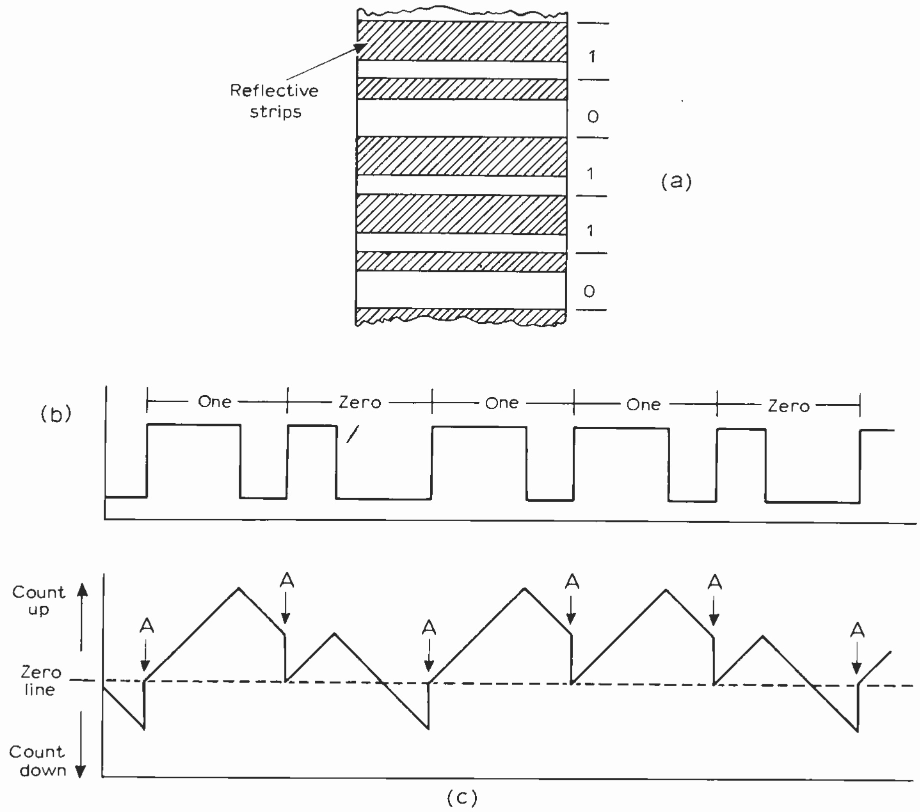


Fig. 10. (a) Width-coded beacon, (b) width-coded beacon waveform, and (c) the counting operation of up/down counter. The counter is read and then reset at points marked A.

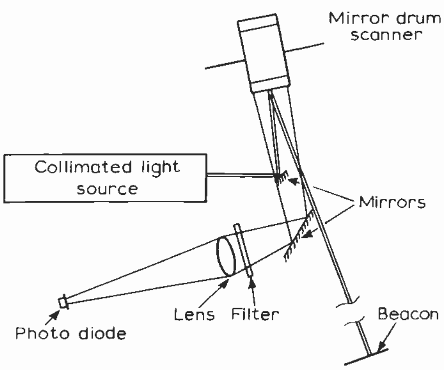


Fig. 8. Basic optical system.

Beacons for the polarized light method consist of alternate equal width strips of reflective and non-reflective material (Fig. 9b). Coding is achieved by covering certain reflective strips with polarizing material and the whole beacon is scanned by a beam of polarized light. This causes a rotation of direction of polarization to occur on light returning from the covered strips while no change occurs on the uncovered ones. By splitting the returned beam after de-scanning, and passing each split beam through suitably arranged polarizing filters, the output of one of the two photo-cells detecting these signals will contain information from each strip, and the other only that from the polarized strips. The signal from the second cell may therefore be clocked into the register using the output from the first cell as a strobing signal.

All estimates existing today predict a continued rise in the number of road vehicles over the next 50 years. In Britain alone this number is expected to top 35 millions. With this order of density it becomes transparently clear that the present "Brownian motion" of traffic can no longer be tolerated and orderly control will have to be applied to a greater extent.

Over the past year several small location systems have been demonstrated. Wheel revolution counting, polarized and width-coded beacon methods have been shown in an attempt to illustrate that solutions do exist to the somewhat difficult problem of vehicle location. The first faltering steps have been taken into the realms of practicability, and it is not unreasonable to suppose that great strides will now be made in this particular subject.

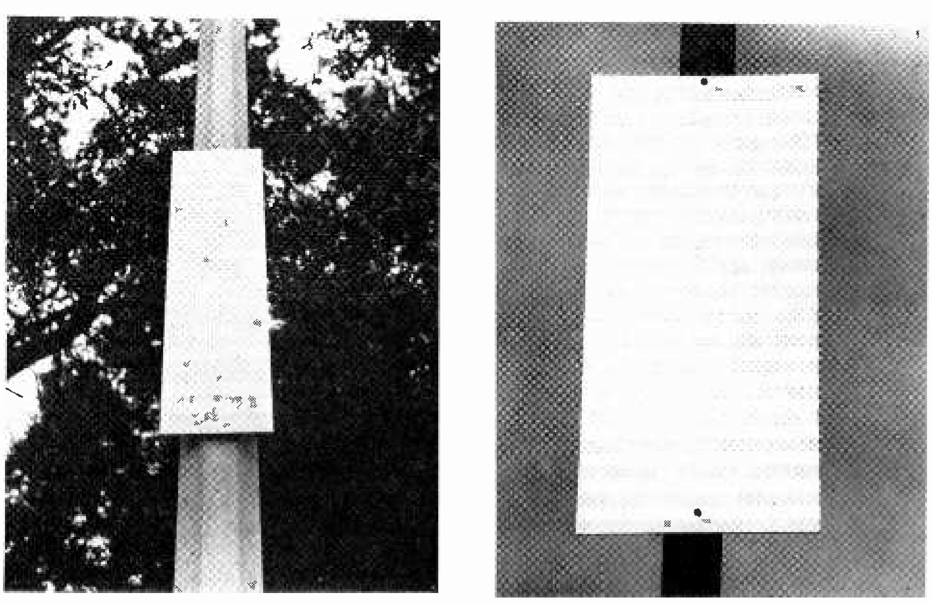


Fig. 9. (a) Width-coded beacon and (b) beacon for polarized light method.

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1 'Differential Gears" by Ernest F. Carter, *Design and Components in Engineering*, 9th February 1967.
2 "The History of Marine Navigation", by Per Collinder (translated by Maurice Michael), Batsford 1954.

Letters to the Editor

The Editor does not necessarily endorse opinions expressed by his correspondents

Electronics in medicine

In expressing my support for the editorial "Electronics in Medicine—the Future" in the June issue, I should like to draw attention to an underlying handicap retarding the development of biomedical engineering technology in the United Kingdom.

Medical electronics is a branch of biomedical engineering, a multi-disciplinary scientific activity. In 1958, following the initiative of Dr. Zworykin, the first of the International Conferences on Medical Electronics was held in which many kinds of specialists from the life and basic sciences have contributed. In hospitals of the National Health Service and Medical Faculties of the Universities, where direct collaboration between medical and engineering personnel is possible and desirable, there is still no recruiting, training or career structure for chartered engineers. Is it merely a rhetorical question to ask how the engineering side of biomedical engineering can be carried out without the participation of professional engineers?

Some hospital physicists, aware of the need for engineering in the medical environment, have retrained themselves as engineers. The situation in the universities is even less satisfactory, for the normal staffing structure in an electronics laboratory is the "one- (or two-) man-band", each one organized independently, one or more to a department according to the demand. There is no need to emphasize the consequent duplication of capital outlay and failure to utilize available skill.

It appears to the writer that the engineer is not yet accepted in biomedical engineering, and that non-engineers do not understand either the need for engineers, or the work they do. The remedy, of necessity, must originate from within the engineering profession. The initiative must be both practical and independent of external support or patronage. Only when engineers, as a profession, are making an indispensable contribution to the inter-disciplinary cause, also providing training and research facilities, will the engineer be welcomed by non-engineers. Five years ago, Dr. Zworykin proposed the setting-up of an Institute of Biomedical Engineering. May I as an engineer, propose that the Council of Engineering Institutions give consideration and support for the creation of such an institute?

R. E. GEORGE,
Physics Dept.,
Medical School,
Guy's Hospital, London S.E.1.

Your editorial in the June issue rightly draws attention to the difficulties faced by companies manufacturing medical electronic equipment, especially for the U.K. market. In it you suggest that equipment produced for the general user has potential applications in medicine, but the converse is just as often true.

The evolution of electronics from radio began before the incentive of war-time radar, helped greatly by physiologists and associates who asked for the impossible, and then provided it themselves. For example, between the wars, Matthews, Offner, Tonnies, Schmitt and others transformed the crude audio amplifier by adding differential inputs, increasing gain and stability, and extending the frequency response towards zero. Since the war, medical workers have been constantly breathing down the necks of solid-state designers, demanding higher input impedances, less noise and many other characteristics that have proved invaluable elsewhere.

It is not only in the fields of circuit techniques and components that biomedical engineering (as it is now known) has influenced electronics in general. Many items of hardware developed specifically for medicine have found other applications, perhaps after minor modifications.

This company manufactures a patient monitoring system that has evolved from initial work in a medical research establishment, but the specially developed indicators and chart printers are now proving to be of a great interest to instrumentation and control engineers in many industries. By exploiting this greatly extended market, it is possible to provide the hospitals with what they want at a price the nation can afford.

ANTHONY S. VELATE,
T.E.M. Engineering Ltd,
Crawley, Sussex.

Rationalizing trade associations

I was most interested in the editorial in your issue for July because a rationalization of the many trade associations in the electronics field is long overdue.

In 1956 I was appointed secretary of R.C.E.E.A.—now E.E.A.—and a year later I put up a scheme (which would probably have led to my immediate retirement) for the administration of all these associations within the general confines of the British Electrical and Allied Manufacturers' Association. I learnt very quickly that this was a non-starter at that time, although the economic factors and almost entirely common membership of E.E.A. with B.E.A.M.A. made it a logical development.

There were two reasons why it was not acceptable then. One you have high-lighted. That is the reluctance of some associations (or their council members) to surrender their autonomy. The second was never publicly recorded, but was, the intense jealousy between member companies, which led them to believe that their own interests were more important than those of the industry, or indeed of the nation. I am convinced that the latter has been the real reason why no rationalization of the electronics industry representation has so far taken place.

However, the last few years have seen a major re-grouping of the companies in this field, and there are consequently fewer heads of businesses to be convinced of the obvious economies to be achieved by ending the proliferation. One would therefore hope that the rationalization of the trade associations, which the E.E.A. is apparently trying to achieve, will go forward. The resulting benefits to the nation should in the long term benefit the member companies also.

The top men who would comprise the "voice that could speak to governments, other associations and foreign organizations" are unlikely to have the time to do it in B.E.A.M.A. *and* E.E.A. etc. Perhaps after thirteen years the jealousies are becoming less pronounced, and the powerful voice may yet emerge.

H. E. F. TAYLOR,
Torquay,
Devon.

Mobile radio and amateur bands

In your issue of July 1970 you quote from the recently published Annual Report (1969) of the Electronic Engineering Association: "The lack of spectrum space continues to be seen as the most likely factor which could seriously limit the expansion of mobile radio communications. Negotiations have therefore now begun with the Ministry of Posts and Telecommunications to secure the use of the 68 to 71.5 MHz and 420 to 450 MHz bands."

The reason why our Association has suggested to the Ministry of Posts and

Telecommunications that 68 to 71.5 MHz should be made available to mobile radio communications arose from the recent announcement that the GEE navigational chain which operated in that band had closed down. In making the recommendation I do not think that any of our members were aware that this part of the frequency spectrum was used by amateurs. Indeed, it seems that the allocation was made without the usual formality, and only upon a non-interference basis.

In regard to the 420-450 MHz band, I feel sure that the amateurs will understand our concern as an industry that substantial frequency space, without which mobile radio cannot expand, is being made available above 470 MHz in the United States and below 450 MHz on the Continent. Mobile radio activity and the size of the industry it can support, and the amount of research and development that the industry can support, are in the last analysis dependent upon the adequate availability of frequencies. If the problem of expansion in the all important u.h.f. sector is not solved then leadership in product development and export will undoubtedly pass from Britain to those industrial countries who are now receiving generous new allocations. It is against this background that our members question whether the country can afford, even on a non-interference basis, the allocation of 22 MHz to amateur television.

Our members have the highest regard for amateur activities including their brilliant pioneer work above 1.5 MHz nearly fifty years ago, and would not wish to see them deprived of all facilities in the u.h.f. bands. We should welcome discussion on the subject with representatives of amateur associations.

R. A. VILLIERS,
The Electronic Engineering Assoc.,
London, W.1.

Class AB—some questions

Following the two articles on a class AB amplifier design by Mr. Linsley Hood and also the correspondence in the August issue, we would like to raise several points concerning the specification.

Total harmonic distortion is specified as less than 0.02% at all power levels below maximum output, but this is presumably (see Figs. 6 and 7) only at 1kHz though not specified as such. What are the distortion levels at 100Hz and 10kHz at full output, for example?

When quoting a noise level for the amplifier, the noise bandwidth of the measurement was unspecified thus rendering the result as meaningless as quoting a frequency response without limits (e.g. ± 3dB).

A value for "square-wave transfer distortion" is given as 0.2% at 10kHz but the power level is not specified. As "square-wave transfer distortion" is a non-standard quantitative measurement, for the result to be meaningful, an explanation is required as pointed out by Mr. Gibbs in his letter in the August issue.

Also results for other amplifiers, for example a good class B amplifier, would be useful for comparison.

MARTIN SMITH and
H. P. WALKER,
Southampton, Hants.

Notwithstanding the perfection of Mr. Linsley Hood's latest amplifier in practice, I would differ with him over some of the points he raises in the July issue.

A Darlington pair has a lower mutual conductance than the output transistor on its own. The converse can only be true of the complementary pair configuration. His first paragraph attributes a higher value to both pairs.

The overall linearity of the output stage of his Fig. 2, when driven from a genuinely low source impedance, does depend on the quiescent current contrary to his expectations. A high drive impedance is the answer, with a low inter-base impedance. This does not impair the cut-off performance as the conducting transistor presents a low base-emitter impedance to the one being cut off.

The output stage of Fig. 3 operates between the common emitter and the common collector modes. The true emitter follower of Fig. 2 has an inherent distortion of about 100 times less than Fig. 3, provided that the source impedance is low enough and the quiescent current is appropriate. Infinite values of bootstrap capacitance are necessary to secure pure common emitter operation; this circuit is predominantly common emitter above 30 Hz. His calculation of class A output power assumes that the output transistors have a constant mutual conductance. Due to the bend in this characteristic at low collector currents they do not cut off as soon as expected. The class A output of either version is nearly 2 amps pk-pk. Using a standing current of 100 mA and no emitter resistors, a class A output of over 5 amps pk-pk is available. (The traditional definition of class A does not preclude current ratios between the two halves of 10^8.)

A high class A power is not, *ipso facto,* a particular virtue. The correct quiescent current is related to the linearity of the output stage under dynamic conditions, and this ought to be significantly lower than that required by full class A operation, in a good class AB design.

The mutual conductance of MJ 481/491 with 0.82-ohm emitter resistors is 1 mho at high currents; this falls to 0.5 mho at a collector current of around 20 mA. If $Tr_{3, 4, 5}$ have high current gains, so that the drive impedance really is low, this is the optimum quiescent current with a bandwidth of a few kHz. Higher quiescent currents worsen the performance. A current of 200 mA is undoubtedly right for bandwidths greater than this, but no compromise would be necessary if the drive impedance was high enough for all combinations of transistors.

Poor matching of the output transistors is extremely unlikely to cause any noticeable deterioration of the performance, except to a distortion meter; low gains may even be advantageous in certain cases. Full class A operation is unnecessary in both these circumstances.

My final point concerns the avoidance of temperature-compensation in the biasing of the output stage. The penalty for this is very poor thermal stability in the 8-ohm version.

D. L. D. MITCHELL,
University of Bradford.

I.Cs in stereo pre-amplifiers

We were very interested in the article in July 1970, by L. Nelson-Jones, describing an application of the RCA CA3048 as a stereo pre-amplifier.

The original RCA stereo pre-amplifier circuit was intended for use in low-cost applications and had a few shortcomings as observed by Mr. Nelson-Jones. A later RCA report published in June 1969 (A Monolithic Integrated Circuit for Stereo Pre-amplifiers, by L. Kaplan) describes several circuits which produce comparable performance to the circuit described by Mr. Nelson-Jones but use passive tone controls. The problem of signal-to-noise ratio was overcome by increasing the loop gain of the first pre-amplifier whilst still maintaining adequate overload capability.

The CA3052 is capable of producing a similar noise performance to the CA3048 when used as a stereo pre-amplifier provided amplifiers A2 and A3 are used for the magnetic inputs. These amplifiers are tested in production for equivalent input noise with full R.I.A.A. compensation, and amplifiers A1 and A4 are checked with a simple 'C' filter connected at the output. We should be happy to provide any further applications information to interested readers.

L. R. AVERY,
RCA Ltd,
Sunbury-on-Thames,
Middx.

Jupiter probe

In the feature "News of the Month" in your August issue under the heading "Space Probe to Jupiter" you state that "This planet is believed to be the only one in our solar system which radiates more energy than it absorbs from the sun, current measurements indicating about twice as much." This is not true as all planets containing any radioactive material must radiate more energy than they absorb from the sun. No other planet, however, radiates as large an excess as Jupiter.

A planet must radiate enough energy to maintain its thermal equilibrium, and it is reasonable to assume that any planet will contain some radioactive core material, the decay of which will produce heat energy. It is evident, therefore, that to maintain thermal equilibrium a planet must radiate as much more energy than it absorbs from the sun as it generates itself in internal radioactive decay.

JAMES M. BRYANT,
Cheltenham, Glos.

Transistor Breakdown-voltage Meter

Constructional details of a useful instrument which provides direct reading of junction breakdown voltage at fixed reverse currents

*by Jens Langvad**, *M.Sc. (Danish)*

Designers of power amplifiers, power supplies or any equipment using transistors working at high supply voltages often require to know the actual breakdown voltage of the transistors. The instrument described here was designed to replace the conventional method of just increasing the voltage slowly to see what happens.

The maximum collector-to-emitter or collector-to-base voltage specified by the transistor manufacturer is often grossly understated. This leads the circuit designer into buying expensive high-voltage transistors in cases where a standard type may do the same job just as well. The cheap BC107, for instance, usually exhibits a collector-to-base breakdown voltage [$V_{(BR)CBO}$] in excess of 80V, although the stated maximum operating voltage is only 45V.

The reason for this is that in the planar manufacturing process it is not easy to keep the $V_{(BR)}$ within narrow limits. By stating a $V_{(BR)}$ well below the production average, but still above the need of most customers, allowance is made for a greater number of freaks and a lower rejection factor.

Breakdown characteristics

For an ordinary silicon planar junction transistor three breakdown characteristics are of interest. These are: (1) the $V_{(BR)EBO}$, (2) the $V_{(BR)CBO}$ and (3) the $V_{(BR)CEO}$ voltage-current curves. (In accordance with accepted terminology the *"O"* stands for open and refers to the third terminal.) The first is quickly dealt with; the *EB* junction breaks down like an avalanche zener diode with a knee voltage of 6-12V, according to type (most BC 107s lie quite accurately at 8V).

The *CB* diode breaks down in a similar manner, but the knee voltage is much higher. The $V_{(BR)CBO}$ is the most important of the three, as it determines the maximum peak-to-peak collector voltage swing one can expect from the transistor.

The *CEO* characteristic is shown in Fig. 1 together with the *CBO*. (There is no scale factor, as this depends entirely on the type of transistor investigated.) At very low currents the two curves are seen to coincide, but after reaching the knee the $V_{(BR)CEO}$ drops with increasing current, thus exhibiting a negative resistance characteristic. If, however, an external resistance is present between the base and the emitter the *CE* curve approaches that of the *CB*. If, for a low-level transistor, R_{BE} is less than 100kΩ (and this will be the case in most amplifier applications) the two curves may be considered as coinciding.

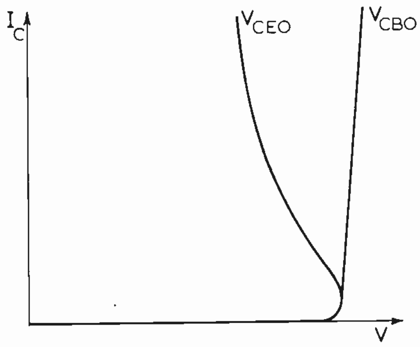

Fig. 1. Typical breakdown characteristics.

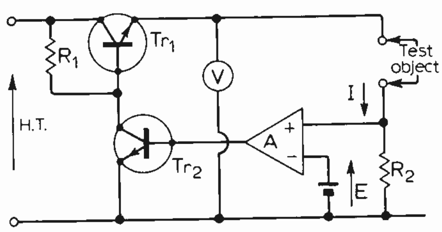

Fig. 2. Principle of breakdown-voltage meter.

Breakdown-voltage meter

The breakdown-voltage meter itself comprises a simple d.c. supply giving the unusually high output voltage of 200V, in connection with a current limiter covering the equally unusual range of 0.1μA—1mA in decade steps. In Fig. 2, Tr_1 and Tr_2 are high-voltage type transistors. *A* is a differential input amplifier working as a voltage comparator and *E* is a reference voltage. When no current flows in the test object the output of *A* is zero, so that Tr_2 is cut off and the series transistor Tr_1 saturated through R_1. When the test object is connected, the current flowing through it will produce a voltage drop across R_2. This voltage cannot exceed *E*, because if it did, the voltage comparator would turn over, saturate Tr_2 and thus drive the supply voltage towards zero.

Provided that the input impedance of *A* is much greater than R_2, the current in the test object is seen to be limited to $I = E/R_2$. Thus $I = 1\mu$A and $E = 0.5$V require $R_2 = 0.5$MΩ. The voltmeter *V* indicates the voltage required to drive *I* through the test object, plus the reference voltage *E*. But the latter is small enough to be neglected.

Practical circuit

The complete circuit diagram is shown in Fig. 3. Any transformer giving 200-250V and 6.3V may be used. The zener-stabilized low voltage supplies the voltage comparator and also works as a reference for the voltage limiter. (This is of the usual kind and serves only to limit the maximum voltage to full scale deflection on the meter.) There are two voltage ranges: 0-100 and 0-200V, and the voltage divider for the limiter and the pad for the meter are switched simultaneously by S_1. Transistors Tr_1, Tr_2 and Tr_3 are high-voltage devices, RCA type 2N440, but any type with a $V_{(BR)CBO}$ of 300V or more would do.

Transistors Tr_4, Tr_5 and Tr_6 constitute the voltage comparator. The f.e.t. input transistor provides the amplifier with an input impedance approaching infinity. The diode-resistor network in the input protects the f.e.t. against voltage surges. S_2 selects the current ranges 0.1μA—1mA. The selected value of current obviously depends on the type of transistor under test, but usually the $V_{(BR)CBO}$ is remarkably independent of the current at which it is measured. Generally speaking, one should use the low-current ranges for low-level planar transistors, and the two highest ranges for germanium or power devices. When measuring $V_{(BR)CEO}$ one can get a fair impression of the shape of the breakdown characteristic just by rotating S_2.

S_3 switches between positive and negative supply for n-p-n and p-n-p transistors respectively, and S_4 connects the transistor either *CBO* or *CEO*. In the mid-position the transistor is disconnected and the supply voltage turned down. This is achieved by S_{4C} shorting the base of Tr_1 to ground. S_4 could be a keyswitch biased in both directions, so as to ensure that the

* Philips Records Ltd.

Capacitors

The prefix *C* and suffix F have been omitted

1—3 × 16μ, 450V 2—125μ, 16V

3—125μ, 16V

Diodes

The prefix *D* has been omitted

1—BAX16 4—BAX16

2—BAX16 5—4 × 10D10

3—BAX16

Transistors

The prefix *Tr* has been omitted

1—2N3440 4—2N3702

2—2N3440 5—BC108

3—2N3440 6—2N3819

Resistors

The prefix *R* and suffix have been omitted.

1—560	11—2.2k
2—33k	12—5.6k
3—100k	13—4.7k
4—1k	14—560
5—100k	15—1k
6—91k	16—100k
7—9.1k	17—4.7M
8—10k	18—470k
9—1M	19—47k
10—1M	20—4.7k
P1—4.7k	21—470

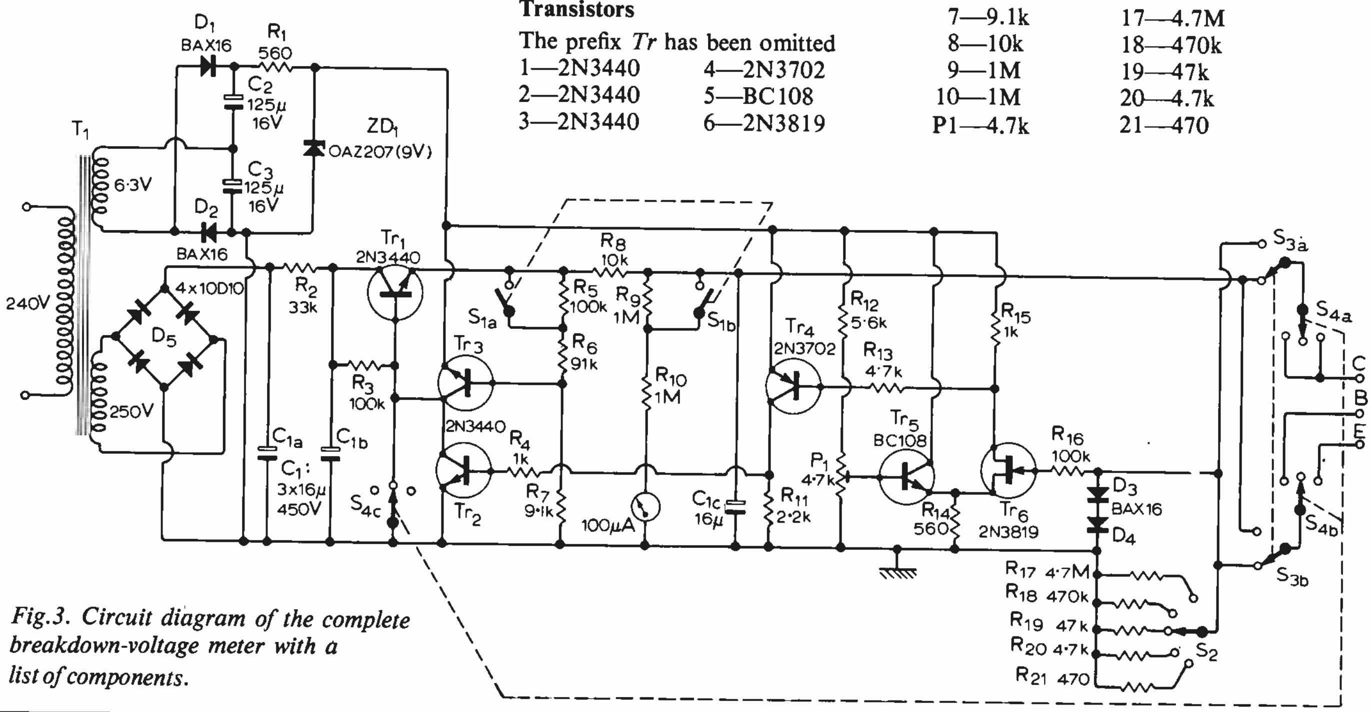

Fig.3. Circuit diagram of the complete breakdown-voltage meter with a list of components.

switch is always in the neutral position when the transistor is inserted. When S_4 is operated, the voltage does not jump but rises fairly slowly owing to the 10kΩ/16μF *RC* network. This is necessary in order to avoid dangerous current surges in the test object.

The test current may be adjusted by altering the base voltage of Tr_5. It is easily set to the correct value by means of a known resistor. With 470kΩ connected between *C* and *B* terminals, S_4 in position *CB* and S_2 in position 100μA, P_1 should be adjusted until the meter reads 47V. (100μA × 470kΩ = 47V). In this context it can be mentioned that the instrument is ideal as a megohm-meter, having a linear scale and covering a range of 0.1 to 2000MΩ f.s.d.

Mounting of the components is far from critical; yet it is recommended to put the whole in a closed metal box and connect this to circuit earth so as to avoid the disturbing influence of hum voltages on the high impedance input. However, insulate the box externally (a coat of paint may suffice), or be very careful with external test-leads, as a short from either of the connecting terminals to the case will prove fatal for at least half of the transistors.

If the components are mounted on a printed circuit board, trouble might be experienced with the low current ranges owing to stray currents on the board. This can be countered by keeping the gate of the f.e.t. and associated leads and components on a "stand-off", away from the etched side of the board.

A Quick Look at Display Devices

Information present inside a piece of equipment—be it a computer, a measuring instrument or what have you—is of no use unless it can be made to do something or it can be presented to the outside world in a way that can be understood by human beings.

Ordinary incandescent lamps can be used individually to show simple on-off or yes-no information or they can be used to illuminate a more complex legend, "No. 3 alternator failure", for instance. Early counter-timers used such lamps to illuminate the digits 0 to 9 either by edge-lighting perspex sheets or by illuminating an engraved or printed window. However the flexibility is limited and the current consumption is high, and most of the power consumed is wasted as heat.

Some of the above remarks apply to simple neon indicators, but here, low brightness and high-voltage replace the high-current and heat disadvantages.

Gas-filled indicators, such as the Nixie, have long reigned supreme in the instrumentation field where a numeric only readout is required but they are of no use when letters and words have to be displayed.

The cathode-ray tube is now much used to display alphanumeric information and the methods of doing so are as varied as they are numerous. Character generators for c.r.ts can now be purchased in integrated circuit form which use the seven-segment method of presentation. Seven lines are generated which can be used to assemble all the numbers from 0 to 9 plus a few letters. More complex electronics can be employed to produce on a c.r.t. upper and lower case alphabets, all the numerals and a varied assortment of symbols. Although expensive with complex scanning, blanking, decoding, addressing and storage circuitry this type of display is much used, often in association with computers.

The single incandescent lamps and the neon lamps can be grouped together in a 5 × 7 matrix to provide a display capable of showing a full range of alphanumeric characters and symbols. The array of lamps in this case is driven by a logic network which receives coded input information and determines which lamps have to be lit to display a given character. The same 5 × 7 matrix approach can be, and is, used with c.r.ts, here the matrix is formed by dots on the tube face.

The latest contender in the field of displays is the light emitting diode and our front cover this month shows an integrated array produced by Standard Telecommunication Laboratories Ltd which, as can be seen, is based on a 5 × 7 matrix. These lamps are forward biased diodes which emit a narrow band of light by the recombination of injected holes and electrons across the energy band gap of a single crystal material. The diodes are manufactured from a mixture of gallium phosphide and gallium arsenide. They have the advantage of requiring voltages and currents that are compatible with those normally found in transistor and integrated circuitry.

Another display medium which we will be hearing a lot about in the future is liquid crystal; a substance which can be held in a layer between two transparent surfaces and turns opaque when an electric field is applied.

Active Filters

13. Applications of the active-ladder synthesis

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

The application of the active-ladder synthesis to low-pass and high-pass filters with and without zeros in the stop band, and to simple band-pass filters is discussed.

Butterworth ladder filters

Because only moderate Q factors are needed, Butterworth responses are in general suitable for realisation by factors. Nevertheless, by misalignment of the factors on the frequency axis or by errors in Q factor, an appreciable slope can be introduced into the pass-band response by component errors which would cause very little change in pass-band response in an equally terminated filter (much as in the example shown in Fig. 3 of Part 12).

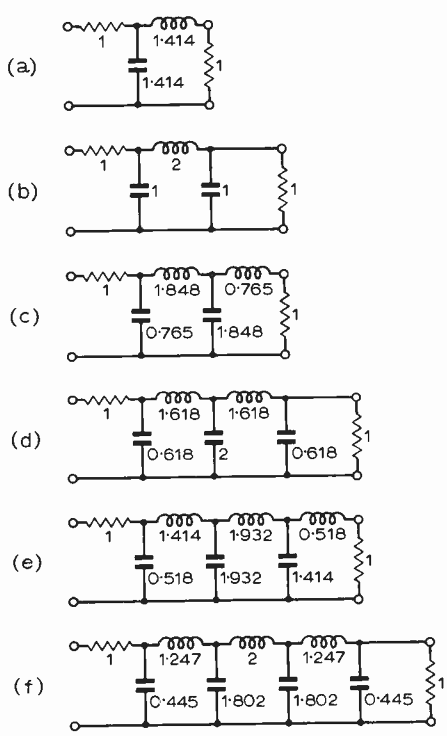

Fig. 1. Butterworth ladder filters normalised for ω_c = 1 rad/s.

Now although Butterworth responses have relatively rounded corners, over much of the pass band they show a high degree of flatness, e.g. 5th-order response is 1% down from the zero-frequency level at about 0·85 of ω_c. If this degree of flatness is to be achieved in practice, therefore, without calling for a tight tolerance on the values of the reactances (or their equivalents in an active filter, the values of the Ts— or CR products), an equally terminated ladder filter is the preferable choice. Diagrams for orders 2 to 7 are shown in Fig. 1 (although the filters of orders 2 and 3 do not really come within the scope of the present argument) and it is interesting to notice that the normalised reactance values coincide with the reciprocals of the Q factors (Part 9, table 6).

The theoretical structure for a 5th-order active ladder has already been given (Fig. 13(d) of Part 12) and may be turned into practical form as in Fig. 2 herewith. In this diagram the inverting stages have been moved to new positions to reduce the number of resistors connected to the second and fourth integrators.

If the frequency response is compared with that of the 5th-order constant-k filter, Fig. 6, it is seen that the shapes are little different, except that the constant-k filter gives a more rapid fall into the cut-off region. When normalised for -3dB at 1 radian/second the element values of the constant-k filter are as given in Fig. 3, and comparison with those of the Butterworth filter give an indication of the tolerances allowable without losing flatness in the pass band or attenuation in the stop band.

In a filter such as that shown in Fig. 2, if $R_{in} = R_5$ the voltage gain is the same as that of the passive model, $\frac{1}{2} = -6$dB. There is no necessity to stick to this figure, however. First R_{in} may be varied over a considerable range. Secondly the gain of any stage in the forward path may be changed provided the gain of the surrounding backward path or paths is/are changed in inverse ratio. Thus the signal level at the output may be raised without altering the shape of the response curve by reducing R_1 and increasing the lowest-positioned of the three resistors R' by the same ratio; by reducing R_2 and increasing the already mentioned resistor marked R' *and* the lower of the two resistors marked R_3; etcetera. Such changes assume, of course, that each affected amplifier has sufficiently high internal gain to behave still sensibly ideally.

As the first and last integrators have a feedback resistor connected across the capacitor, they provide in effect a simple lag and an unwanted sign change. More economical ends may be used, therefore, as shown in

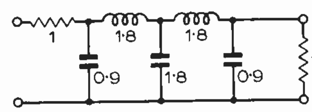

Fig. 3. Constant-k filter normalised for -3dB relative to zero-frequency level at 1 rad/s.

Fig. 4. These are lag and integrator circuits (Part 5) and they replace the end two integrator loops of the basic active ladder. The input resistor is now part of a time constant and so cannot be varied independently. Other means of varying the overall gain remain, however.

For a sharper corner and increased attenuation in the stop band Chebyshev filters may be used, and element values may be taken from reference works (e.g. Ref. 1). There is also something to be said for taking as a model a "classical" constant-k filter.

Constant-*k* filters

Constant-k or A-type filters (Ref. 2) have attractive regularity, Fig. 5, which allows all the integrators of the active ladder to have the same T (except the end ones, which are of half value). When the central elements

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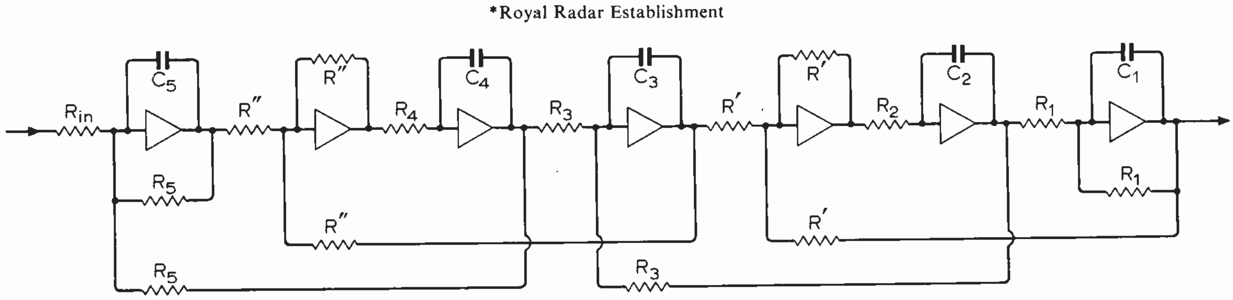


for ω_c = 1rad/s

$T_1 = C_1R_1 = 0{\cdot}618$; $T_2 = 1{\cdot}618$; $T_3 = 2$; $T_4 = 1{\cdot}618$; $T_5 = 0{\cdot}618$; R', R'' and R_{in} arbitrary

Fig. 2. Active 5th order Butterworth filter modelled on Fig. 1(d).

of the passive models are designated $2L$ and $2C$, as shown, the nominal cutoff frequency is given by

$$\omega_n = 1/\sqrt{(LC)} = 1/T \text{ (say).} \qquad (1)$$

Hence, since $R = \sqrt{(L/C)}$, $L/R = CR = T$. In the basic active ladder, therefore, all the integrators have CR products equal to $2T$, except the end ones which have value T.

From the responses shown in Fig. 6 it can be seen that these filters are not of equal-ripple type. They are not, therefore, optimum designs: a Chebyshev filter of equal order and with the same maximum depth of ripple in the pass band gives a steeper initial fall into the cutoff region, though the ultimate rate of fall is necessarily the same; or alternatively for the same steepness of initial fall the Chebyshev filter gives a smaller maximum depth of ripple.

The 3rd-order filter is identical with a Butterworth filter; the high-frequency and low-frequency asymptotes intersect at the nominal cutoff frequency, and $\omega_n = \omega_c$ (the

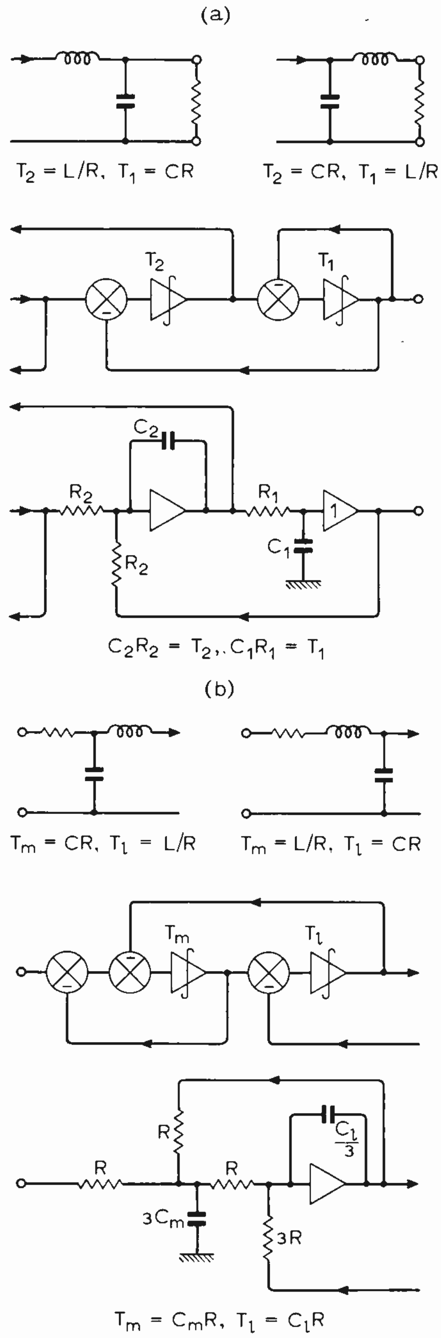


Fig. 4. Methods of making economical ends.

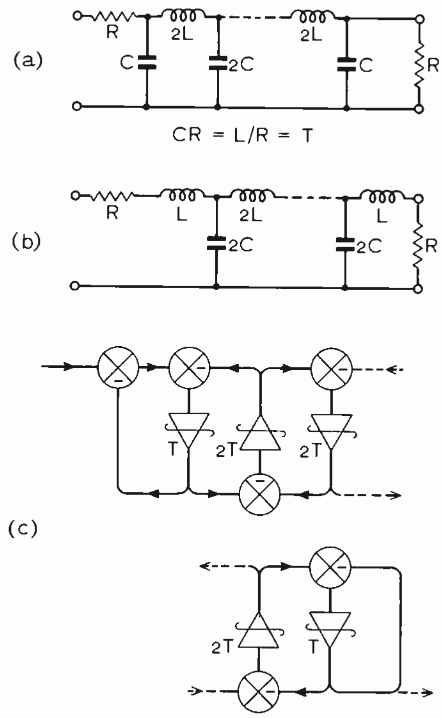


Fig. 5. Constant-k low-pass ladders.

−3dB frequency). As the order increases, an increasing proportion of the reactances (or integrators) have the value 2, and the high-frequency asymptote moves down the frequency scale. This causes the initial fall into the cutoff region to be steeper than the asymptote: for the 5th-order about 35 dB/octave, compared with 30 dB/octave; for the 7th-order about 60 dB/octave, compared with 42; for the 9th-order about 90 dB/octave, compared with 54. By the 9th-order a pronounced and unpleasant ripple has appeared towards the end of the pass band; but up to the 7th-order the amplitude response is quite good enough for many purposes, and the fact that the integrators can be made up using a number of resistors all of one value, and a number of capacitors also all of one value, can make these filters a convenient choice in experimental work.

Also shown in Fig. 6 are curves for the voltage at the output of the centre integrator, which is the replica of the voltage across the centre capacitor (if there is one) of the passive model, or of the current through the centre inductor. These show an increasingly large resonant peak as the order increases. This means that the centre integrator could overload before full output is reached at the final integrator. This possibility can be avoided, however, by increasing the forward gain of one or more of the loops at the output end of the filter and reducing the gain of the feedback parts of the loops as already described.

Some peaking occurs at the centre of Butterworth filters, but to a less degree as they are less resonant. This can be seen by comparing the Q factors of a Butterworth filter (Part 9, Table 6) with those of a resistance-terminated constant-k filter given herewith (Table 1). The high Q factors are, of course, the price paid for the steeper initial cutoff, and if the integrators do not have high zero-frequency gain this will show as an imperfection in the realised response more than with a Butterworth filter.

Effect of finite gain

In a realisation by factors it is easy to estimate the effect of finite gain and make correction by adjusting element values (provided more than the minimum necessary

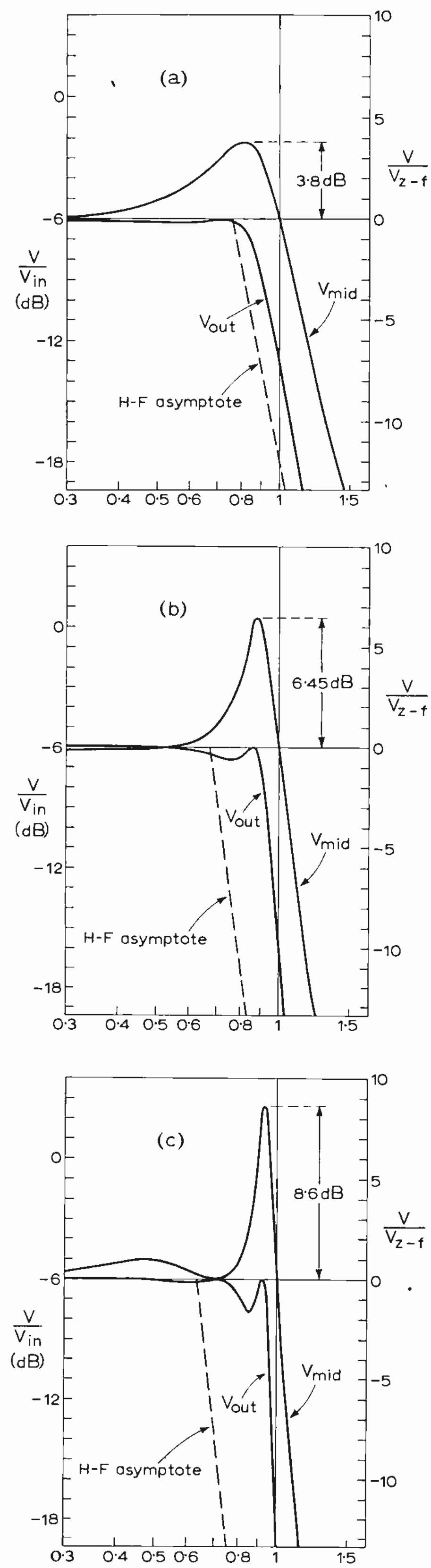


Fig. 6. Constant-k filters: output voltage and voltage at centre. (a) 5th order; (b) 7th order; (c) 9th order.

gain is available)—see Part 9. In a ladder it is a straightforward matter to calculate the effect of finite gain in an integrator, which as shown in Part 7 is equivalent to finite Q in the corresponding reactance. But when the model is an equally terminated filter, and all but the reactances should be of infinite Q, no adjustment of element values can compensate for finite Q in these reactances or finite gain in the corresponding integrators. In the active ladder, the loss may be compensated by raising the effective amplifier gain to infinity by positive feedback. This may be applied through a sign inverting amplifier, as shown in Fig. 7; or

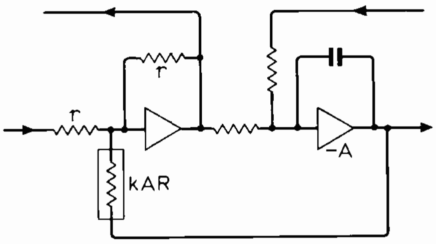


Fig. 7. A method of applying positive feedback to compensate for finite gain in an integrator.

if a differential input terminal can be found (e.g. the emitter of the input transistor) the positive feedback may be applied directly over the integrating amplifier. What is aimed at is that each of the inner resonant loops when freed of loading from the rest of the filter should have infinite Q, and in practice it is sufficient to apply positive-feedback correction of approximately double magnitude to only one integrator of each loop. It need hardly be said, however, that nowadays the best course of action, at any rate for a low-frequency filter, is to use an operational amplifier of ample intrinsic gain so that no compensation is needed. It should, of course, not be forgotten that capacitors do not have zero loss. But the availability of polystyrene capacitors showing Q factors from 2000 to 5000 makes it unlikely that much difficulty will be experienced on this score; and polycarbonate capacitors showing Q factors of about 500, and having the advantage of much greater compactness in capacitances of 0·1 μF and over, are usually quite satisfactory.

The effect of finite amplifier gain in distorting the pass-band response is shown by some experimental results obtained by the present authors many years ago. At that time transistors were in a very primitive state, and the only sensible choice of amplifier seemed to be a single pentode valve, Fig. 8. With the loss in the coupling and mixing networks it is probable that the effective amplifier gain was 50 to 100, and the measured response is shown in Fig. 9, curve (b). To increase the effect the valves were connected as triodes, giving an effective gain of probably 10 to 20, and the result is shown in curve (c). The curves have been adjusted vertically for easy comparison, i.e. the extra zero-frequency loss is not shown, and it is clear the main effect is felt near the "pole" frequency of the highest-Q factor and indicates a damping down of the Q of that factor.

Adding zeros in the stop band

The performance of a filter is often improved at small expense by adding zeros in the stop band, as mentioned in earlier parts; and in a passive low-pass ladder filter a zero is obtained if either (1) a series arm inductance is tuned with a parallel capacitance (so that, ideally, the impedance of the arm becomes infinite at the required frequency) or (2) a shunt arm capacitance is tuned by a series inductance (so that the impedance of the arm tends to zero at the required frequency). Strict correspondence between the internal working of the passive filter and an active counterpart would require an extra integrator for every added reactance. However, the present writers' colleague, Dr R. L. Ford, has shown that satisfactory results can be obtained by making some additional interconnections to the simple (all-pole) structure, in much the same manner as

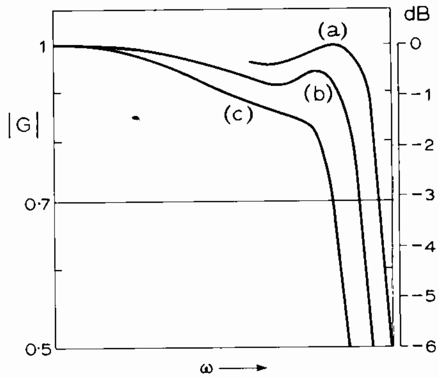


Fig. 9. Responses of 7th-order low-pass active ladder filter: (a) theoretical, (b) measured using pentodes, (c) measured using triodes.

notch response can be obtained from a two-integrator loop (Part 7).

The basic idea is that in Fig 10, for example, the current in C_2' is proportional to $(pV_3 - pV_1)$ and therefore to $(I_3 - I_1)$, the difference of two quantities which, at least potentially, are already available in counterpart in the active system. Hence the counterpart of the required difference $(I_3 - I_1)$ can be formed and added with a suitable scaling factor to the counterpart of I_2 to represent the addition of I_2' to I_2 without the use of any additional integrators (or differentiators). This is shown in Fig. 10(b). The additional links are r_1 and r_3, and their correct values relative to r_2 can be found by the following analysis.

In the passive network

$$I_2'/pC_2' = V_2 = V_3 - V_1 \qquad (2)$$

so for the counterpart quantities in the active system we should have

$$I_2'R = (V_3 - V_1)pT_2' \qquad (3)$$

where $T_2' = C_2'R$, R being the scaling factor chosen for converting the currents of the passive network into voltages in the active system.

The voltages shown as available in Fig. 10(b) are

$$I_3R = V_3pT_3 \text{ and } I_1R = V_1pT_1.$$

$I_2'R$ may therefore be obtained by implementing the equation

$$I_2'R = V_3pT_3\left(\frac{T_2'}{T_3}\right) - V_1pT_1\left(\frac{T_2'}{T_1}\right). \qquad (4)$$

Table 1. Factors of constant-k filters with resistance terminations for nominal cutoff frequency of 1 rad/s.

$$\frac{1}{2(1+pT_0)(1+pT_1/q_1+p^2T_1^2)(1+pT_2/q_2+p^2T_2^2)\ldots}$$

n	T_0	T_1	q_1	T_2	q_2	T_3	q_3	T_4	q_4
5	1·54	1·14	2·47	1·414	0·707				
7	2	1·10	6·47	1·414	1·414	1·82	0·642		
9	2·41	1·063	13·3	1·29	3	1·709	1·13	2·19	0·613

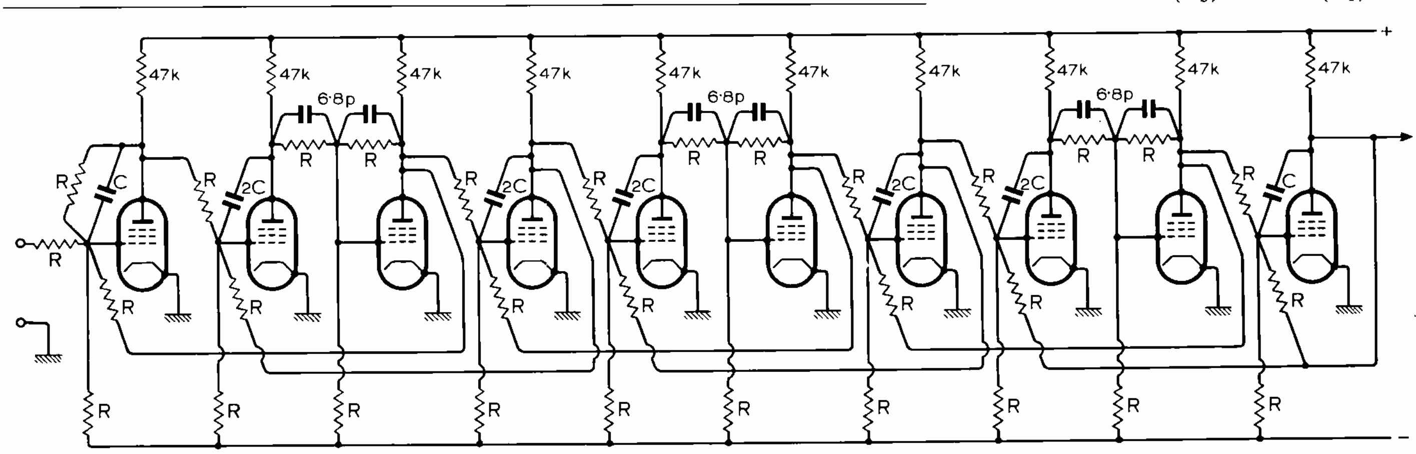
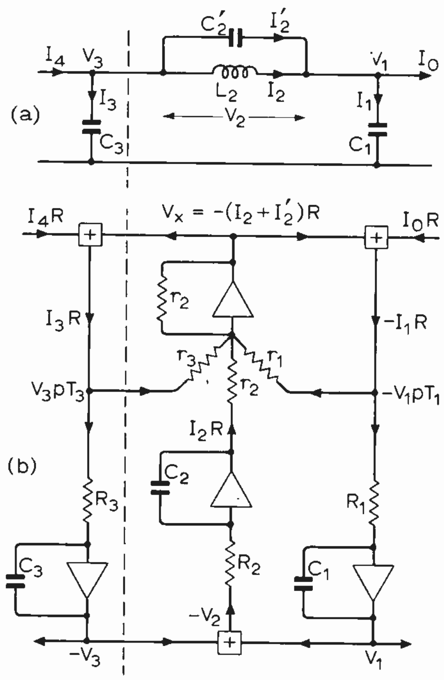


Fig. 8. Very early version of 7th-order low-pass active ladder filter. R = 390kΩ, C = 1000pF, 2C = 2,000pF, $V_1 - V_{10}$ = CV138.

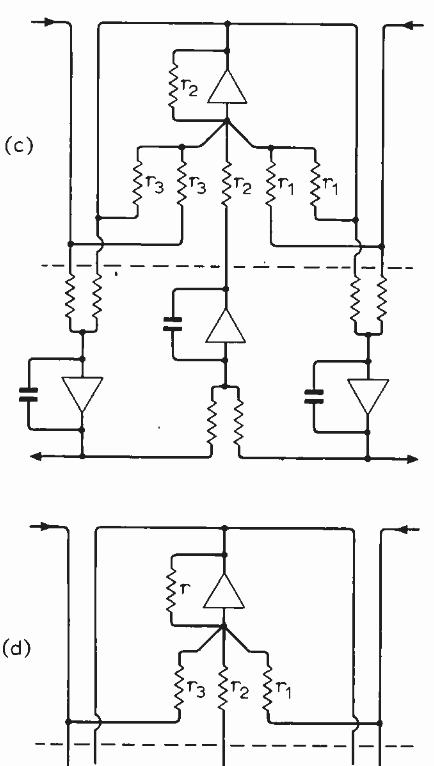
Now for the system as shown in Fig. 10(b)

$$V_x = I_2R\left(\frac{r_2}{r_2}\right)+V_3pT_3\left(\frac{r_2}{r_3}\right) - V_1pT_1\left(\frac{r_2}{r_1}\right). \qquad (5)$$



	Passive	Active
T_1	C_1R	C_1R_1
T_2	L_2/R	C_2R_2
T_2'	$C_2'R$	
T_3	C_3R	C_3R_3

$$\frac{r_3}{r_2} = \frac{T_3}{T_2'} \qquad \frac{r_1}{r_2} = \frac{T_1}{T_2'}$$



$$\frac{1}{r} = \frac{1}{r_1}+\frac{1}{r_2}+\frac{1}{r_3}$$

Fig. 10. A method of introducing a zero of transmission by adding only resistors.

So $V_x = I_2R+I_2'R$ as required if

$$\frac{r_3}{r_2} = \frac{T_3}{T_2'} \text{ and } \frac{r_1}{r_2} = \frac{T_1}{T_2'}. \qquad (6), (7)$$

In practice the synthesis must be modified somewhat, since I_3R, the quantity representing $I_4+(-I_2)$ (we suppose for a moment that C_2' is not present) does not appear as a single voltage, the summation being done by giving the T_3 integrator two input resistors. This situation is met by providing two input resistors r_3 for the inverting amplifier, as shown in Fig. 10(c), and the same is done for $-I_1R$ as shown on the right-hand side of the diagram. The result at this stage is rather cumbersome. But two of the added resistors are in parallel with the feedback resistor r_2, so the system may be simplified as shown in Fig. 10(d), the resistor r being made equal to the parallel combination of r_1, r_2 and r_3.

Fig. 11 illustrates how the method can be applied to reproduce a 5th-order filter with two added stop-band zeros. The parameters have been chosen to realise a Darlington characteristic of the same shape as that shown in Fig. 1 of Part 1, and used as an example in Part 9, but scaled to a lower frequency to set the first zero at 50 Hz. The required time constants, in milliseconds, are given in the table.

Variable tuning

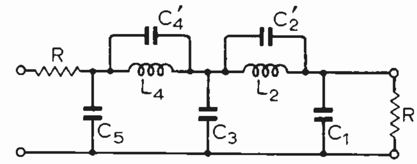
The filter shown in Fig. 11 also serves to illustrate how variable tuning may be applied. The requirement is to vary the *T*s of all five integrators in unison. (T_2' and T_4' follow automatically.) Methods for tuning an integrator were described in Part 8 (Figs 10 and 11). In particular, it was shown (Fig. 10(d) of Part 8) that by "plotting down" the input to an integrator by a factor k (<1) the T of the integrator is increased to T/k. When several inputs are summed at the virtual earth of an integrator amplifier it is obviously necessary to vary all these inputs. However, if thought is given to the placing of the potentiometers, each one may generally control an input to two integrators. This is achieved in Fig. 11 by breaking the circuit at the places marked X and feeding the potentiometers from the outputs of the adjacent amplifiers, the onward signal lines being reconnected to the sliders. It will be noticed that there also has to be a potentiometer in the input line; but the total number has been reduced to six, against the total of eleven separate inputs to the integrators.

The problem of providing a multi-gang potentiometer can be avoided by using the electronic switching method shown in Fig. 11 of Part 8. Using this method filters have been built covering a tuning range of 10 to 1; but because the action of "potting down" reduces the zero-frequency loop gain of the two-integrator loops (and hence their quality—see also Parts 5 and 7), some droop towards the corner of the pass-band response may be expected as the cut-off frequency is reduced, unless an adequate surplus of zero-frequency gain is provided.

High-pass

In Part 7 it was shown how a two-integrator loop can be used to produce the basic 2nd-order high-pass response. As a reminder of this, Fig. 12 shows an undamped high-pass section and the corresponding functional diagram. The essential feature is that the output, V_L, is obtained by subtracting the voltage across C from the input V_i. This puts both integrators in the feedback path.

If the same principle is to be followed when a system of integrators is used to copy the working of a high-pass ladder structure, for example that shown in Fig. 13, it follows that the functional diagram must be arranged so that V_o is derived from V_{in} by successively subtracting voltages, V_6, V_4



Parameters for first zero at 50Hz

T_1	4·77	T_2'	0·8215	T_4'	2·447
T_2	5·62	$\frac{T_1}{T_2'}$	5·808	$\frac{T_3}{T_4'}$	3·058
T_3	7·50				
T_4	4·14	$\frac{T_3}{T_2'}$	9·118	$\frac{T_5}{T_4'}$	1·484
T_5	3·63				

($T_1 = C_1R$, $T_2' = C_2'R$, etc in milliseconds)

ω_p = 33·9Hz

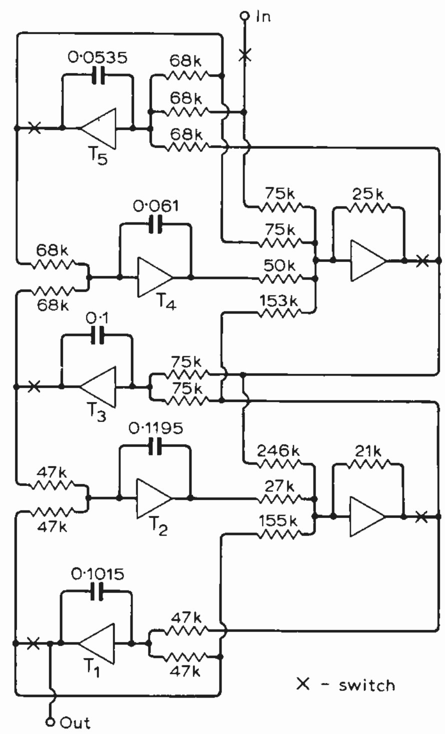


Fig. 11. Realisation of a 5th-order Darlington low-pass response.

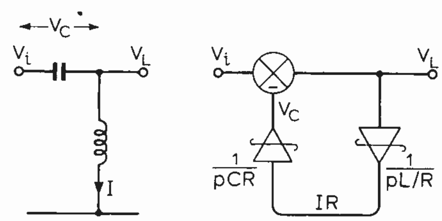


Fig. 12. High-pass sections.

and V_2, corresponding to those across the series-arm impedances of the passive circuit. As usual, it is convenient to work backwards from the output end, and so, assuming V_3 given, it is required to form the relationship

$$V_0 = V_1 = V_3 - V_2. \quad (8)$$

To do this, V_2 must be obtained by copying the working as follows:

$$I_1 p L_1 = V_1 \quad (9)$$

and hence

$$I_1 R = V_1/pT_1 \text{ (where } T_1 = L_1/R). \quad (10)$$

$$I_2 R = I_1 R + I_0 R \text{ (where } I_0 R = V), \quad (11)$$

$$V_2 = \frac{I_2}{pC_2} = \frac{I_2 R}{pT_2} \text{ (where } T_2 = C_2 R). \quad (12)$$

Similarly V_3 is obtained, assuming V_5 is given,

$$V_3 = V_5 - V_4, \quad (13)$$

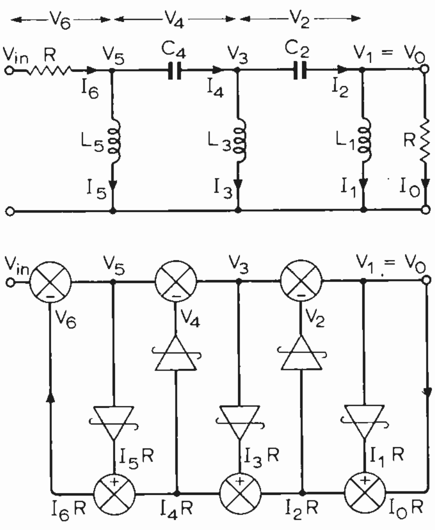
and so on.

The resulting functional diagram is deceptively simple. Turning it into a practical circuit using Blumlein feedback integrators and sign inverting stages leads to a formidable number of interconnections. At the output end, however, these may be reduced to more manageable proportions by using a modified integrator as shown in Fig. 14 which gives the voltage transfer ratio

$$1 + 1/pT_1.$$

In this way both $-V_0$ and $-V_0/pT_1$ are fed back together on a single wire (c.f. Fig. 7, Part 7).

It is also possible without much extra complexity to make additional connections to produce stop-band zeros, and quite high-performance filters have been built by this method. All things considered, however, other methods of building an active high-pass filter, which will be described in a later article, are generally to be preferred.



T_5	T_4	T_3	T_2	T_1
$\frac{L_5}{R}$	C_4R	$\frac{L_3}{R}$	C_2R	$\frac{L_1}{R}$

Fig. 13. High-pass ladders.

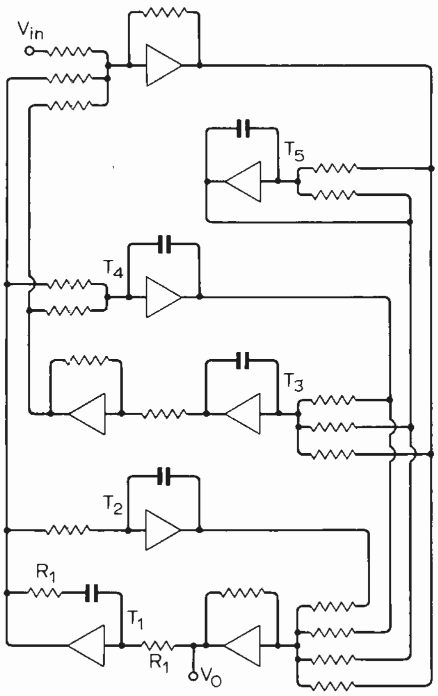


Fig. 14. Electrical version of 5th-order h-p active ladder filter.

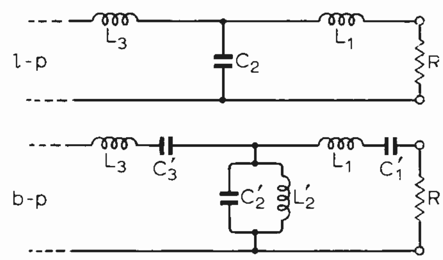
Band-pass

The transformation of a low-pass ladder to band-pass was described in Part 2. Briefly, each reactance in the low-pass ladder is formed into a tuned branch by the addition, either in series for a series arm, or in parallel for a shunt arm, of a reactance of the opposite sort. Each of the tuned branches so formed is tuned to the same frequency, the geometric centre frequency of the required band-pass characteristic. Thus the shape and bandwidth of the band-pass characteristic are determined by the original low-pass filter, and the added reactances simply transfer the characteristic to the required position on the frequency scale. This process is illustrated in Fig. 15, where the output ends of passive ladders of alternative types, T and Π, are shown. The table shows the *T*s which must be equal if the filters are to have the same bandwidth and (for the band-pass filters) centre frequency.

The working of these band-pass filters could be copied step by step, each reactance being considered individually; but the result can be reached more quickly by first noticing two related correspondences. The first, Fig. 16(a), is the familiar one (see Part 7, for example) between a two integrator loop and a series tuned circuit. When the parameters are appropriately matched, the active system in response to an input voltage V produces an output voltage IR, which is proportional to the current I which the same voltage produces in the passive circuit. The second, Fig. 16(b), is the correspondence between a two-integrator loop and a parallel tuned circuit. Here, when the parameters are appropriately matched, the active system produces an output voltage V in response to an input voltage IR which is proportional to the current I which produces the same voltage V across the passive parallel tuned circuit. In both cases R may, in principle, have any arbitrary value.

Thus it is seen that an integrator representing an inductance is transformed to represent a series tuned circuit by connecting across it in a feedback link a second integrator, and that an integrator representing a capacitance is transformed to represent a parallel tuned circuit by the same addition. In each case the two-integrator loops so formed are tuned to the same frequency as the passive tuned circuit

It follows that the equivalent of the passive transformations shown in Fig. 17 is to connect a second integrator across each integrator of the active low-pass ladder to form a two-integrator loop. Each of these loops is tuned to the required centre frequency, and ideally gives infinite gain at this frequency just as the single integrators in a low-pass ladder ideally give infinite gain at zero frequency.



	T_3'	T_3	T_2'	T_2	T_1'	T_1
l-p		$\frac{L_3}{R}$		C_2R		$\frac{L_1}{R}$
b-p	$C_3'R$	"	$\frac{L_2'}{R}$	"	$C_1'R$	"
l-p		C_3R		$\frac{L_2}{R}$		C_1R
b-p	$\frac{L_3'}{R}$	"	$C_2'R$	"	$\frac{L_1'}{R}$	"

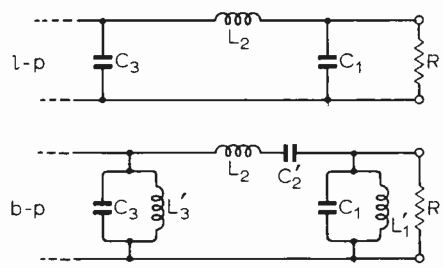


Fig. 15. Derivation of band-pass structures from low-pass.

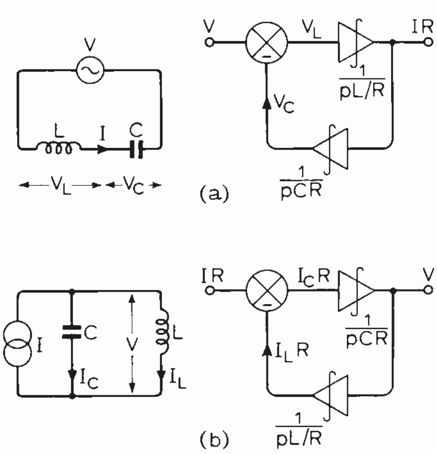


Fig. 16. Active equivalents of the two types of tuned branch.

An example is shown in Fig. 17(b), where for convenience the arbitrary scaling resistance is taken as equal to the terminating resistance of the passive model, Fig. 17(a). The two-integrator loops representing the series-tuned branches develop I_1R from V_1, I_3R from V_3, etc.; and those representing the parallel-tuned branches develop V_2 from I_2R etc. The interconnections via the several difference elements provide that (1) each two-integrator loop which represents a series-tuned branch is fed with the voltage across that branch, i.e. $V_1 = V_2 - V_0$, $V_3 = V_4 - V_2$, etc.; and that (2) each two-integrator loop which represents a parallel-tuned branch is fed with a voltage proportional to the current through that branch, $I_2R = I_3R - I_1R$, etc.

For the sake of completeness Figs 17(c) and (d) illustrate the derivation of an active ladder from the alternative form of passive model; but since, for the same response, T_1 ($= L_1/R$) in the one form must equal T_1 ($= C_1R$) in the other, Figs 17(a) and (d), and so on for all the Ts, the resulting active system, Fig. 17(d), is identical to that of Fig. 17(b).

The functional scheme derived above can be realised by a variety of means. If attention is confined to conventional feedback arrangements of the virtual-earth sort, it is necessary to take account of the sign reversal that accompanies each operation, and also to order the circuits so that subtractions are replaced by summations. A rule that must be followed is that all loops must contain an odd number of amplifiers, one or three. Each two-integrator loop must include a third amplifier for sign inversion, and a straightforward application of these rules gives Fig. 17(e). Clearly a large number of possible variations exist, though normally the circuits will be drawn to give a similar high degree of regularity. At the input end the action of the terminating resistance is easily reproduced as in Fig. 18, and again results in a resistor placed across the capacitor of an integrator.

When the filter modelled is of relatively narrow bandwidth the Ts of the integrators representing the reactances of the original low-pass filter, T_1, T_2, T_3 etc., are much larger than the Ts of the integrators representing the reactances added in the band-pass transformation, T_1', T_2', T_3' etc. (because the low-pass corner frequency, which is also the band-pass width, is much smaller than the band-pass centre frequency). This, as we know from analysis of a single two-integrator loop, does not make good use of the integrator amplifiers. The flexibility of the active system is such, however, that a much larger resistance than R may be chosen to give the two integrators in each loop approximately equal Ts. This increases relatively the voltages at the outputs of the top row of integrators, and the resistances in all the remaining links connected to these points must be increased to compensate, if the bandwidth of the filter is to be unaltered.

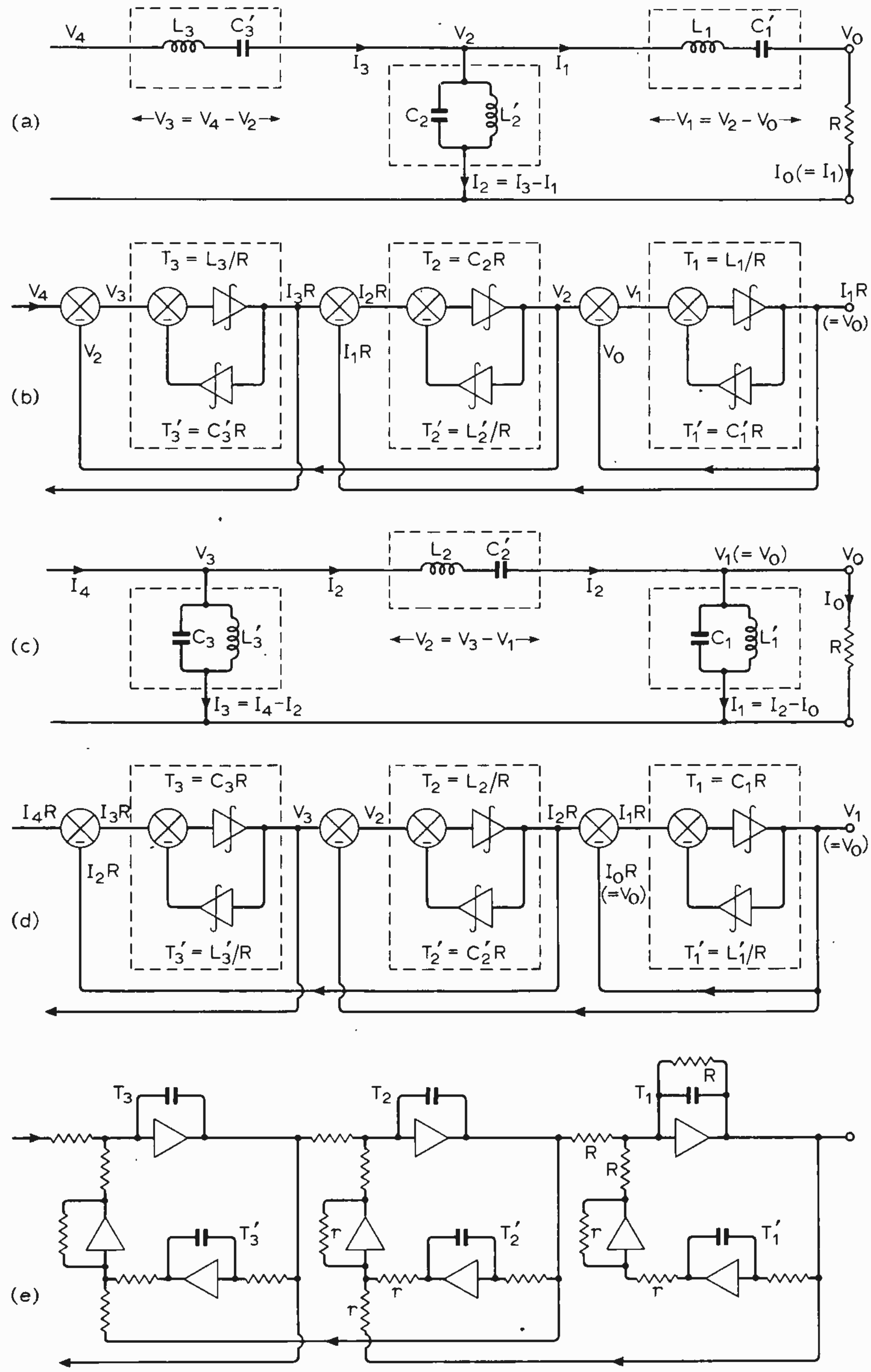
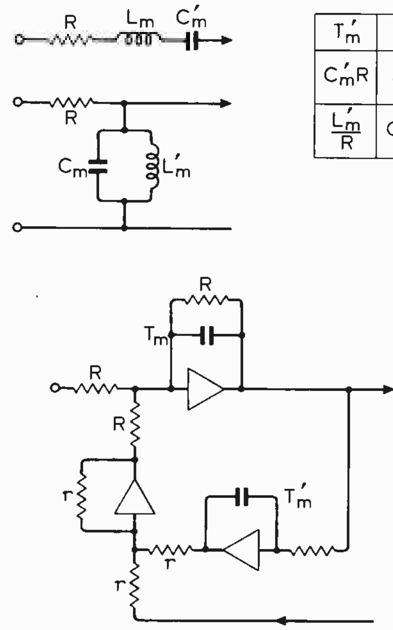


Fig. 17. Derivation of active band-pass ladder.



T_m'	T_m
$C_m'R$	$\frac{L_m}{R}$
$\frac{L_m'}{R}$	C_mR

Fig. 18. Modelling the input end of a band-pass ladder filter.

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1. "Simplified Modern Filter Design" (book) by P. R. Geffe. Iliffe, London, and Rider, New York, 1963.
2. "Tchebyshev Filters and Amplifier Networks", by V. Belevitch. *Wireless Engineer*, April 1952, Vol. 29, No. 343, pp. 106–110.

Improving the 13A Oscilloscope

A few extra parts increase the bandwidth of this popular 'scope to 5MHz

by N. W. Vale

The needs of scientific research and colour TV construction, both carried out on a limited budget, led the author to develop a modification which considerably improves the bandwidth of the Hartley 13A double-beam oscilloscope currently available at low cost on the surplus market. The instrument typically has a 3-dB bandwidth of 10 Hz to 3.5 MHz, although it will display a 4.5-MHz signal of 3V peak-to-peak.

The modified oscilloscope retains a maximum sensitivity of 300mV peak-to-peak, and it has a virtually flat response from 20 Hz to 4.5 MHz. The 3-dB points are at 10 Hz and 5 MHz.

Experiments with *Wireless World* colour TV circuits show that the modified oscilloscope will successfully display and lock video signals at the detector stage, giving a clear indication of the colour burst. It was found that all the important waveforms in the colour decoder could be readily displayed. In other tests the instrument was found to give a useful display at frequencies up to 10 MHz.

All extra components for the modification are easily fitted inside the oscilloscope case and the external appearance and control functions remain unaltered.

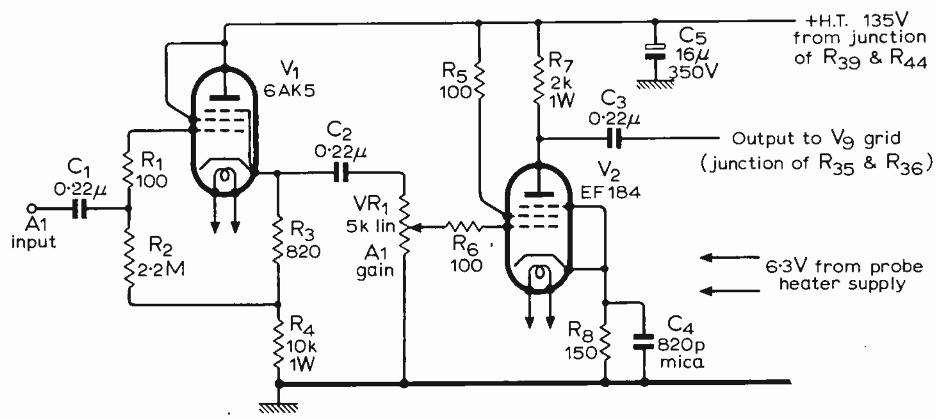


Fig. 1. The circuit of the new wideband amplifier.

General description

The bandwidth of the original instrument can be improved by reducing the value of the feedback resistor (R_{46} in the circuit diagram normally fixed inside the lid of each instrument) in the h.f. deflection amplifier.

Naturally, such a modification gives an appropriate loss of amplification. The deflection amplifier gain may be restored by adding a simple compensated wideband amplifier. A transistor design was originally considered for this application but was rejected for the wider dynamic range and better compatibility of a valve amplifier.

The compensated amplifier finally used is of conventional design and relies on the high slope of a frame grid r.f. pentode to obtain a good sensitivity while a low value of anode load is used to obtain the required bandwidth. Fig. 1 gives the circuit of this amplifier.

The 6AK5 is a triode-connected cathode-follower input stage, to obtain smooth gain control at high frequencies and maintain a high input impedance.

To provide compensation at high frequencies, the cathode resistor of V_2 is bypassed by C_4 the value of which has been selected to give compensation for the h.f. losses which occur in the oscilloscope and its active probe input circuits.

It is suggested that the modified oscilloscope be used with its cathode follower probe plugged in at all times. This is to avoid changes in the h.t. supply voltage to the new amplifier. In any case, it is advisable to use the probe for all work requiring the full frequency response of the modified instrument.

Practical details

Remove the negative feedback resistor R_{46} from the instrument and replace it with a 1.8 kΩ non-inductive carbon resistor rated at 0.5W (R_9).

Interconnect contacts 1 and 2 of the probe switch S_4A in the oscilloscope. This keeps the h.t. supply connected to the probe valve in all switch positions. The object of this is to prevent cathode poisoning of the probe valve, which might occur if the valve was allowed to run with heater only supplies for any length of time, and to keep the h.t. requirements for the input section as steady as possible.

Remove and discard C_{35} (a large paper dielectric type). Fit two suitable valve holders in the region previously occupied by C_{35} and build in the new amplifier. The aluminium chassis bracket which passes the site of the new valves may have an appropriate piece cut out of it to allow easier access.

Remove the existing A1 gain control (R_{38}) and connect the wire which originally fed this control from R_{37} to the chassis. Fit RV_1, the new A1 gain control, using the same panel hole.

Wire a 20-k Ω 10-W resistor (R_{10}) in parallel with resistors R_{39} and R_{40} of the oscilloscope. This compensates for the increased h.t. load due to the added valves.

Remove and discard C_{34}, the original input coupling capacitor to V_9, and connect new amplifier output to V_9.

Connect the input of the new amplifier to contact 6 on S_4B (i.e. to A1 input) and check that the other connections to the new amplifier are as shown in Fig. 1.

Shopping list

R_1	100Ω	R_7	2kΩ*	C_3	0.22μF
R_2	2.2MΩ	R_8	150Ω	C_4	820pF#
R_3	820Ω	R_9	1.8kΩ†	C_5	16μF*b*
R_4	10kΩ*	R_{10}	20kΩ§	VR_1	5kΩ lin.
R_5	100Ω	C_1	0.22μF	V_1	6AK5
R_6	100Ω	C_2	0.22μF	V_2	EF184

*1W rating. †0.5W rating. §10W rating.
All other resistors rated at 0.25W.
#mica. *b*350V working.
All capacitors 250V working polyester unless otherwise specified.

Electronic Building Bricks

4. Transducers—analogue and digital

by James Franklin

In Part 2 we saw how information could be represented by electrical variables, both in static and dynamic form. A "travelling representation" is called a signal. Two examples given were pulses of electrical energy representing number of objects travelling on a conveyer, and a continuous variation of electrical energy representing sound-wave energy. To obtain these electrical signals a converting device was needed in each case—a photo-electric cell to convert the light energy and a microphone to convert the sound energy. There are in fact a great many converting devices in use, particularly in industrial control equipment. A familiar one at home is the gramophone pickup. Some convert the mechanical or other energy directly into electrical energy, as shown in Fig.1(a), while others—for example, the oven thermostat—use it to control electrical energy coming from a separate source, as shown at (b).

In Part 3 we saw that electrical energy is constituted by electrons, and that a general movement of many electrons is an electric current, measurable as a flow rate in coulombs per second (amperes). Current is an electrical variable, and as

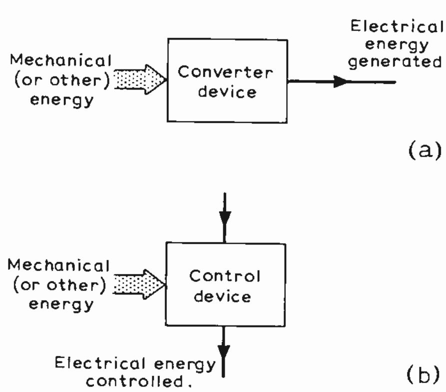


Fig. 1. Obtaining information in electrical form from information represented by some other type of energy: (a) direct energy conversion; (b) control of electrical energy.

such can be controlled. Thus in Fig.1(b) we could replace the words "electrical energy" by "electric current" or indeed by words denoting any other electrical variable. Also in (b) we could replace the words "mechanical (or other) energy" by words describing a particular physical variable, such as force or temperature.

We can, in fact, develop from Fig.1 a much more general concept—a device that receives some physical variable representing information and transforms this into an electrical variable representing the same information. The variables change but the essential information does not. Any actual device which does this is called a *transducer**. This is our first "building brick", and we are showing it, according to plan, just as a functional block—Fig.2.

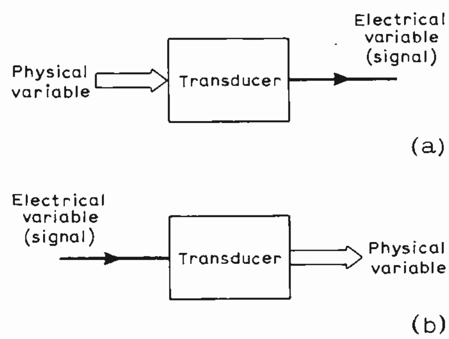


Fig. 2. Electronic building brick No. 1—the transducer; (a) an input transducer; (b) an output transducer.

It must be quickly added that "transducer" can also mean a device that works in the reverse direction, converting an electrical variable into a physical variable, as shown at (b) in Fig.2. This may be thought confusing, but in practice the transducer's direction of operation in an electronic system becomes obvious from the position the device occupies in a block diagram, say at the input as against the output. Domestic examples of the Fig.1(b) type of transducer are the loudspeaker and the television picture tube, though these are not often described as transducers. But the term is used frequently in industrial electronics (Fig. 2 (a) could be a "pressure transducer") and sometimes in sound reproduction.

From Part 2 we saw that information may be represented in a signal in two ways. In the electronic counting system the number of objects was represented by that *number* of pulses of electrical energy. In fact the exact form of the energy-time

* From Latin *trans* (across), *ducere* (to draw).

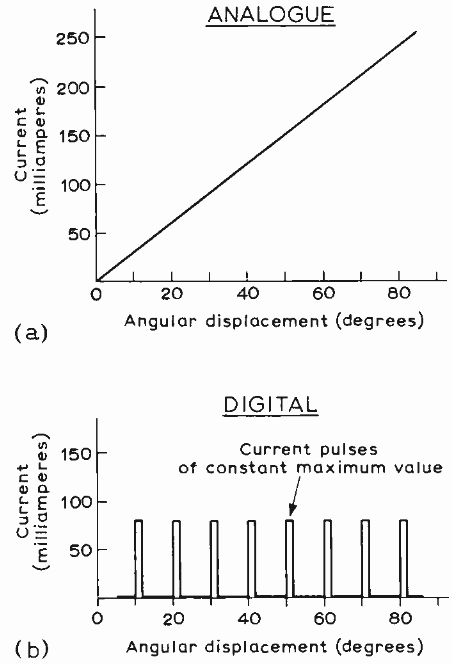


Fig. 3. Graphs illustrating the action of (a) an analogue transducer and (b) a digital transducer, both responding to angular displacement of a shaft.

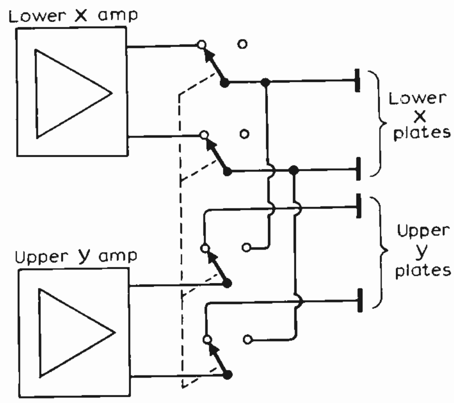
graph did not matter very much: the pulses could equally well be triangular or some other shape provided their number was correct. This type of representation, in which the number of electrical events gives the essential information, is called a *digital*† signal.

With the other type of representation, the continuously varying electrical energy obtained from a microphone responding to a violin note, the successive values of electrical energy were proportional to the successive values of sound energy —they had to be, otherwise distortion of the information would occur. In other words the electrical energy time-graph was similar in form to the variation with time of the sound energy (supposing this were plotted as a graph). As such the electrical variable was a sort of model, or analogue, of the sound variable. Such a representation is called an *analogue*‡ signal.

Some transducers are designed to work with analogue signals, others with digital signals—it depends on what is needed for a particular electronic system. For example, for a system in which a digital computer is the main element, digital transducers might well be the more convenient. To illustrate the difference between the modes of operation, Fig.3 shows the relationships in two transducers, both representing a given physical variable, rotation of a shaft: (a) an analogue transducer giving a proportional electric current (3 milliamperes per degree of rotation), and (b) a digital transducer giving a related number of pulses of current (one pulse per 10 degrees). Both transducers use current as the electrical variable, the analogue type directly, the digital type merely as a medium for denoting number.

† From Latin *digitus* (finger). The link with counting is obvious.
‡ From Greek *ana* (up to), *logos* (proportion).

Circuit Ideas



Switching for X Y plotter.

Hybrid push-pull deflection amplifier

A requirement arose for a push-pull deflection amplifier for an electrostatic cathode ray tube system where it was not practicable to provide the additional negative h.t. and heater supplies which would be required by the usual constant-current pentode coupling circuit. To satisfy the requirement a hybrid circuit was devised which uses a double triode valve in a standard long tailed pair configuration and a two terminal constant current circuit using transistors[1,2] in the common cathode lead. Assignment of component values is very simple. First anode resistors and anode currents are chosen to meet the output requirements and this fixes the standing grid-to-cathode bias and the necessary input voltage swing. It is then possible to determine the offset voltage for the grids relative to ground from the inequality:—$V_{gg} + V_{gk\ (min)} \geqslant V_{cc\ (min)}$

where V_{gg} is the required offset voltage

V_{gk} is the grid-to-cathode voltage

$V_{cc\ (min)}$ is the minimum voltage which can appear across the constant current circuit for correct operation

provided that a value for $V_{cc\ (min)}$ is fixed. $V_{cc\ (min)}$ is determined by the zener diodes, the use of those with the lowest available reference voltage setting a lower limit for $V_{cc\ (min)}$ of about 7V. If a smallest value for $V_{gk\ (min)}$ of 2V is assumed, a minimum value of 5V is established for V_{gg}. Use may be made of this offset voltage to provide a trace shift or picture centring facility as shown in the circuit diagram. The components shown provide a 460V peak-to-peak output swing for a 28V peak-to-peak input swing.

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Hants.

1. P. Williams, "Letters to the Editor", *Wireless World*, p.456, Sept. 1966.
2. P. Williams, "Ring-of-two Reference", *Wireless World*, p.318, July 1967.

Conversion of a double-beam 'scope for XY plotting

For those unable to buy an oscilloscope having identical X and Y amplifiers, the ability to do XY plotting is somewhat limited. The figure given illustrates a simple modification to a double-beam oscilloscope. The idle Y amplifier is made use of, and switched to the X plates, which in turn have merely been disconnected from their own amplifier. Leaving the spare beam disconnected at the upper Y plates usually means that it will appear near centre on the screen, but with the brilliance turned down, no difficulty should ensue. Where the output of the normal X and Y amplifiers is developed with respect to chassis, care should be taken to avoid reversals in the switching.

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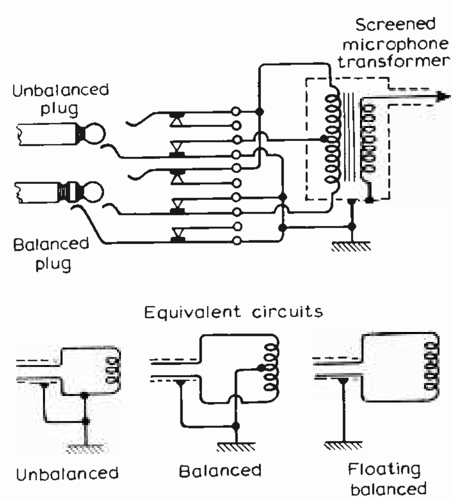
Bletchley,
Bucks.

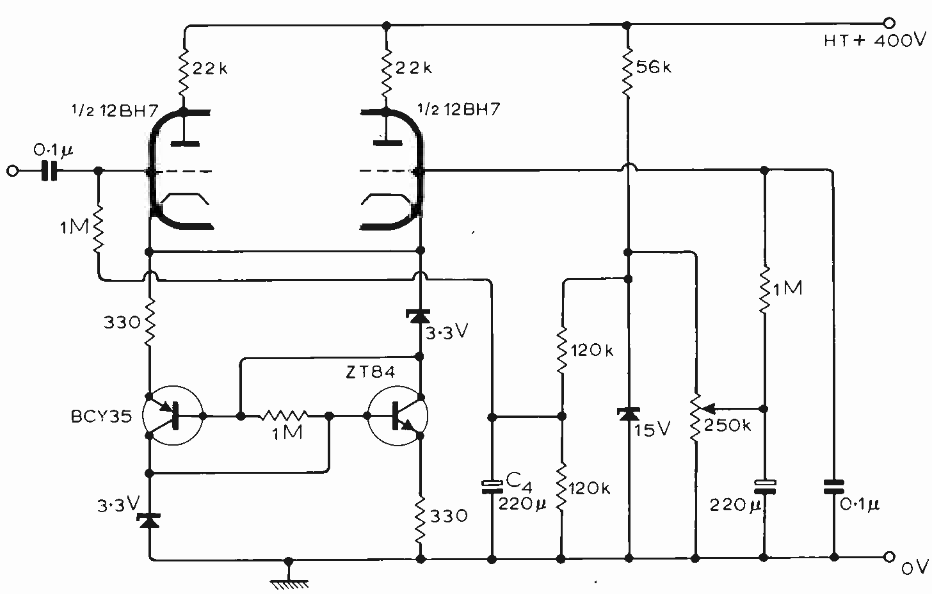
Connecting microphones to transformers

It is sometimes necessary to be able to connect either a balanced or unbalanced input plug to the primary of a microphone transformer. This can be done with the circuit shown which allows either to be plugged into its appropriate socket. The jack sockets have normally closed contacts as shown, and the connections are arranged so that the whole of the primary winding is included and the appropriate points and cable screening are earthed in each case. A floating balanced input can also be obtained by inserting an empty plug into the unbalanced socket.



Versatile microphone transformer arrangement.

A. C. GOTT,
Southall,
Middx.



Transistors supply constant current to valve amplifier cathodes.

The F.E.T. as a Class-A Audio Amplifier

When to use and how to bias

*by P. L. Matthews**

In spite of the publicity given to field-effect transistors when they were introduced, they are still dominated by bipolar types in audio circuits. One of the main reasons for this is that present methods of manufacture lead to wide variations in their primary characteristics—the zero gate voltage drain current I_{dss}, pinch-off voltage V_p, and mutual conductance $_{gmo}$—which are approximately related by $2I_{dss} = g_{mo}.V_p$.

Probably the most widely used f.e.t. is the 2N3819, with V_p and I_{dss} spreads of 0·5 to 7·5 V and 2 to 20 mA respectively. Without going into calculations it can be seen that if devices with extreme parameter values are to be incorporated into a conventional self-biased stage, the design will have to tolerate a drain current variation of the order of 10 to 1, which is obviously impracticable. The extra cost of a tightly specified selection of the 2N3819 makes it less of an economic proposition when a couple of cheap plastic bipolars will often do the job just as well. However, there are some instances in which an f.e.t. must be used, and this article aims to show which devices should be chosen for a given application, and how they should be biased.

The perfect performance

In order to determine how bias variations alter the performance of an f.e.t. amplifier it is useful to consider the operation of a theoretically perfect device in an isolated self-biased stage. The circuit of Fig. 1 depicts an n-channel f.e.t. operated from a supply voltage of $+V_{dd}$, with load and bias resistors R_L and R_S respectively, when capacitor C is sufficiently large to bypass signal frequencies. The input shunting resistor R_g, which would typically be about 1 MΩ, must nevertheless be small enough to ensure that $R_g.I_{gss} \ll V_{gs}$, and the gate terminal can then be considered to be at zero d.c. potential. In addition the primary characteristics of the f.e.t. are taken as V_p, I_{dss} and g_m.

*Newmarket Transistors Ltd.

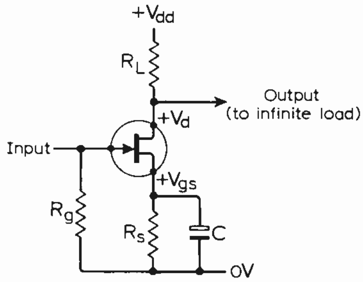


Fig. 1. Standard biasing circuit for f.e.t. audio stage.

a \ b	1/4	1/2	3/4	1	2	3	4
1/4	0·25	0·38	0·48	0·54			
1/2	0·32	0·50	0·65	0·76	1·07		
3/4	0·33	0·57	0·75	0·90	1·33	1·60	
1	0·35	0·61	0·82	1·00	1·53	1·89	2·34
2		0·70	0·98	1·22	2·00	2·59	3·06
3			1·05	1·33	2·26	3·00	3·61
4				1·41	2·44	3·28	4·00

$$C = \frac{b}{2a}\left[(4a+b) - \sqrt{b(8a+b)}\right]$$

Fig. 2. Values of c for various values of a and b, according to the equation.

From the above it can be seen that, provided the input signal remains within the linear region, the maximum pk–pk output voltage swing in the linear region ($V_{ds} > V_p$) is

$$V_{dd} - [V_{ds(sat)} + V_{gs}]$$
$$= V_{dd} - [(V_p - V_{gs}) + V_{gs}]$$
$$= V_{dd} - V_p.$$

The voltage drop across R_L is half the swing, or $\frac{V_{dd}-V_p}{2}$, and this gives the optimum value of R_L as

$$R_L = \frac{V_{dd} - V_p}{2I_d}$$

But $A_V = g_m \times R_L$

$$= \frac{2I_{dss}}{V_p}\left(1 - \frac{V_{gs}}{V_p}\right) \times \frac{V_{dd} - V_p}{2I_d}$$

$$= \frac{V_{dd} - V_p}{V_p - V_{gs}} \text{ (with substitution for } I_d\text{).}$$

Again, considering pk–pk signal voltage,

$$V_{in(max)} = \frac{V_{out(max)}}{A_V}$$

This gives $V_{in(max)} = V_p - V_{gs}$.

This result shows that no matter what other specifications the f.e.t. has, provided that the load resistance is substantially lower than the input impedance of the following stage, and that the drain voltage falls exactly at the centre of the output swing, an input signal of $V_p - V_{gs}$ (pk–pk) gives the maximum output of $V_{dd} - V_p$ (pk–pk). Consequently, a device with a low pinch-off voltage should be selected for use with a low supply voltage so that a larger output swing will be available. As a general rule, it is not worthwhile operating an f.e.t. from a supply voltage less than twice its maximum pinch-off voltage, so the 2N3819 would have to run off at least 15V. You have to be careful, of course, not to exceed the maximum voltage rating for the device —in this case it is 18 V.

It is now easy to calculate the harmonic distortion (predominantly second) produced by the above amplifier at maximum output.

$$D_2 = \frac{V_{in}}{8(V_p - V_{gs})} = \tfrac{1}{8} = 12\tfrac{1}{2}\%.$$

It may seem remarkable that the above simple results should hold for any depletion mode f.e.t. but in practice a slight departure from the theoretical value would be expected due to variations in the specification of V_p, and to the fact that the assumed equations may only be approximately true.

The problem of parameter spreads

The selection of suitable bias conditions for the f.e.t. is now considered. With a particular device, it can be seen that as V_{gs} is increased toward V_p, more voltage gain is obtained at the expense of output swing, and vice versa. A higher supply voltage increases both gain and output, but, as the designer is often tied to the supply of 12 V or so available in most equipment and because the drain–source breakdown voltage (which is only about 20 V for many popular devices) imposes limitations, this is not always a practicable proposition. Nevertheless, a quick calculation will show that quite a reasonable gain and output can be obtained with a typical f.e.t. but the problem of parameter spreads now becomes clear. If a circuit is set up with components selected to suit a device with a chosen V_p and I_{dss} substitution of a dissimilar device is likely to cause severe bias problems. In particular, the variation in drain current will upset the vital centring of the drain voltage between V_{dd} and V_p, causing a considerable reduction in output voltage swing. Unfortunately this problem cannot be solved in general as simply as in the preceding calculations, and the best technique is to determine the maximum spreads of I_{dss} specified for the device in question, and then to calculate the optimum resistance values for a typical centre spread unit. As an example, in the case of the 2N3819, it would be best to take I_{dss} = 11 mA and V_p = 4 V, as these are the middle values. It also proves useful to specify $V_{gs} = \frac{V_p}{2}$ in most cases, as this places the input voltage swing safely in the centre of the permitted region.

Suppose, in the circuit of Fig. 1 that a device with parameters I_{dss} and V_p are inserted, R_S is adjusted to make $V_{gs} = \frac{V_p}{2}$, and R_L sets the drain voltage to the middle of the output swing. The drain current is therefore $\frac{I_{dss}}{4}$. If now the f.e.t. is removed and replaced by a device with parameters $I_{dss'}$ and $V_{p'}$, where $I_{dss'} = (a.I_{dss})$ and $V_{p'} = (b.V_p)$ (a and b are small positive numbers) then the new drain current $I_{d'}$ can be related to the original by $I_{d'} = c.I_d$.

Substituting into the standard transfer equation,

$$I_{d'} = I_{dss'}\left(\frac{V_{p'} - V_{gs'}}{V_{p'}}\right)^2$$

giving

$$cI_d = a\,I_{dss}\left(1 - \frac{c}{2b}\right)^2$$

and thus

$$c^2\left(\frac{a}{b^2}\right) - c\left(\frac{4a}{b} + 1\right) + 4a = 0.$$

The solution of this quadratic is

$$c = \frac{b^2}{2a}\left(\frac{4a+b}{b} - \frac{\sqrt{8a+b}}{b}\right),$$

and further simplification gives

$$c = \frac{b}{2a}\left[(4a+b) - \sqrt{b(8a+b)}\right].$$

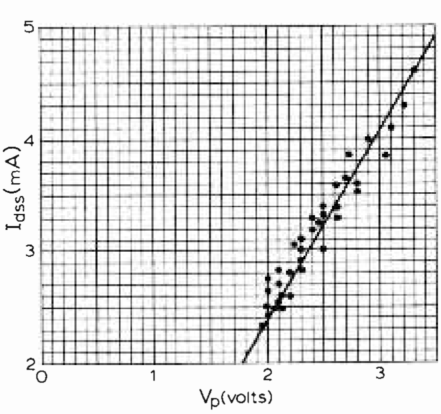


Fig. 3. Graph showing proportionality between V_p and I_{dss} for NKT80110 family—typical for most popular devices.

Approximate values of c for various values of a and b are given in Fig. 2.

This discussion assumes to a certain extent that the highest and lowest values of V_p and I_{dss} occur simultaneously, so that they are considered to be characteristic of an extreme device. In theory this is not necessarily the case, but the graph in Fig. 3 illustrates that V_p can be considered to be roughly proportional to I_{dss} for devices from the Newmarket NKT80110 family, and this can be shown empirically for most popular devices. In any event it is possible to allow for discrepancies of this nature merely by substitution of appropriate "worst cases" in the above equation.

A practical example

An example will help to clarify the general procedure. Consider the circuit shown in Fig. 4(a) which incorporates a junction f.e.t. type NKT80212. The typical V_p and I_{dss} for this device are 700 mV and 200 μA respectively, and setting $V_{gs} = \frac{V_p}{2}$ = 350 mV gives a drain current of 50 μA. The values of R_S and R_L can be calculated from this information to be 7 kΩ and 83 kΩ respectively and the nearest E12 preferred types have been specified in the diagram.

Consider now what happens when this typical device is replaced by a "lower spread" unit with V_p and I_{dss} of 500 mV and 100 μA respectively as shown in Fig. 4(b). For this unit, $a = \frac{1}{2}$, $b = \frac{5}{7}$ and this gives c = 0·6 (approximately) so the new drain current will be about 30 μA. The source and drain voltage can now easily be calculated and it is seen that the drain voltage is above the output swing centre, causing clipping of the waveform unless its amplitude is reduced to 2·5 V pk–pk. A similar argument applies to the "upper spread" unit with $a = 1\frac{1}{2}$, $b = 1\frac{2}{7}$ and c = 1·35 (approximately), giving I' = 68 μA, but this time the drain voltage is too low and restricts the output to 2·5 V pk–pk. Thus the circuit will be able to yield an output of only 2·5 V pk–pk at most for any NKT80212 f.e.t. without individual circuit adjustment, as distinct from just over 8 V pk–pk for the typical device considered first.

The distortion level appears to be rather complicated to calculate, but it is sufficiently accurate to consider the ratio of available output to maximum output and multiply this by the previously determined value of $12\frac{1}{2}\%$. In this case the value would be

$$D_2 = \frac{2{\cdot}5}{8} \times 12\tfrac{1}{2}\% = 4\% \quad \text{(approximately).}$$

The justification for this calculation is understandable in the case of the typical device, as D_2 is proportional to V_{in}, but this can be formally extended to cover the limit cases if desired.

The effect of temperature

Variations in temperature, supply voltage, resistor values, etc., can readily be incorporated into the above argument, but, as the first of these can be very important, a brief outline of the method is given. The pinch-off voltage of an f.e.t. decreases with

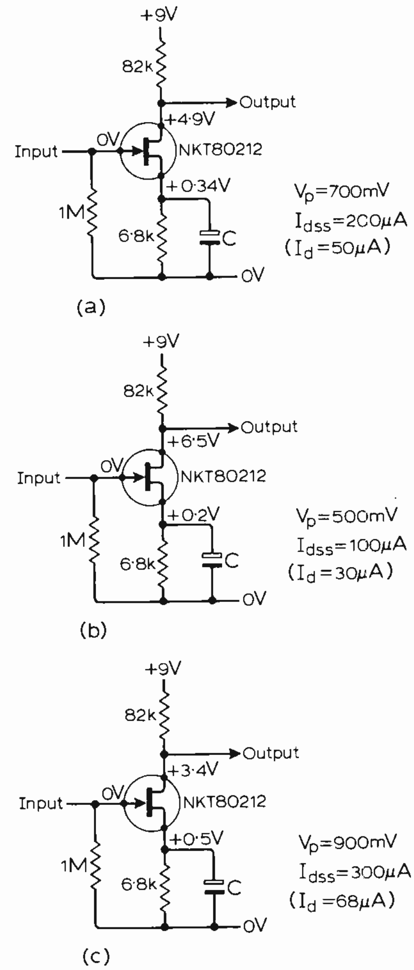


Fig. 4. Circuits illustrating the effect of parameter spreads in f.e.t.s.

temperature by a factor similar to that of a p–n junction, a typical value for diffused devices being −2 mV/C°. Unfortunately it is not as easy to define a simple factor for I_{dss} as this depends on such things as carrier mobility and impurity concentrations of the particular device, but many manufacturers provide a graph of the relationship. As a guide, however, it can be estimated that I_{dss} will fall linearly with increase in temperature up to 75°C, when its value is $\frac{3}{4}$ of that at 25°C. Conversely, the I_{dss} at −25°C is $1\frac{1}{4}$ times the value at 25°C. If the previously designed amplifier was required to operate from −25° to +75°C it would be necessary to consider the typical device at 25°C as before, the lower spread device at +75°C and the upper at −25°C. These would give a $V_{p'}$ and $I_{dss'}$ of 400 mV, 75 μA and 1 V, 375 μA for them, respectively, so a, b, and subsequently c would be altered accordingly.

Conclusion

"You pays your money, you takes your choice". Either you select an economical "full spread" device and accept its restricted usefulness, or pay more for a special selection with its corresponding advantages. It's not surprising that bipolar transistors are still the first choice for most audio circuits.

Personalities

Paul Adorian, F.C.G.I., F.I.E.E., F.I.E.R.E., managing director of Rediffusion Ltd since 1966, is to retire at the end of October when he will be succeeded by **Hugh Dundas,** D.S.O., D.F.C., D.L., who has been on the board of the company since 1966. Mr. Adorian, who will be 65 in November, entered the field of radio communications as a student engineer with Standard Telephones and Cables. He joined Rediffusion in 1932 as a development engineer, becoming in succession assistant chief engineer and chief engineer before joining the board. He is also chairman of Redifon Ltd, manufacturers of flight simulators and communications equipment, and of Redifon-Astrodata Ltd and a director of the British Electric Traction Company Ltd. Mr. Dundas, his successor, joined Rediffusion's executive staff in 1961 from the Beaverbrook Press. He is also a director of Rediffusion's parent company, British Electric Traction, and of Thames Television.

Sinclair Radionics has bought a major shareholding in the AIM Associates Group. **Tim Eiloart,** founder of the group, has resigned from the position of chief executive but remains outside chairman of Cambridge Consultants Ltd, the research and development company of the AIM group, and **Richard Cutting** is appointed managing director. **Gordon Edge** and **Roy Hawkins,** who were joint managing directors of Cambridge Consultants, have left the company and been appointed to the Technical Division of PA Management Consultants Ltd. **David Southward,** another of the founder-directors of the company, remains managing director of AIM Electronics, AIM bioSciences and Cambridge Audio Laboratories, the Group's manufacturing subsidiaries. **Clive Sinclair,** managing director of Sinclair Radionics, which he founded eight years ago, has become chairman and chief executive of the AIM group. Mr. Sinclair has stated that the two organizations will operate independently "because of their very different fields of activity".

Commander David W. Malim has become chairman of Marconi Space and Defence Systems Ltd, one of the four systems companies within GEC-Marconi Electronics Ltd. Educated at the Royal Naval College, Keyham, he served as an officer in the Royal Navy throughout the war and was on the staff of the British Joint Services Mission in Washington from 1946 to 1949 when he went to the Admiralty as weapon engineer officer. On his retirement in 1959 Commander Malim became manager and director of Lancashire Dynamo Ltd, in Manchester, joining Elliott Automation Ltd in 1962 and becoming joint managing director of Elliott Space and Weapon Automation Ltd in 1963. English Electric bought Elliott Automation in 1968 and after the merger with GEC in 1969 all the electronics interests in the group were reorganized. Elliott Space and Weapon Automation became part of GEC-Marconi Electronics Ltd and Commander Malim was appointed joint managing director of GEC-Elliott Space and Weapon Systems with **Arthur S. Walsh,** M.A. (Cantab.), who now becomes managing director of Marconi Space and Defence Systems. Mr. Walsh, who is 44, was educated at Selwyn College, Cambridge, where, after service in the Royal Signals from 1944 to 1948, he graduated in natural sciences in 1951. He joined the GEC Laboratories at Stanmore in the radar microwave group and became a group leader in 1956. Mr. Walsh was appointed assistant manager of the Applied Electronics Laboratories at Stanmore in 1964, becoming manager a year later, and was technical director of GEC-AEI (Electronics) Ltd from 1966 until his appointment as joint managing director of GEC-Elliott Space and Weapon Systems Ltd in 1969.

Leo G. Dive, the senior assistant in the B.B.C. Engineering Information Department, is to be the Corporation's engineering representative on the B.R.E.M.A. commercial committee in succession to **Hugh Greatorex,** who retired recently. Mr. Dive, who was from 1964-1966 senior engineer in the B.B.C's New York Office, has also taken over responsibility for the management and direction of the B.B.C's mobile colour demonstration unit which is at present on a summer tour of exhibitions and promotions. Enquiries about the unit should be addressed to Mr. Dive at the Engineering Information Department, B.B.C., Broadcasting House, London W1A 1AA.

Eric J. Wightman has been appointed engineering director of the Industrial Instrument Division of Smiths Industries Ltd. Mr. Wightman, who is 43, has spent a little over three years with the Industrial Instrument Division, first as chief engineer and latterly as technical manager. Before joining S.I. he spent two years as chief engineer, data systems, at Solartron and previously was with the Gyroscope Division of Sperry Rand for seven years.

K. Milne, Ph.D., M.I.E.E., has been appointed engineering manager of the Radar Equipment Division of Plessey Radar Ltd in succession to **R. L. Burr,** who has become technical co-ordinator for Plessey Radar. Dr. Milne began his career in 1946 as chief microwave aerial designer with Associated Electrical Industries Ltd. He was awarded his doctorate in 1951 for his thesis on wide-angle scanning properties of microwave lens aerials. In 1953 he became engineeer-in-charge of systems analyses of radar systems, including c.w. and volumetric scanning radar. He joined Decca Radar Ltd in 1960. Particular projects with which he was concerned included satellite earth terminals and electronic scanning radars. He was appointed manager of Plessey Radar's Space Systems Department in 1966.

J. W. V. Denton is appointed national sales manager by Data Recognition Ltd, of Reading. Prior to joining D.R. he was manager, information systems, for Motorola Control Systems Ltd and before that was with Cossor Communications for four years as their regional sales manager. Mr. Denton received his training in communications and radar in the Royal Navy.

Sir Robert Cockburn, K.B.E., C.B., has been appointed chairman of the Council of the National Computing Centre. He succeeds **J. M. A. Smith,** who recently retired. Sir Robert, who is 61, was director of the Royal Aircraft Establishment, Farnborough, from 1964-1969. A graduate of London University, he engaged in research in communications, radar and atomic energy between 1937 and 1948 at R.A.E., the Telecommunications Research Establishment and the Atomic Energy Research Establishment, respectively. He was the Scientific Adviser to the Air Ministry from 1948 to 1953 and between 1954 and 1959 was successively principal director of scientific research (guided weapons and electronics), deputy controller of electronics, and controller of guided weapons and electronics at the Ministry of Supply. Sir Robert, who received his knighthood in 1960, was chief scientist, Ministry of Aviation, from 1959 to 1964, and vice-chairman of the Space Committee, Ministry of Defence, from 1964-1965.

Transitron Electronic Corporation, of Wakefield, Mass., has announced the appointment of **Grahame F. Hazell** as vice-president for European Semiconductor Operations. Mr. Hazell, who is 36, is a native of Ipswich, Suffolk, and a graduate in physics of the University of Nottingham. For the past 10 years he has been employed by Texas Instruments, initially on various marketing assignments and for the past few years as operations manager of the TI facilities in Bedford and, more recently, Plymouth. The Transitron companies reporting to European Semiconductor Operations are in the UK, France, West Germany, Holland and Sweden.

George Siddall, who joined the Electrical Research Association as head of its Electronics Department in 1962, has been appointed assistant director technical to help Dr. B. C. Lindley, the director, in the formulation, co-ordination and control of the Association's technical policies and operational activities. He will have special responsibility for the Applied Sciences Division and for the technological planning unit.

Robert W. Beattie has resigned from the managing directorship of Electrosil Ltd and chairmanship of Miniature Electronic Components Ltd. He is succeeded in these positions by **John E. Carl,** B.Sc., a 42-year-old American who recently came to this country from the parent company Corning Glass Works. After graduating from Alfred University, New York, in 1951 with a degree in ceramic engineering he joined the research and development department of Corning. In 1968 he moved to the electronic components plant at Bradford, Pennsylvania, where he remained as plant manager until coming to the U.K. recently as production executive at Pallion, Sunderland.

OBITUARY

John Goodman, assistant managing director of the Dubilier Condenser Co. (1925) Ltd, died recently at the age of 56. Mr. Goodman joined the company in 1932, and was appointed assistant managing director in 1966. His father was one of the founders of the company in 1912.

World of Amateur Radio

Awards and certificates

Although amateur operating proficiency awards and certificates have a long and respected history—for example, the "worked all continents" award dates back to the early days of the International Amateur Radio Union, founded in 1925—there are fears that the situation has rather got out of hand in recent years. As far as can be judged there are now between 800 and 1000 different awards issued by national and international societies and groups, local clubs, and in association with various amateur radio publications. While many of these do undoubtedly encourage useful activities and provide amateurs with valuable competitive yardsticks, there is a growing belief that unless care is exercised the rising flood of awards may bring the system into disrepute. The recent introduction of new awards, such as the "five-band DXCC" requiring some 500 QSL cards, valid only if of recent date, is threatening to overwhelm the amateurs in the "rarer" countries with requests for more and more cards. For several years some of these amateurs have appointed "QSL managers" to undertake the tremendous task of verifying thousands of brief radio contacts each year. But even so (and there are arguments against the system of QSL managers as this is open to abuse), it is becoming more and more difficult to obtain the rarer cards from overseas stations swamped by requests.

The R.S.G.B.—which has done much to uphold the continued value of proficiency awards—has recently announced new rules for all of its awards (details from R.S.G.B., 35 Doughty Street, London W.C.1), but one notes with some alarm that yet another major new award "The I.A.R.U. Region I Award" has been added to the list—this time at the request of the I.A.R.U. Region I Bureau.

R.A.E. courses

Would-be amateur transmitters should note that a number of evening courses covering the syllabus of the Radio Amateurs Examination will be starting at local educational centres in many parts of the country during September. Typical of these is one at Birkenhead Technical College where enrolment takes place from September 7 to 10, and classes are held on Thursday evenings. In this area, as in a number of others, Morse classes are held in conjunction with the local amateur radio society, in this case the Wirral A.R.S. (non-members enquiries to Alf Fisher, G3WSD, 34 Glenmore Road, Oxton, Birkenhead. We have been notified of courses at several London centres including Acton Technical College, Wembley Evening Institute and Gascoigne Recreation Centre, Barking. Readers in other parts of the country should make inquiries of their local education authority to discover if there are any local courses.

Transceivers

Although Collins introduced an h.f. transceiver about 10 years ago, it is only quite recently that fairly substantial numbers of amateurs have swung over to the use of compact, combined transmitter-receivers.

While transceivers are generally regarded as fairly ambitious projects for home construction, many of the problems can be overcome by group construction. For example, the Nottingham Amateur Radio Club is planning a constructional project this winter under the guidance of Bob Sills, G31QM, for a number of members to build their own five-band, largely solid-state, s.s.b. transceivers at a cost in the region of £30.

An interesting recent development in the field is the marketing in the United States of a compact 2-watt c.w.-only low-power transceiver for 3.5 and 7 MHz, for use as a fixed or vacation station. The key feature of this all-semiconductor equipment, which runs from dry batteries and is made by Ten Tec for sale at about $55 (£23), is the use of its single variable-frequency oscillator to form the transmitter driver and a simple homodyne (synchrodyne) receiver based on a dual-gate m.o.s.f.e.t. heterodyne detector, and with a single integrated circuit providing all the a.f. amplification.

At the other end of the scale is the recent 300-watt p.e.p. CX7 transceiver by Signal/One, a subsidiary of the computer firm N.C.R. This makes full use of s.i.cs, including 16 digital and 14 linear types, 60 silicon transistors (including dual-gate f.e.ts in the front-end) with one ceramic/metal power valve which has a beryllia coupler block to transfer anode heat to its own extruded heat sink and with many other features to put it into the *de luxe* class.

In Brief: Winner of the 1970 (33rd) annual B.E.R.U. Contest was R. J. Mills, VQ8CR, of Mauritius, who made 517 contacts in spite of very poor propagation conditions. Runner-up was D. M. MacVicar, VP7DX, while D. L. Courtier-Dutton, G3FPQ, was third and the leading British station. . . . A new v.h.f. beacon transmitter, GB3DM, is now operating from the I.T.A. transmitting site at Burnhope, Co. Durham. The 30-watt transmitter on 145.975 MHz feeds two four-element Yagi aerials beaming north and south, mounted at a height of 98 ft. The station can be received at good range (reports to D. Long, G3PTU, Croesor, Iveston Lane, Iveston, Leadgate, Consett, Co. Durham). . . . Among recent stations heard on h.f. bands have been FB8XX (Kergulen Islands), 4N2BR (a Yugoslav expedition to one of the Adriatic Islands), HS5ABD Chaing Mai, Thailand, 9Q5QR Kinshasha, Congo, and the Japanese EXPO '70 station JA3XPO. Also reported active recently have been CEOAE Easter Island, VPICP British Honduras, VR5LT Tonga Island and several CE9A stations on South Shetland Island. . . . An intense E-layer disturbance on July 6th resulted in long-range reception of many v.h.f. broadcasting stations on 70 and 95 MHz, and another opening to TF3EA in Iceland. . . . Among the mobile rallies being held in September are: R.S.G.B. Scottish Mobile Rally (19th) at the David Livingstone Memorial, Blantyre, Lanarkshire (details G. A. Hunter, GM3ULP, "The Bungalow", Broomside Braes, Camp Road, Motherwell); Peterborough (20th) at Walton Senior School, Mountsteven Avenue off Lincoln Road, Peterborough; and Harlow (27th) at Magdalen Laver Village Hall (details B. G. Capper, G8CUA, 124 Peterswood, Harlow, Essex). . . . A special event station, GB3WAC, will be operating from the Sea Cadet's HQ, Scotch Yard, Tonypandy, during the World Archery Championships (Sept. 10th-12th). . . . A special station, GB3CWR (or GB3CWI) will be run in connection with the Cumberland Federation of Women's Institutes "Golden Jubilee" celebrations (Sept. 6th-12th). . . . Contests during September include the V.H.F. National Field Day (6th) in conjunction with the international I.A.R.U. contest; "Worked all Europe" DX Contest (12th to 13th, phone section): 3.5-MHz Field Day (13th); and National Final of the R.S.G.B. D/F contest (20th. by Slade Radio). . . . A new world record for 13 cm is being claimed for a 249-mile contact in the United States (W4HHK and WA4HGN/P). . . . A 144-MHz moon-bounce contact has been made between New Zealand (ZL1AZR) and California (K6MYC).

PAT HAWKER, G3VA

Literature Received

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

ACTIVE DEVICES

Power microcircuits are the subject of some literature we have received from AEI Semiconductors. Carholme Rd. Lincoln.

PM5A. Thyristor /rectifier diode combination, 800 V, 5A WW401
PM6A. Diode/diode. 1.4kV. 6A WW402
PM7A. Bridge rectifier. 1.4kV. 7A ..WW403

Motorola have produced a new journal called *Semiconductors* which is to be produced every 3 months. The first issue contains articles on integrated circuits for industry and computers. gives circuits for a 400MHz wideband amplifier. a 1Hz low-pass filter. and low-voltage inverters. and describes an integrated circuit f.m. stereo decoder ...WW404

A 96-page manual called "Power transistors for amplification switching and control" has been produced by RCA. It covers physical theory. structure. packaging. limiting factors. and the operation and requirements of power transistors in amplification. switching and in control circuitry. The price is $2. RCA Solid State Division. Somerville. N.J.08876. U.S.A.

A range of logic modules employing high threshold logic. d.t.l.. t.t.l. and e.c.l. are the subject of a catalogue from Jasmin Electronics Ltd. 1 Meriden Close. Hainault. Ilford. Essex. The products consist of integrated circuits on printed circuit cards WW405

"Reliability Report—digital integrated circuits —May 1970" may be obtained from the National Semiconductor Corporation. 2900 Semiconductor Drive. Santa Clara. California 95051 ....WW406

"Planar Power Switching Transistors" is the title of a new Mullard publication for design engineers. E.E.D.. Mullard Ltd. Mullard House. Torrington Place. London WC1E 7HD WW407

Marconi-Elliott Microelectronics Ltd. Witham. Essex. have recently started production of 54-74 series t.t.l. and have published a data sheet on their range WW408

PASSIVE COMPONENTS

Bulletin 94033 from Brush Clevite Co. Ltd. Thornhill. Southampton. Hants. describes a 10.7MHz ceramic filter (type FM4) intended for use in f.m. i.f. strips. The component has a 3dB bandwidth of typically 235kHz (40dB. 900kHz max.) and an insertion loss of about 3.5dB WW409

An alarm annunciator capable of displaying. 12. 18 or 24 legends is described in a data sheet from Mimic Diagrams and Electronics Ltd. Maxim Rd. Crayford. Kent. The unit contains logic for alarm sequences WW410

Condensed catalogue 8010 from Penny and Giles Ltd. Mudeford. Christchurch BH23 4AT. Hampshire. deals with a range of precision conductive plastic potentiometers ....WW411 From the same company comes a fuller catalogue containing details of rectilinear and rotary potentiometers; pressure. force. acceleration and rotary transducers; and servo and signal conditioning equipment WW412

A wall chart devoted to wirewound trimming potentiometers may be obtained from Electrosil Ltd. Pallion. Sunderland. Co. Durham WW413

"Manual of inverter transformers and modules" has been published by Gardners Transformers Ltd. Christchurch. Hampshire BH23 3PN WW414

A pocket booklet describes the products of B & R Relays Ltd. Temple Fields. Harlow. Essex. WW415

Limit switches. microswitches and proximity switches together with various process timers are the subject of a new 20-page catalogue from Omron Precision Controls. 313 Edgware Rd. London. W.2. ....WW416

We have received the following literature concerned with audio equipment for professional use from Vitavox Ltd. Westmoreland Rd. London. N.W.9.

B80 and B M100 microphones WW417
B50 series microphones WW418
B60 and B64 microphones WW419
Microphone stands WW420
Pressure units data sheet WW421
Type 190 circular horn WW422
Radial diffuser. lightweight hailer and 1000Hz horn WW423
Series 550 multicell horns WW424
Series 220 multicell horns WW425
Bitone loudspeaker systems (halls and cinemas) WW426
Mini-bitone speaker system (special effects in halls and cinemas) WW427
Dividing networks (cross-over) WW428
15-inch (380mm) ceramic magnet loudspeakers WW429
12-inch (305mm) ceramic magnet loudspeakers WW430
3-inch (84mm) WN350 loudspeaker unit WW431
Trade price list WW432

An integrated circuit patchboard which will house up to 14 packages (or more with a plug-in extension board) and which contains pulse generation. input and indication facilities is described in a leaflet from Elliott Instruments Ltd. Station Industrial Estate. South Woodham Ferrers. Nr. Chelmsford. Essex WW433

APPLICATION NOTES

A leaflet "Use of 10.7MHz ceramic coupled mode filters in linear i.c. i.f. strips" is obtainable from the Brush Clevite Co. Ltd. Thornhill. Southampton. Hants WW434

"Power Microcircuit" describes the use of the PM5A. 6A and 7A devices mentioned in the Active Devices section. AEI Semiconductors. Carholme Rd. Lincoln WW435

Application report No. 6 from Brookdeal Electronics Ltd. Market St. Bracknell. Berkshire is called "Measurement of contact potential by the vibrating capacitor method" WW436

EQUIPMENT

High-quality audio power amplifiers for use in the broadcasting and sound recording industries with outputs up to 100W are the subject of a leaflet from H/H Electronic. Industrial Site. Cambridge Rd. Milton. Cambridge CB4 4AZ WW437

The following catalogues of American equipment are available from Wessex Electronics Ltd. Stover Trading Estate. Yate. Bristol BS17 5QP.

Wiltron Company. Sweep generators 0.1 to 18GHz. phase meters. vector network analysers. line stretchers. v.s.w.r. automatic testers. directional detectors and r.f. detectors WW438
Anzac Electronics. Broadband r.f. and microwave signal processing devices. 0.5MHz to 3GHz WW439
Wavetek Company. Function generators from sine. square and triangle manual instruments to fully programmable waveform synthesizers from 0.0015Hz to 10MHz WW440
Philco-Ford. Sierra Operations. High-power signal generators 50MHz to 3GHz. wattmeters. coaxial loads and attenuators WW441
Birtcher Corporation. Transistor and integrated circuit test equipment. computer controlled automatic test equipment WW442

The data sheets listed below were sent to us by Beglec N.V.. 718 Houba de Strooperlaam. Brussels 2.

Rodec Fader. 22Hz to 32kHz WW443
Rodec Stereo Mixer. CU887. Four-channels with bass and treble controls. comprehensive switching facilities and twin VU meters WW444
Rodec Mixer. CU881. Four-channel stereo mixer WW445

Audio equipment. an amplifier (model 9000) and an f.m. tuner (model 1500). is described in literature from Bryans Amplifiers Ltd. 18 Greenacres Rd. Oldham. Lancs WW446

Two leaflets from Culan Electronics Ltd, By Ormiston, East Lothian, Scotland, describe a.c. power controllers up to 3kW..................WW447

SE Laboratories (Engineering) Ltd, North Feltham Trading Estate, Feltham, Middlesex have available a leaflet which briefly describes the test equipment they produce WW448

New Products

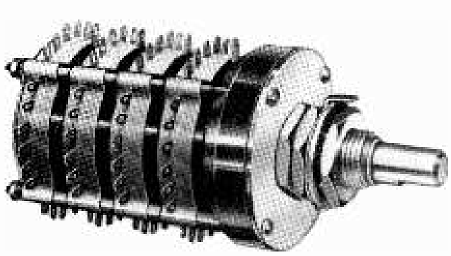
Flying Spot Scanner Tube

A new screen phosphor is used in the Mullard, five-inch, flying spot scanner tube type Q13-110GU. The tube has magnetic focusing, a metal-backed screen, and a spark gap to prevent internal flashover between the anode and the grid. It operates with an anode voltage of 25kV and has a resolution better than 1000 lines. The new phosphor, a cerium activated yttrium aluminate, has an emission peak at 550nm and an extremely short decay time of less than 0.1μs. It is mixed with a blue phosphor (caesium activated yttrium silicate), which also has a short decay time. Because of the short decay time of the phosphors, the flying spot scanner tube can follow fast moving objects without blurr. Mullard Ltd., Mullard House, Torrington Place, London W.C.1.

WW 302 for further details

Miniature Rotary Switch

Highland Electronics have announced a miniature rotary switch in the Highland/Grayhill range—series 53. The switches have the following features: up to

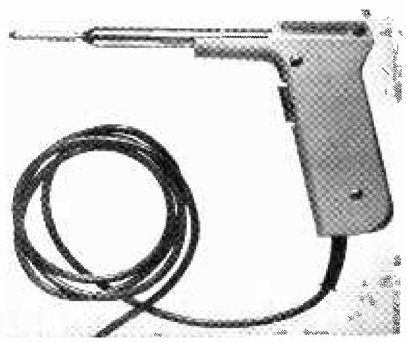


24 positions (15° indexing angle); up to 12 decks; up to 12 poles per deck; diameter over tags 1.35in; gold plated contacts. Panel and spindle seals are available. Highland Electronics Ltd, 33-41 Dallington Street, London E.C.1.

WW 303 for further details

Quick-heating Iron

The Ersa Sprint quick-heating soldering iron, made in West Germany, is available from Home Radio (Components). Warm up time is less than 15 seconds. When the press switch is released the iron starts to



cool, but remains sufficiently hot to continue to melt solder for about 1 minute. The technique for using it therefore seems to be to press the switch for 15 seconds, then release it just at the point of starting to make the soldered joint. The switch could be held down for the whole time of soldering, particularly if there was rapid heat conduction from the joint, but this would cause the bit to oxidise more rapidly. The iron proper weighs less than 5 ounces (7 ounces with lead and plug), and when laid down on the bench after use it is balanced so that the bit does not touch the bench. Although the iron is marked 220V the suppliers state that it is suitable for all a.c. voltages from 220V to 250V and are prepared to guarantee it for use within this range. Spare parts are available and any part can be changed very quickly. Price £3 19s 6d. Home Radio (Components) Ltd, 234-240 London Road, Mitcham, Surrey, CR4 3HD.

WW329 for further details

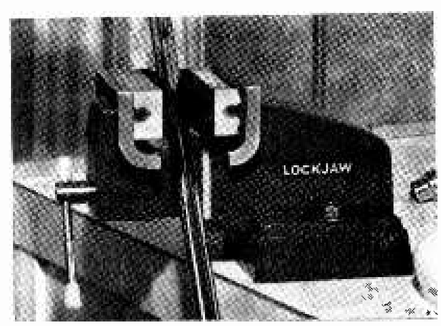
Op. Amp Dual Output Supply

A dual output supply to power operational amplifiers is available from Lambda Electronics. The outputs are ± 12 to 15V d.c. at 400mA for each line. The voltage difference between the lines is maintained to within 1% of absolute voltage or 0.1% change for all variations of line, load, and temperature. The unit, designated LXD-3-152, gives series connected dual outputs. Convection cooling is used and no external heat sinks or forced-air systems are required. Regulation is 0.1% of line or load; ripple and noise 1.5mV r.m.s., 5mV peak to peak. Remote sensing is provided. Current limiting is fixed and current ratings at different operating temperatures are clearly specified. The unit measures 80 × 95 × 130mm. A 24 to 40V version is also available. Lambda Electronics, Marshland Road, Farlington, Hants.

WW 318 for further details

Lockjaw Vice

A vice, called the Lockjaw, has been introduced which is suitable for industrial or domestic use and incorporates a number of novel features. The vice has 104mm ($4\frac{1}{16}$ in) jaws, is 200mm (8 in) long and 100mm (4in) high and weighs about 1.75kg. The silicon metal jaws have indents and slots machined in them so that tubes and discs can be held without excessive pressure and therefore without damage. The jaws can be easily prised off and the rear jaw can be reversed (turned upside down). If this is done the rear jaw is free to pivot so that it automatically adjusts to the shape of any tapered objects that have to be clamped, again avoiding excessive damage. The vice as described above has a recommended retail price of 86s. For more delicate work rubber-faced jaws are available as an extra (14s a pair). There are three ways of mounting the vice. In its standard form it can be bolted to a bench in the normal way. A suction base can be purchased (30s) which allows the vice to be firmly attached to any flat



non-porous surface without damaging the surface in any way. Finally a substantial G-clamp is available (30s) which can be bolted to the vice so that the vice can be mounted on any convenient bench, girder, etc. Vice and Workholding Co. Ltd, 149a Crayford Rd, Crayford, Kent.

WW 301 for further details.

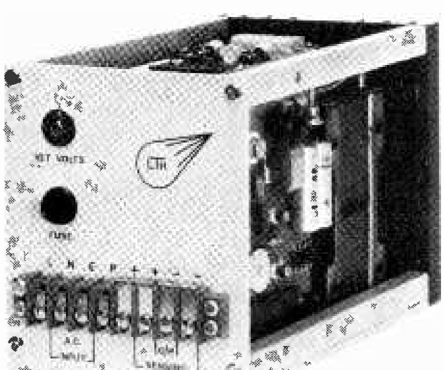
Spot-frequency Marine Receivers

Two multi-channel, single-sideband receivers are introduced by Eddystone Radio. Both have crystal-control and operate on switch-selected spot frequencies. One of these models, type EC964/1, provides a choice of 52 spot frequencies in the medium- and high-frequency bands, while the other, type EC964/3 offers a choice of 28 spot frequency channels in the m.f. maritime band. The two receivers meet the British Post Office specifications TSC 102 and TSC105 respectively. The EC964/1 is primarily intended for use in ocean-going ships, while the EC964/3 is suitable for

ships operating in coastal waters where the higher frequencies are not normally used. Both receivers cover the international distress and calling channel (2182 kHz), other frequencies being chosen by the customer before delivery. Channels are switch-selected to simplify operation. Reception facilities cover double-sideband and single-sideband voice transmission, upper-sideband being accepted in the latter mode. An integral monitor loudspeaker is fitted and outlets are provided for telephones and remote line circuits. Operation can be from any standard a.c. supply or from low-voltage d.c. using a G.P.O. approved d.c./a.c. converter type 978. GEC-Marconi Electronics, Eddystone Radio Ltd, Marconi House, Chelmsford, Essex.
WW330 for further details

Modular Power Supplies

LTH Electronics have launched a series of compact, high specification stabilized power supply modules. Known as the LRA series, it is designed for installation into customers' equipment and is built as a 19in rack-mounting module. Both single and twin units are available with preset voltages in the range of 1-30V and 30-50V at 1, 2, 5 and 10A. The stability ratio is greater than 200,000:1 and typically 500,000:1. A fast acting, automatic-reset, over-current circuit affords complete protection against short circuit and overload. Remote sensing enables the correct voltage to be maintained at the load terminals and integral over-voltage units are available for the protection of



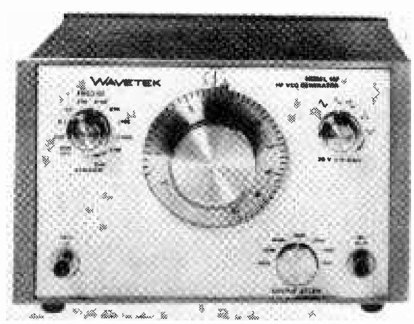
expensive external loads. The output resistance is less than 0.0005Ω. Ripple and noise is less than 200μV. LTH Electronics Ltd, Eltelec Works, Chaul End Lane, Luton, Beds.
WW327 for further details

Quartz-crystal Units

The McKnight Crystal Company supply a range of hermetically sealed, nitrogen filled, metal-cased AT-cut quartz-crystal units covering the frequency band 1400kHz to 20MHz. These units are available in several holder styles including miniature Def. types D, J and K. A wide range of frequency tolerances is available for oscillator or filter applications. Frequency calibration is to customers' requirements. McKnight Crystal Company, Unit 21, Shipyard Estate, Hythe, Southampton SO4 6DE.
WW333 for further details

Function Generator

Model 142 HF VCG generator from Wavetek (available through Wessex Electronics) is a precision signal source with a frequency range of 0.0005Hz to 10MHz. In addition to sine, square and triangle waveforms model 142 offers a variable duty cycle on all output waveforms. This can be continuously adjusted from 5% through the usual 50% to 95%. This means that pulse outputs can be generated with on-off ratios of 1:19, 19:1 or anywhere between. A sawtooth



waveform with a rise-fall ratio of 1:19, 19:1 is also possible. In addition to the 15V pk-pk into 50Ω, the 60dB step attenuator allows clean signal levels as low as 15mV (pk-pk). An external sweep signal (either d.c. programming or a.c. modulation) can be applied to sweep the output over a 1000:1 ratio. Square-wave rise and fall time is less than 20ns. Wessex Electronics Ltd, Stover Trading Estate, Yate, Bristol BS17 5QP.
WW332 for further details

Time Scale Generator

The Comark time scale generator, type 1401, provides a 'rule-scale' timing waveform from 1μs to 10s with an accuracy of 0.005%. There are three simultaneously available waveforms. The rule-scale waveform consists of equispaced

rectangular pulses of three distinct amplitudes, every fifth and tenth being accentuated without change of duration. An 8-position switch selects the scale by specifying the least significant interval. The unscaled outputs are 1:4 mark/space ratio pulses from seven separate outputs with periods from 1μs to 1s. The squarewave output has unity mark/space ratio and an interval of 10 times the periods indicated by the selector switch. The output stage has a rise time of 40ns and a fall time of 30ns. The internal 1MHz reference can be replaced by an external clock signal for increased accuracy or to permit generation of long time periods. Two instruments, used in cascade, will generate periods up to 10^8 sec. Comark Electronics Ltd, Brookside Avenue, Rustington, Littlehampton, Sussex.
WW 324 for further details

3.5-W Amplifier

A low-cost version of the General Electric (U.S.A.) PA246 5-W amplifier, the PA263, is available from Jermyn. Designed for supply voltages up to 30V it will deliver 3.5W r.m.s. continuously with about 9mV input. It may be used with various supply voltages, loads and inputs by adjusting external components. Price for quantities 1-24 is 33s (plus 2s 6d post and packing for small orders). Jermyn Industries, Vestry Estate, Sevenoaks, Kent.
WW328 for further details

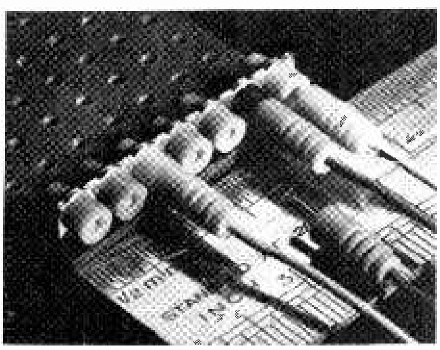
Dual Pulse Stretcher

A noise-free pulse of almost any width can be generated by the type MC 675 pulse stretcher from Motorola. The width of the output pulse generated by this circuit is determined by the width of the input pulse and a time constant derived from an external capacitor and either an external or an internal resistor. Two modes of operation are made possible by the inclusion of a two-input NOR gate at the output. When the signal levels at both inputs of this gate are low and with the input pulse fed to the pulse-stretcher input terminal only, the unit operates as a

conventional pulse stretcher to generate a pulse the duration of which is related to the input pulse duration as described above. The unit can also operate as a monostable multivibrator. In this configuration, no output pulse is generated for an input noise pulse of a duration (in seconds) equal to 10 times the value of the external capacitor (in farads). Propagation delay is 110ns and the power dissipation is 180mW. Encapsulation is a plastic TO-116 package. The temperature range is –40 to +75°C. The price is 45s each for quantities of 100 and above, Motorola Semiconductors Ltd, York House, Empire Way, Wembley, Middx.
WW331 for further details

Printed Circuit Sockets

Oxley Developments Company have added a subminiature socket, type 30S/093/PCB, to its range of miniature printed circuit sockets. Designed to accept the standard Oxley 30P/093 plug, the sockets mount on a standard 0.150 in (3.8mm) module printed circuit board. Plugs are inserted in a plane parallel to the



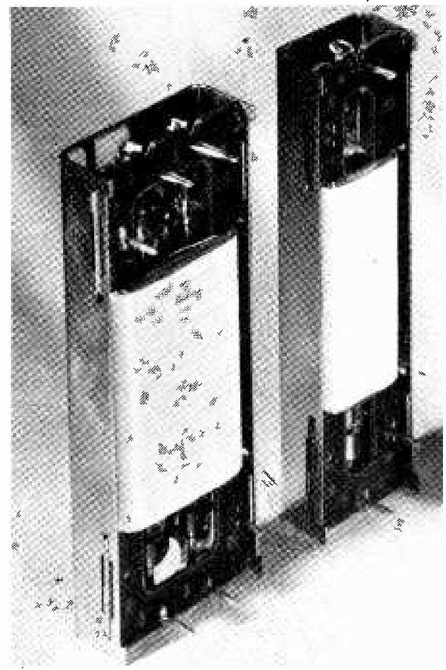
printed circuit board to allow compact board stacking. Single or multiple units with up to ten outlets can be supplied: p.t.f.e., insulating bushes are available in eleven colours to BS 2746; socket contacts are gold-plated brass. Oxley Developments Co. Ltd, Priory Park, Ulverston, Lancs.
WW 313 for further details

50-MHz Digital Frequency Meter

Orbit Controls have announced the first of a new family of electronic instruments for 19in rack mounting with a panel height of only 1¾in. Digital frequency meter type 71C 100 is a direct reading 5-decade instrument covering the spectrum 20Hz to 50MHz. Input sensitivity over the range is 10mV, and the dynamic range is better than 60dB. Three programmable gate times and four programmable display times are available. Selection of gate and display time is remotely programmed via a multipole socket on the rear panel. The instrument is designed so that it may be used to indicate the incoming frequency to a communications receiver by monitoring the local oscillator frequency. To provide direct reading, up to six frequency offsets are provided and the particular offset, equal to the receiver intermediate frequency, may be selected to add to or subtract from the local oscillator frequency as appropriate. Selection of frequency offset is, again, remotely programmed via a multipole socket on the rear panel. The front panel of the instrument is clear of controls other than a test button which enables the quartz crystal derived timing signal to be switched to the input circuit. Input, outputs. switch, etc., are located on the rear panel. The display module may, if required, be removed from the instrument and connected by an external cable to a further socket on the rear panel to provide a display remote from the instrument. An alternative version, type 71C 101, is also available, having gate and display times selected by means of front panel push-buttons. In this version the frequency offset, if required, is prewired internally. In other respects the two instruments are identical. Orbit Controls Ltd, Alstone Lane Industrial Estate, Cheltenham, Glos. GL51 8JQ.
WW 309 for further details

Low-cost Dry Reed Relays

A range of dry reed relays has been introduced by ITT Components Group Europe. The HRE 399, HRE 599, HRE 899 and HRE 831 are low-cost general-purpose industrial relays incorporating up to four "normally open" (make) switches. They are designed for printed-circuit mounting on boards up to 2.5mm thick, with the mounting lugs and electrical terminals being on a grid spacing of 2.54mm. The depth of the relays is 10.5mm, and magnetic and mechanical protection is provided by a metal shield. The HRE 399 relay has a maximum switching current of 0.6A a.c./d.c., maximum switching voltage of 220V a.c./d.c. and a maximum switching power of 12VA. Maximum switching current for the HRE 599 is 1A a.c./d.c.; the maximum switching voltage is 250V a.c.



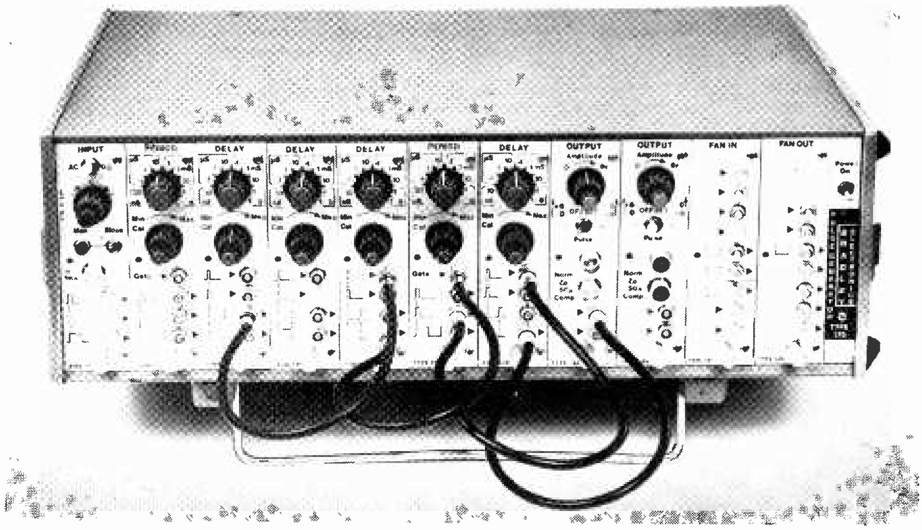
or 90V d.c. and the maximum switching power is 24VA. For the HRE 899 relay the maximum switching current is also 1A a.c./d.c., with the maximum switching voltage being 250V a.c. or 150V d.c. and the maximum switching power 60VA. The HRE 831 has the same performance as the HRE 899 relay. ITT Components Group Europe, Power Components Division, West Road, Harlow. Essex.
WW 308 for further details

Signal Amplifier

The Bournlea Dynamic d.c.-a.c. amplifier has a bandwidth from d.c. to 500kHz, 25V pk-pk output swing and employs f.e.t. input stages to provide a drift of less than 50μV/°C and a noise level of 25μV r.m.s. at 60dB gain. Alternative overall gain settings of 0, 20 or 40dB are selectable by switch, the gain accuracy being 0.5dB. The 1MΩ input resistance is accurate within $\pm$2% and makes it possible to use the amplifier as an accurate high gain current to voltage converter down into the nA and pA region. Either a.c. or d.c. input coupling can be selected, the low frequency cut-off being at 1Hz in the a.c. mode. The output d.c. potential is adjustable by a single front panel control the setting of which is unaffected by the input source resistance. The output resistance is less than 20Ω. Operation is from self-contained batteries. The amplifier is available from The Cardon Instrument Co., Earls Colne, Colchester, Essex.
WW 307 for further details

Modular Pulse Generator

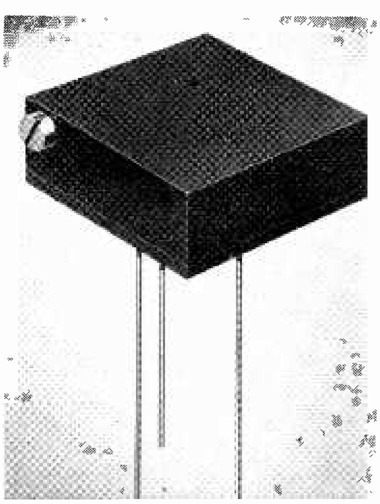
The Bradley Modular Pulse Generator 176 is a 50 MHz instrument for generating various complex pulse patterns. The use of modules in a basic main frame provides flexibility—the modules can be arranged in any order and additional units added as required. The main frame which



includes the stabilized power supply, can accommodate up to 11 modules. These are interconnected by coaxial cables. Six modules are currently available, the basic units being a period module capable of producing any p.r.f. from 1Hz to 50MHz, and a delay/width module providing pulse widths or delays from 10ns to 1s. Other modules include a high-impedance variable-level input unit, and an output module providing up to 6V and having a risetime of 2ns. Fan-in and fan-out units provide expansion, pulse-width subtraction, and inversion. Separate sockets are used for each interconnection, and all outputs are short circuit proof. G. & E. Bradley Ltd, Electral House, Neasden Lane, London, N.W.10.
WW 326 for further details

Square Trimmer Resistor

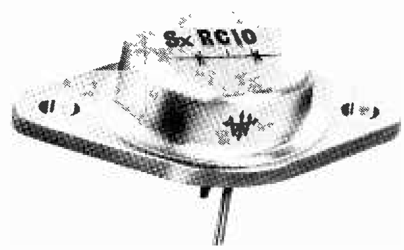
Reliance Controls have introduced a new ½-in square wirewound fully sealed trimmer. The trimmer is available in three versions, designated CW65, CW66 and CW67, dependent upon the pin configuration required. Suitable for printed circuit board mounting, the various styles allow the designer to make the maximum use of available space. Resistance range of the CW65/66/67 trimmers covers 10Ω to 25kΩ. Mechanical adjustment is 25 turns and temperature range −55 to +155°C. Wattage rating (whole element uniformly



loaded) is 1W at 20°C, derating to zero to 155°C. Insulation resistance is 1000MΩ at 500V d.c. The wiper is of gold plated beryllium copper, and the terminals are gold plated. Reliance Controls Ltd, Drakes Way, Swindon, Wilts.
WW 316 for further details

Centre-tap Silicon Diode Assemblies

The Semiconductor Division of Westinghouse Brake and Signal Co. have introduced a series of silicon double-diode rectifiers rated at 20A with voltage ratings of 100-600V. This series of centre-tap assemblies is designated



SxRC10 and SxRN10. Higher than usual overload to mean-current rating ratio is claimed. Housing is a standard TO-3 glass/metal package which permits mounting on normal transistor heat sinks. Westinghouse Brake and Signal Co. Ltd, 82 York Way, King's Cross, London N.1.
WW 306 for further details

Thyristors for High-frequency Inverters

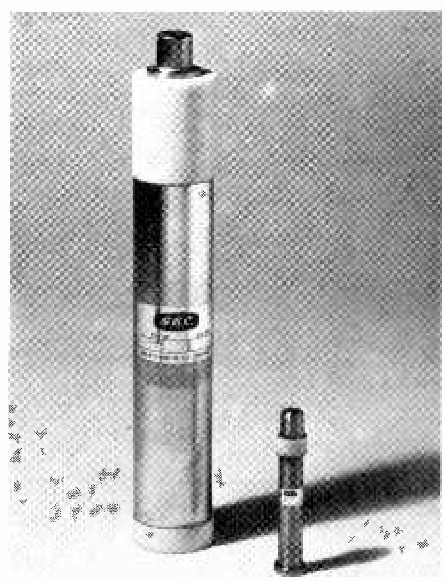
Two new families of thyristors available from Mullard have high *di/dt* and *dv/dt* ratings of 100A/μs and 200V/μs respectively with crest working voltages of 800, 1000, and 1,200V. Types BTW23 and BTW24, are p-gate, reverse blocking, avalanche devices intended for use in high-frequency inverters and motor-control circuits. The BTW23 thyristors have a maximum forward current rating of 70A and an avalanche rating of 40kW when a square pulse of 10μs duration is applied and the junction temperature is 25°C. The thyristors have an encapsulation similar to the TO-49 with an M12 thread on the stud. The BTW24 thyristors have a forward current rating of 30A and an avalanche rating of 20kW (10μs square pulse when junction temperature is 25°C). These thyristors are contained within TO-103 encapsulations. Mullard Ltd, Mullard House, Torrington Place, London W.C.1.
WW 322 for further details

100-W Power Transistor

Microwave Associates have announced the first 100-W power transistor at 1GHz, designed specifically for pulse applications. The device will provide 10dB gain as an amplifier, and 60 watts output as an oscillator, at 1GHz. Microwave Associates Ltd, Cradock Road, Luton, Beds.
WW 310 for further details

Corona Stabilizers

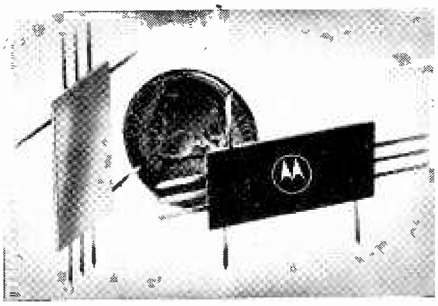
The M-O Valve Company has extended its range of metal/ceramic corona stabilizers with the introduction of the SC8 and SC9 series. The SC8 (the larger in the photograph) provides stabilized voltages in the range 25 to 50kV in six 5kV stages and the SC9 in the range 350 to 2000V in 11 stages of from 50 to 200V. Specific voltage types can also be made to



customers' special requirements. After one minute the voltage remains constant to within 0.25% of the initial value for at least two hours, and long-term stability is better than 1% per 1000 hours. The temperature coefficient of the SC8 is 0.005%/°C while for the SC9 it is measured at 0.3V/°C over the temperature range −40 to +90°C. The M-O Valve Co. Ltd, Brook Green Works, London W.6.
WW 323 for further details

Microwave Duplexer

Type MCH5890 microwave duplexer from Motorola consists of two



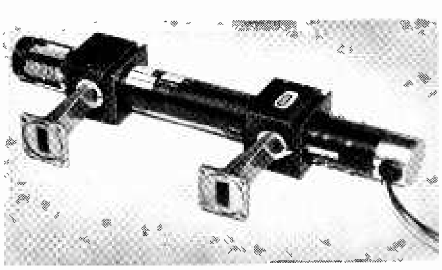
step-recovery diodes, one connected in series between the transmitter terminals and the receiver terminals, and the other one shunting both pairs of terminals. The device can accept a 40W input power. During transmission the insertion loss between aerial and transmitter terminals is only 0.1dB while the isolation between the aerial and receiver terminals is 25dB. Motorola Semiconductors Ltd, York House, Empire Way, Wembley, Middx.
WW 319 for further details

M.O.S. Arrays for Electronic Organs

Two m.o.s. arrays specifically designed for electronic organs are available from WEL. Array type MA70 is a $\sqrt[12]{2}$ divider giving true semi-tones. Array type MA60 is a six-stage binary divider giving six outputs each one octave apart. (An electronic organ can be made with four MA70 circuits and twelve MA60s.) For organ designers requiring a free-phase system 30 MA70s could be used. Prices are about £3 10s. for the MA70 and 25s for the MA60 (for quantities in excess of 1,000 pieces). WEL Components Ltd, 5 Loverock Road, Reading, Berks.
WW 320 for further details

Lightweight X-band Travelling-wave Tube

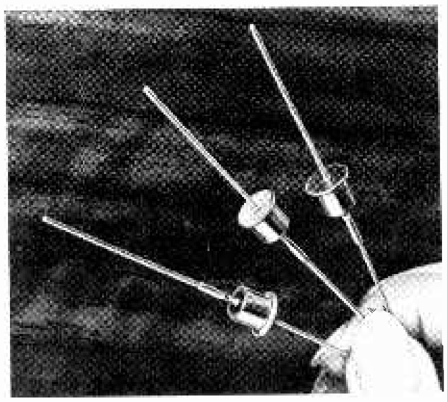
The M-O Valve Company has introduced a low-cost lightweight X-band travelling wave amplifier, type TWX34, giving a minimum gain of 35dB and a saturated power output of 1W over the frequency range 7.5 to 11 GHz. A typical application is in the ground station of a transportable satellite communications system where continuous operation in ambient temperatures of up to 60°C and ability to withstand shock, vibration and high humidity conditions are required. A low wattage oxide coated cathode has been incorporated and total power supply consumption under full r.f. drive conditions is less than 25W. Power supply connections are by flying leads, but other



types of connection can be provided on request. The M-O Valve Co. Ltd, Brook Green Works, London W.6.
WW 321 for further details

Switching Diode

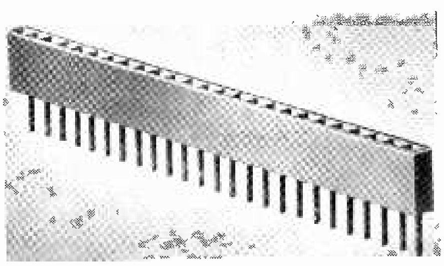
A silicon planar, epitaxial, whiskerless diode, type BAV44, is announced by Mullard. It can rapidly switch currents up to 1A, and is primarily intended for use in servo amplifiers, digital voltmeters and oscilloscopes. Maximum continuous



reverse voltage is 65V and the maximum voltage drop with a forward current of 1A is 1.25V. Switching time (t_{rr}) from I_F= 1.0A to I_R = 1.0A and measured at I_{rr} = 100mA is 22ns. Diode capacitance is 7.5pF. Mullard Ltd, Mullard House, Torrington Place, London W.C.1.
WW 325 for further details

Integrated Circuit Socket

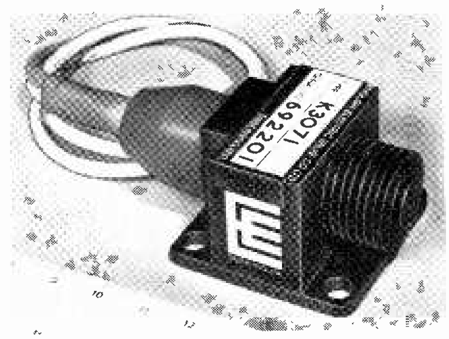
Employing a new concept in socket design for integrated circuits and semiconductor devices, the A23/2033 now available from Jermyn, is a 25-contact single strip socket capable of accepting either flat or round



leads. The body of glass-filled nylon is fitted with gold flash phosphor bronze contacts, giving a life of up to 10,000 insertions with an electrical resistance of 15m Ω. Jermyn Industries, Vestry Estate, Sevenoaks, Kent.
WW 305 for further details

Oscillator Klystron

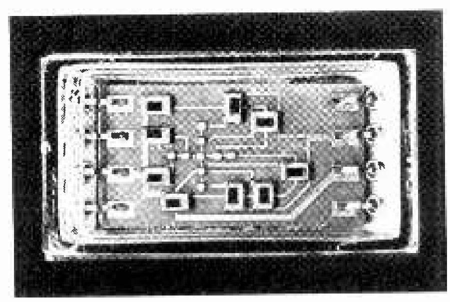
A forced-air cooled, fixed frequency, two cavity oscillator klystron (K3071), for operation in airborne Doppler and beacon radar equipment, has been introduced by English Electric Valve Co. The tube incorporates a heater design claimed to give a lower level of heater f.m. than



hitherto attainable. In addition, the sturdy construction results in a tube having low noise and low microphony. A further improvement is its longer life expectancy compared with earlier types. Operating frequency of the K3071 is 8800 ±5MHz and a typical output power is 1.5W, but variants can be supplied for operation at any frequency in the 8 to 9.5 GHz range. Silicone rubber moulded connections are used for unpressurized high-altitude operation. English Electric Valve Co. Ltd, Chelmsford, Essex.
WW 315 for further details

Solid-state Plus/Minus Sign

To indicate plus or minus information, a light-emitting diode display module has been added to Hewlett-Packard's numeric indicators. The model 5082-7018 plus/minus sign comes in the same package as the 5082-7000 series of numeric indicators. It operates at less than



5V. A plus or a minus is displayed by applying a drive voltage to one (minus sign) or two (plus sign) input leads. A third lead is the earth connection. Brightness can be varied by changing the d.c. drive between 1.6 and 4V. Hewlett-Packard Ltd, 224 Bath Road, Slough, Bucks.
WW 317 for further details

Schmitt Trigger I.C.

SN7413N Schmitt trigger combines two 4-input positive nand gates. Each gate has different input levels for positive and negative going signals. Hysteresis (the difference between two threshold levels) is typically 800mV. Noisy pulses can be accommodated without false trigger regardless of change of temperature. The units also fire reliably with rise and fall times slower than 1V per second. The unit is housed in a 14-pin dual-in-line plastic package and costs 10s 1d each in quantities of 100. WEL Components Ltd, 5 Loverock Road, Reading, Berks.
WW 311 for further details

Real and Imaginary

"Have we eaten of the insane root that takes the reason prisoner?"

Now that the dust of the General Election has settled, with the computers, both electro-mechanical and human, duly confounded,* we are now looking forward with differing degrees of trepidation to the delights the future holds under our new masters.

The political parties are unanimous in plumping for increased production for export as the prime means of salvation. This being so, I thought that for a few minutes we might do worse than to take a look at our own electronics industry to see whether anything could be done to improve existing performance.

The position of industry in general can be likened to that of a farmer who possesses twenty cows. Under Communisn, the State takes the cows and sells some of the milk back to the farmer. Labour buys the milk from the farmer and pours it down the drain. The Conservative policy is to get the farmer to sell five cows and buy a bull. The Liberal outlook is much the same except that they would employ A.I. All, however, subscribe (by implication at least) to the premise that the bigger the organization the more efficient it is; so the trend has been towards the scooping of a multitude of little electronics concerns into a few gigantic ones.

We all know the arguments for mergers—streamlining, rationalization, the avoidance of duplication of effort, are all stock phrases—but, according to a recently published study,† most takeovers and mergers are made for the security and convenience of the management of the bidding company. Usually (it says) the major motives are to gain bigger shares of their markets, reduce competition or protect themselves from other predatory firms.

Be that as it may, the concept that the greater the size, the greater the efficiency, is rubbish, and those who argue for it take no account of human nature. As an instance, those of you who run your own small businesses won't need to be told that economies in small matters make the difference between extracting a living and going broke.

* With one exception; see this page, May issue.
†("Management and Merger Activity" by Gerald Newbould.)

Not so in the big organization. To the clerk who is writing a letter about a £250,000 order, a second sheet of paper is too trifling a matter to be considered and it is far too easy to reach for a telephone to save the fag of writing.

In such circumstances the management's typical reaction is either nil, which is bad, or to create a Gestapo department to deal with it, which is often worse. Control departments of this character are frequently parasitic growths, sucking out an existence from the main stem of profits, for they can, all too easily, cost more to maintain than the savings they effect.

A typical example of the control system now in vogue is P.E.R.T. (the initials stand for Programme Evaluation and Review Technique). In essence this is a semi-diagrammatic method of ensuring that all the bits and pieces for a large contract come together at the right times and places so that the job flows smoothly to completion on schedule. It also sets out to keep tabs on costs and to aid the solution of various management problems. It is the sort of thing that no trendy company would dream of doing without, on the grounds that present-day systems orders are on so large a scale that it would be impossible to control them efficiently in any other way. But some while ago I was at a gathering which included several veteran engineers and I was amazed to hear from one of them of the size and complexity of some of the radio communication systems which were undertaken between the wars, and even back in the 1914-18 period. Nobody had ever heard of P.E.R.T. in those days, but the jobs got done in far quicker time than they would take today. At that time it was part of engineering training to be able to visualize the demands for materials well before they arose, and to order accordingly. The man who couldn't do that didn't become a senior or chief engineer or if he did, he didn't last long. He was sacked.

Now P.E.R.T. in principle is applied commonsense and, used rationally, is well enough. It is at its interface with human nature that it can fall down badly. Elevate it to the status of a department and it can so easily become a coterie of high priests whose main—although perhaps unconscious—aim is to preserve the mystique and to enlarge the sphere of influence of the cult.

The computer is another control system which is wide open to misuse. Computers tend to be run by an alien race of high priests who, protected from close investigation by reason of the jargon of their calling, can easily let enthusiasm outweigh sober judgement and advocate the use of highly expensive machines for jobs that a desk calculator could do in less time and at a fraction of the cost.

A big combine tends to be riddled with control departments; all were instituted with the best of intentions; some, possibly, save the company money. But it would at any rate do no harm for such a firm to take a cool, hard look at all such areas to see whether they are costing more to maintain than the amounts they save. The grave danger, too, with efficiency-control systems is that they can easily become a cloak for managerial ineptitude and face-saving. When things go wrong their presence can make the difference between the personal admission of "I was to blame" and the shrugging-off comment "Unfortunately there was an undetected error in the computer programming". Such an outlook leads by logical progression to the present prevalent practice of calling in an external firm of management consultants to advise on company reorganization. By definition a manager's function is to manage, and do so to such effect that he makes a profit for his company. If he cannot do this, then he should go.

Are we, in fact, making too much of a fetish of large-scale electronics groupings and business efficiency systems? These approaches of course are an aping of American techniques, but what is sauce for the American goose is not necessarily sauce for the British gander; it should not be forgotten that, in spite of its size and vaunted efficiency, the American electronics industry would collapse like a house of cards if disarmament became a reality and government-sponsored space programmes were stopped.

Now I come to think of it, the humble tea-break is, in its way, symptomatic of the whole business of pseudo-efficiency. Time was when the tea-lady and her trolley did the rounds twice daily, dispensing cheer to all. But, "Inefficient!" said the experts, and proceeded to prove on paper how much cheaper machines would be. "Splendid!" said management and so it came to pass that machines were installed and the tea-lady was out.

After the inevitable teething troubles, things go famously for a while and the revenue rockets. Management rubs its hands, overlooking the fact that a times-three increase in takings signifies that a considerable proportion of the proletariat are now taking several tea-breaks *per diem* instead of the statutory two.

This bit of automation was originally an American idea which has been widely copied over here. But one world-famous firm has, to my certain knowledge, seen the light and has gone back to the once-derided trolley. I hardly like to mention this, but it's located in the U.S.A.

Wireless World, September 1970

F. M. Tuners

Outline details of some up-to-date units

It is fair to suggest that the poor relation of most hi-fi enthusiasts' systems is the radio tuner, if indeed a tuner is included at all.

While at any time the upper crust of the hi-fi fraternity are willing to enter into protracted and sometimes pretentious discussion of purely mechanical parameters such as wow and flutter, mass, compliance, trackability and so on, radio reception is hardly regarded by some as worthy of consideration. Yet this latter via the radio tuner is a truly electronic signal source that could be more actively utilized. Most of the country is now covered by v.h.f. transmitters so that it is possible for the majority of listeners to receive a reliable signal*. If the transmission is a "live" one originating in a studio local to the transmitter, the output from a hi-fi system can be of superbly high quality, especially on stereo.

Hi-fi probably came into its own as a "following", with the introduction of the LP disc and thus made possible music quality in the home far superior to that supplied by long- and medium-wave a.m. radio. But medium-wave radio has nothing to do with hi-fi and it is only at v.h.f. with f.m. transmissions that radio, mainly because of better noise performance at those frequencies, can compete with the record. It is true that some manufacturers supply an a.m./f.m. tuner with medium-, long- and sometimes short-wave reception, in addition to the v.h.f. band. However, no one pretends that quality programmes are possible on the a.m. bands and these are usually included merely to give maximum coverage of the available transmissions, since the v.h.f. band provides only B.B.C. programmes Radio 2, 3 and 4, and some local radio. Also v.h.f. has limited area coverage and foreign programmes generally can only be received on a.m. bands.

Importance of tuning

The distortion content of an f.m. transmission is potentially extremely small. One particular requirement to avoid distortion on f.m. is accuracy of tuning. Unlike the simple a.m. demodulator, if the i.f. signal is not centred on the linear portion of the f.m. discriminator characteristic the demodulated audio signal will be non-linear. Again unlike a.m. reception it is not easy with an f.m. tuner for the listener to find the centre frequency by ear and most tuners feature visual and automatic electronic facilities to assist in tuning and to prevent oscillator drift with time.

*See "B.B.C. Band-two Broadcasting Stations", in the August issue, page 401.

Sensitivity and noise

It is in the matter of audio dynamic range and signal-to-noise ratio that disc recordings could be said to be superior to that of broadcast signals. Because of the need to maintain the broadcast signal at a reasonably high average it is necessary for the B.B.C. to limit the audio dynamic range (the range from the softest to the loudest passage) to a particular ratio; 30-35dB having been quoted. Recordings have not the same limitations of dynamic range. On the question of noise, one big advantage of the f.m. system over a.m. is that extraneous signals which display an a.m. characteristic can be removed from the input signal by a limiter circuit. These include static and man-made interference such as that caused by ignition systems and electrical apparatus. By limiting the maximum amplitude of all signals, the discriminator is presented with a signal of constant amplitude, with variations in frequency only.

Random noise (white noise) contains a.m. and f.m. components and the f.m. when detected is heard as a background noise or hiss. Because of the nature of f.m. detection and the use of a.g.c. circuits the audio output from the detector is reasonably constant and the difference between strong and weak stations manifests itself by the amount of background noise present. To obtain good clean reception (and the worst is always cleaner than a.m.) it is desirable for the tuner to enjoy good sensitivity and low noise when the amplifying stages are working at high gain. This "usable" sensitivity is invariably quoted by tuner manufacturers as so many μV for so many dBs quieting, for example: 2 μV for 30dB quieting being typical.

Noise increases as the audio frequency range is extended, and this is counteracted by emphasizing the high-frequency end of the audio range before transmission and de-emphasizing by the same amount in the receiver. By this means the frequency response effectively remains level but the accompanying noise is attenuated. The de-emphasis figure is quoted as 50μsec which represents the time-constant of the de-emphasis circuit.

As has been stated then the greatest advantage of f.m. over a.m. is the virtual elimination of interference and this combined with transmissions at v.h.f. has allowed a greater audio bandwidth. Some programmes, however, which are routed over many miles of land-line networks before reaching the transmitter, suffer from attenuation of the high frequencies. The U.K. system bandwidth is 15kHz with a deviation frequency of $\pm$ 75kHz. The ratio of deviation frequency to the highest audio frequency (deviation ratio) has a direct influence on noise performance†. The i.f. bandwidth of an f.m. tuner is typically 200kHz at the 6dB points. The response should be symmetrical, and the linear portion of the discriminator characteristic should cover at least 100kHz to prevent distortion. (Grundig claim for their RT100 tuner that the distortion factor is below 1% when it is detuned by 50kHz.) Although present-day tuners are very sensitive and can be operated from relatively crude aerial installations, distortion can arise as a result of multipath reception and for this reason alone it is wise to consider the erection of an efficient outdoor aerial preferably with some directional properties. All v.h.f./f.m. transmissions are horizontally polarized which means that even the simple dipole offers some rejection of reflected signals on a horizontal plane but "aircraft flutter" can be troublesome in some areas. A particularly unpleasant form of distortion can occur from reflected signals which makes piano music sound as if the speaker windings were dragging the magnet pole piece.

Basic requirements

Having settled on a suitable aerial, all that is necessary for enjoyable reception on an existing hi-fi amplifier and speakers is a quite simple, well designed tuner. It

†For further reading: "F-M Simplified" by Milton S. Kiver. Published by D. Van Nostrand Co. Inc., Princeton, New Jersey, U.S.A.

should have good sensitivity so that the limiter is always saturated, adequate bandwidth with good response, a stable oscillator and a discriminator with a high order of linearity. To save further on cost the tuner could be powered from the main amplifier if the amplifier power supply ratings allow but for convenience it is better to buy a tuner unit with its own mains power unit. Such a tuner can be obtained for a modest sum. If stereo reception is required the tuner must incorporate a stereo decoder. Most modern tuners are suitable for stereo reception but if cost is a prime factor a "stereo ready" tuner could be purchased and used on mono. It could be converted to stereo at a later date by means of a plug-in decoder module available from the manufacturer.

Refinements on the more expensive tuners are extras which improve reception in weak-signal or difficult areas and electronic aids to relieve the user of a certain amount of guesswork. Among these are visual tuning indicators which may take the form of an ordinary meter movement, an electronic tuning eye or even a system of coloured lights to indicate which side of the centre frequency the tuner is off-tune and when it is on-tune. Probably the largest single influence on good stereo quality is the decoder itself. Manufacturers favour different methods of extracting the left and right-hand channels from the composite signal but a point to watch out for here is the degree of channel separation. This should be typically 30dB at 1kHz.

In the early days of stereo transmissions some tuners gave a visual indication of when a stereo signal was present and it was then necessary to switch the tuner manually to stereo. Nowadays it is more usual for the stereo signal to switch-in the decoder automatically, at the same time lighting a pilot lamp to indicate that stereo is being received. Many tuners with automatic switching still retain a manual mono override switch because in fringe reception areas or situations where the stereo signal is very weak it may be preferable to hear the programme in mono with less background noise, which is the case when the decoder is muted. It should be mentioned that stereo transmissions in any case contain more noise than mono transmissions when received on a mono tuner because the audio information occupies slightly less of the carrier modulation depth.

To assist listeners in setting up their equipment correctly, the B.B.C. transmits special test signals at scheduled times. Details of these are set out in cols. 1 and 2 opposite.

If it appears that we are dwelling too long on the possibility that programmes are likely to be degraded by noise let us put things in perspective by saying that in the poorest conditions the noise on f.m. is less than is the norm on a.m., especially after dusk. With a "usable" signal, noise on f.m. is virtually inaudible.

Use of new devices

Noise performance has again been improved as of late by the use of field effect transistors. These devices also make it easier to apply automatic gain control than do bipolar transistors and for these and other reasons many makers use an f.e.t. as an r.f. amplifier. Ceramic filters in place of conventional i.f. transformers are coming in and so too are i.cs. The use of i.cs no doubt eases production problems and ceramic filters can be designed with wideband characteristics and sharp cut-off to provide the necessary selectivity without the use of a large number of tuned circuits in the i.f. stages.

More and more manufacturers are now employing capacitance-diode tuning. This allows the ganged tuning capacitor to be dispensed with and tuning to be carried out by a single potentiometer. Sometimes four or more potentiometers are included, each with a separate tuning scale and press-button thus allowing the user to preset his tuner to four or more stations.

Oscillator frequency drift is corrected in most f.m. tuners by an automatic frequency control circuit (a.f.c.) which has

Stereo test tone transmissions

To facilitate channel identification and adjustment of channel cross-talk a 250Hz tone is transmitted every day except Wednesday and Saturday in the left channel only from about four minutes after the end of Radio 3 until 23.55. This test may be interrupted from time to time. On Wednesday and Saturday the following test sequence is transmitted by the B.B.C.

Test No.	Time (Approx.)	Left channel (A)	Right channel (B)	Purpose
1	23.42	250Hz at zero level	440Hz at zero level	Identification of left and right channels and setting of reference level
2	23.44	900Hz at +7dB	900Hz at +7dB, antiphase to left channel	Adjustment of phase of regenerated subcarrier and check of distortion with signal wholly in the (A–B), i.e. S. channel
3	23.48	900Hz at +7dB	900Hz at +7dB, in phase with left channel	Check of distortion with signal wholly in the (A+B), i.e. M. channel
4	23.49	900Hz at +7dB	No modulation	Check of A to B cross-talk
5	23.50	No modulation	900Hz at +7dB	Check of B to A cross-talk
6	23.51.20	Tone sequence at −4dB: 60Hz 900Hz 5kHz 10kHz This sequence is repeated	No modulation	Check of A-channel frequency response and A to B cross-talk at high and low frequencies
7	23.52.20	No modulation	Tone sequences as for left channel on Test 6	Check of B-channel frequency response and B to A cross-talk at high and low frequencies
8	23.53.20	No modulation	No modulation	Check of noise level in the presence of pilot tone
	23.55	Reversion to monophonic transmission		

The schedule is subject to variation or cancellation to accord with programme requirements and essential transmission tests.

The zero level reference corresponds to 40 % of the maximum level of modulation applied to either stereophonic channel before pre-emphasis. All tests are transmitted with pre-emphasis. Periods of tone lasting several minutes are interrupted momentarily at one-minute intervals.

With receivers having separate controls of subcarrier phase and cross-talk, the correct order of alignment is to adjust first the subcarrier phase to produce maximum output from either the A or the B channel during test-2 and then to adjust the cross-talk (or 'separation') control on tests-4 and -5 for minimum cross-talk between channels.

With receivers in which the only control of cross-talk is by adjustment of subcarrier phase, this adjustment should be made on tests-4 and -5.

Adjustment of the 'balance' control to produce equal loudness from A and B loudspeakers, is best carried out when listening to the announcements during a stereophonic transmission, which are always made from a centre-stage position. If this adjustment is attempted during the tone transmissions, the results may be confused because of the occurrence of standing-wave patterns in the listening room.

The outputs of most receivers include significant levels of the 19kHz pilot tone and its harmonics. These components do not interfere with normal listening but do affect most signal-level meters. It is essential, therefore, to provide filters with adequate loss at these frequencies if instruments are to be used for the above tests.

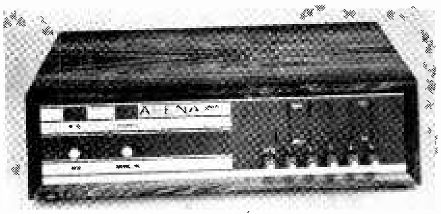
a pull-in range of some 150-200kHz. Provision is commonly included to switch-out the a.f.c. so that the correct tuning point can first be found without the masking effect of the automatic tuning, which is then switched-in to take care of drift. Some listeners may wish to receive an f.m. station in an adjacent transmission area or those living in south-east coastal regions may sometimes receive Continental f.m. stations. These transmissions are necessarily much weaker than the local station and it is useful to be able to switch-out the a.f.c. in this situation to prevent the tuner being "captured" by the strong local signal.

Available units

We circulated all known makers and suppliers of f.m. tuner units in the U.K. for details of their products and from each of those who replied we have selected one model for a short review in the following pages. We stipulated that the tuner must have a v.h.f./f.m. band but it could also cover a.m. bands. Unless stated otherwise the tuners mentioned are for a.c. mains operation. Tuner/amplifiers are excluded. Some specifications are suffixed (I.H.F.) where this has been given on the particular manufacturer's returns. It may well be however, that some unmarked specifications for tuners from other manufacturers are also to I.H.F. requirements but have not been so defined. I.H.F. are the initial letters of an American body, the Institute of High Fidelity, which lays down methods of measurement for audio equipment. In some cases it can provide for enhanced published performance figures.

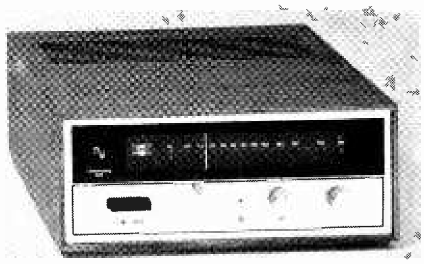
Prices quoted are the suggested retail prices in force at the time of going to press, and include purchase tax.

Arena F211 (f.m. stereo)



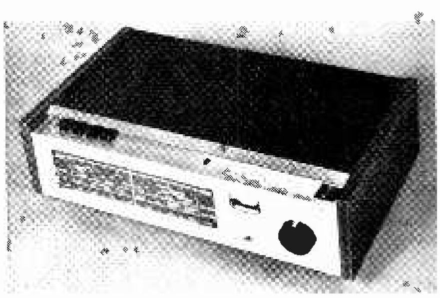
Sensitivity: 3μV for 20dB quieting at 40kHz deviation.
I.F. bandwidth: 250kHz.
Discriminator bandwidth: 600kHz.
A.M. suppression: better than 50dB, limiting from 4μV.
Frequency response: 20-15000Hz.
Channel separation: better than 30dB (250-6300Hz)
Carrier suppression: better than 50dB.
Price: 35gn.
Automatic stereo switching with indicator and mono override switch; switched a.f.c.; capacitance-diode tuning with five press-buttons and scales. U.K. agents: Highgate Acoustics, 184-188 Great Portland Street, London W.1.

Armstrong 524
(f.m., decoder optional)



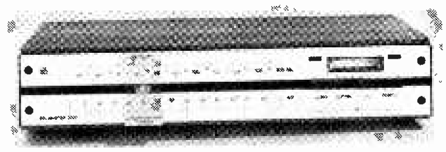
Sensitivity: 1.5μV (mono), 5μV (stereo) for 30dB s/n ratio at 75kHz deviation.
Limiting: Full limiting at 5μV.
I.F. bandwidth: 220kHz at 6dB.
I.F. rejection: 90dB.
Aerial inputs: 70-80Ω and 300Ω.
Output voltage: 300mV (average).
Output impedance: 1kΩ.
Price: £40 4s 6d.
When decoder is fitted, channel separation is 30dB at 1kHz, suppression at 38kHz is 40dB; automatic switching and stereo indicator; tuning meter and interstation noise suppression. Armstrong Audio Ltd, Warlters Road, London N.7.

Bush (a.m./f.m. stereo)



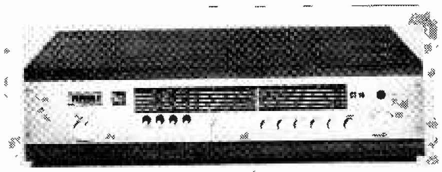
Sensitivity: 3μV for 26dB s/n ratio (mono).
Aerial input: not quoted.
Output voltage: 2V (max).
Output impedance: 1MΩ.
Power supply: external from Bush stereo amplifier via 5-pin DIN type socket.
Price: £36 15s.
Tuner intended primarily for use with Bush stereo amplifier. Features a.f.c. on f.m. with switch; stereo indicator and tuning meter. Above details refer to f.m. section. Coverage on a.m. is long- and medium-wave bands with internal ferrite rod aerial. Rank Bush Murphy Ltd, Power Road, Chiswick, London W.4.

Beomaster 5000 (f.m. stereo)



Sensitivity: 0.8μV for 20dB s/n ratio, 75kHz deviation.
I.F. bandwidth: 200kHz at 3dB.
Detector bandwidth: 1MHz.
Crosstalk: 40dB at 1kHz.
Aerial input: 75-300Ω.
Output voltage: adjustable between 55mV and 1V.
Pilot and carrier suppression: 50dB.
Price: £105 (teak), £106 (rosewood).
Ceramic filters and f.e.t. r.f. amp. and mixer incorporated; a.f.c.; automatic stereo switching with level adjustable between 1 and 100μV; interstation muting. Bang & Olufsen U.K. Ltd, Eastbrook Road, Gloucester GL4 7DE.

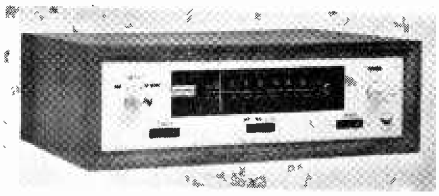
Dual CT16 (a.m./f.m. stereo)



Sensitivity: 1.5μV for 26dB quieting at 22.5kHz deviation (mono).
Image rejection: 50dB.
I.F. bandwidth: 200kHz.
Channel separation: better than 45dB.
Pilot tone suppression: better than 50dB.
Aerial input: 240Ω.
Output voltage: 0.8V for 30% modulation.
Output impedance: 47kΩ.
Price: £106.
Automatic stereo switching and stereo indicator; audio response 40-15000Hz ±1.5dB. Above details refer to f.m. section. On a.m. frequency coverage is 150-350kHz (l.w.), 500-1650kHz (m.w.), 6.7-15.4MHz and 5.6-6.6MHz. Dual Electronic Industries, St. Georgen House, Mill Road, Stockenchurch, Bucks.

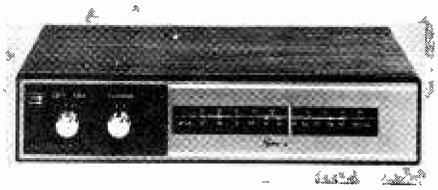
Eagle AFT60 (a.m./f.m. stereo)

Sensitivity: 2μV for 20dB quieting (stereo).
Channel separation: 28dB at 1kHz.
Image rejection: 55dB.
Aerial input: 75Ω unbalanced.
Output voltage: variable 0-280mV.
Price: £53 18s 10d.
Above details refer to f.m. section. Cover-



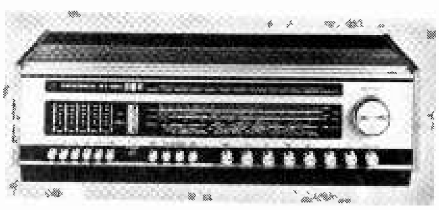
age on a.m. 600-1600kHz (m.w.) with sensitivity of 500μV. Teak cabinet; f.e.t. front end. U.K. agents: B. Adler & Sons (Radio) Ltd, Coptic Street, London W.C.1.

Elizabethan System 4
(f.m. mono)



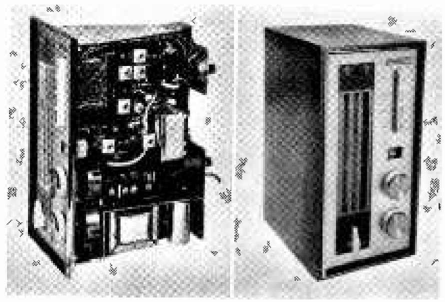
Sensitivity: 14μV for 3dB from limiting.
Output voltage: 100mV.
Price: £28
A.f.c. ±250kHz range. Part of the Elizabethan Coniston Suite. Lee Products (Great Britain) Ltd, 10-18 Clifton Street, London E.C.2.

Grundig RT100 (a.m./f.m. stereo)



Sensitivity: 1.5μV for 26dB quieting at 15kHz deviation.
Image rejection: 58-66dB.
I.F. bandwidth: 160-200kHz.
Discriminator bandwidth: 650kHz.
A.M. suppression: better than 58dB at 1kHz.
Pilot tone suppression: 40dB at 19kHz, 60dB at 38kHz.
Channel separation: 35dB minimum at 1kHz.
Aerial input: 240Ω.
Output voltage: 0.65V for 40kHz deviation.
Output impedance: 2kΩ (lowest permissible load impedance 22kΩ).
Price: £181 11s 9d.
Automatic stereo switching with variable trigger level (6-60μV); override mono switch and indicator; f.e.t. front end with a.f.c.; tuning meter and light aids; variable-capacitance diode tuning with press-button selection of up to six stations. Above details refer to f.m. section. Frequency coverage on a.m.: 145-350kHz (l.w.), 510-620kHz (m.w.), 3.15-8.8 and 8.6-22.5MHz. Grundig (Great Britain) Ltd, 15 Orchard Street, London W1H 9AE.

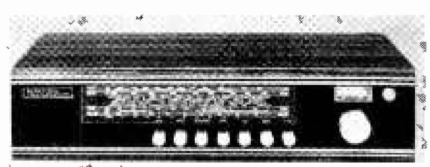
Goodmans Stereomax
(a.m./f.m. stereo)



Sensitivity: 2μV for 30dB quieting.
I.F. bandwidth: 300kHz at 6dB.
Discriminator bandwidth: 600kHz.
A.M. rejection: 40dB.
Channel separation: 37dB at 1kHz.
Pilot tone suppression: 50dB.
Aerial input: 300Ω balanced, 70Ω unbalanced.
Output voltage: 250mV for 30% modulation.
Output impedance: 100kΩ.
Automatic stereo switching and stereo indicator; switched a.f.c. (pull-in range 100kHz); muting; tuning meter. Above details refer to f.m. section. Frequency coverage on a.m. 545-1650kHz (m.w.). External power socket for feeding other units. Goodmans Loudspeakers Ltd, Axiom Works, Lancelot Road, Wembley, Middx.

Heathkit K/AJ-14 (f.m. stereo)
Sensitivity: 5μV.
Image rejection: 45dB.
I.F. rejection: 80dB.
Frequency response: 20-15000Hz at 3dB (I.H.F.).
Aerial input: 300Ω balanced, 75Ω unbalanced.
Output voltage: 0.7V for 1000μV input.
Output impedance: 20kΩ.
Channel separation: 30dB at 1kHz.
Price of kit: £24 18s, post and packing 5s.
Mono/stereo switch and stereo indicator; a.f.c.; walnut or teak cabinet available. Heath (Gloucester) Ltd, Gloucester GL2 6EE.

Körting T500 (a.m./f.m. stereo)

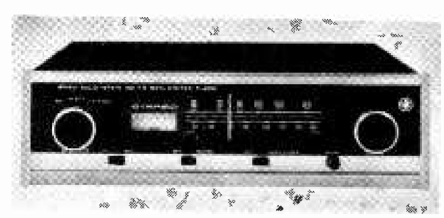


Sensitivity: 3μV for 26dB s/n ratio at 12.5kHz deviation.
Bandwidth: approximately 140kHz.
Price: £79 15s.
Automatic stereo switching with indicator; tuning meter; frequency coverage on a.m. 145-355kHz (l.w.), 510-1620kHz (m.w.) and 5.85-7.4MHz. Special feature on a.m., automatic bandwidth (3-6kHz) dependent on input signal strength. U.K. agents: Europa Electronics Ltd, Howard Place, Shelton, Stoke-on-Trent ST1 4NW.

Leak Stereofetic (f.m. stereo)
Sensitivity: 2.5μV for 30dB s/n ratio.
Image rejection: 72dB.
I.F. rejection: 80dB.
A.M. suppression: 50dB.
Aerial input: 75Ω unbalanced.
Output voltage: 0.9V r.m.s. at 75kHz deviation.
Output impedance: 200Ω.
Channel separation: better than 35dB at 1kHz.
Price: £68 1s 4d.
Switched a.f.c.; tuning indicator combined stereo indicator; f.e.t. front end and i.cs in i.f. stages and decoder; ceramic filter i.f. stages; automatic stereo switching with "mono lock" switch. Features "quasi-stereo", a method of progressively reducing noise while retaining some stereo effect. H. J. Leak & Co. Ltd, Bradford Road, Idle, Bradford, Yorks.

L & H "Signalmaster" 6087
(f.m., decoder optional)
Sensitivity: 2.5μV for 20dB s/n ratio.
Image rejection: better than 55dB.
I.F. bandwidth: 250kHz.
Output voltage: 200mV (nominal).
Power supply: 9V battery or 32V external.
Price: 30gn.
With decoder channel separation is better than 32dB at 1kHz; a.f.c. switch; stereo indicator; available in teak or rosewood. Britimpex Ltd, 8/12 Rickett Street, London S.W.6.

Nikko FAM-12F (a.m./f.m. stereo)



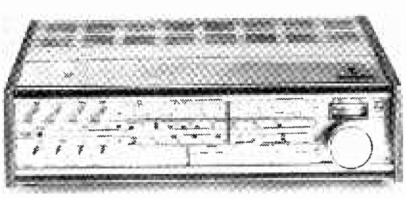
Sensitivity: 1.8μV for 20dB quieting.
Image rejection: better than 45dB at 1MHz.
Channel separation: 40dB at 1kHz.
Output voltage: 0.5V at 30% modulation.
Price: £68 8s 3d.
Automatic stereo switching with indicator and mono override switch; tuning meter; switched a.f.c.; muting; noise filter; f.e.t.

front end. Above details refer to f.m. section. Frequency coverage on a.m.: 530-1605kHz (m.w.). U.K. agents: Howland-West Ltd, 2 Park End, South Hill Park, London N.W.3.

Pioneer TX-900 (a.m./f.m. stereo)

Sensitivity: 1.7μV (I.H.F.).
Image rejection: 95dB.
Selectivity: 65dB.
S/N ratio: 60dB at 30% modulation.
Harmonic distortion: 0.3% at 400Hz, 100% modulation.
Aerial input: 300Ω balanced.
Output voltage: 1V at 400Hz, 30% modulation adjustable on each channel.
Channel separation: 38dB at 1kHz.
Price: £154 18s 3d.
Front end f.e.t; crystal filter i.f. stages; automatic stereo switching with mono override switch; adjustable stereo muting and interstation muting; switched a.f.c. Above details refer to f.m. section. Frequency coverage on a.m.: 525-1605kHz (m.w.). Shriro (U.K.) Ltd, Lynwood House, 24/32 Kilburn High Road, London N.W.6.

Philips RH691 (a.m./f.m. stereo)

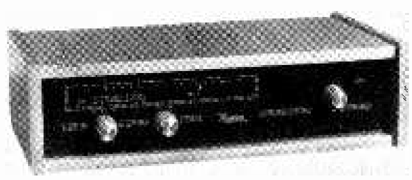


Sensitivity: 7μV for 26dB s/n ratio at 15kHz deviation.
Distortion: less than 1% at 75kHz deviation.
Channel separation: 35dB at 1kHz.
Carrier suppression: 40dB at 38kHz (30dB at 19kHz).
Audio response: 20-15000Hz ± 3dB.
Output voltage: 1.4V at 40kHz deviation.
Output impedance: 10kΩ.
Price: £83.
Switched a.f.c; automatic stereo switching with mono override switch and indicator; tuning meter; interstation muting. Above details refer to f.m. section. Frequency coverage on a.m.: 150-400kHz (l.w.), 525-1604kHz (m.w.) and 5.9-18.2MHz. Philips Electrical Ltd, Century House, Shaftesbury Avenue, London W.C.2.

Quad FM (stereo)

Sensitivity: 2μV for 30dB quieting.
Crosstalk: better than 30dB at 1kHz.
Suppression: better than 40dB at 38kHz.
Aerial input: 75Ω.
Output voltage: 100mV.
Output impedance: 100kΩ each channel (50kΩ mono).
Price: £51.
Automatic switching and stereo indicator with override manual switch; neon tuning indicator; a.f.c; fully tropicalized. Acoustical Manufacturing Co. Ltd, Huntingdon, Hunts.

Radon 404 Mk2
(f.m., decoder optional).



Sensitivity: 15μV (usable).
Bandwidth: 300kHz at 3dB.
I.F. bandwidth: 170kHz.
A.M. rejection: 26dB for 20mV input.
Aerial input: 75Ω.
Price: £39 18s with decoder fitted.
Switched a.f.c; permeability tuning. When decoder is fitted stereo switching is automatic with indicator. A Mark 3 version of this tuner has the stereo decoder fitted as standard. Radon Industrial Electronics Co. Ltd, Brooklands Trading Estate, Orme Road, Worthing, Sussex.

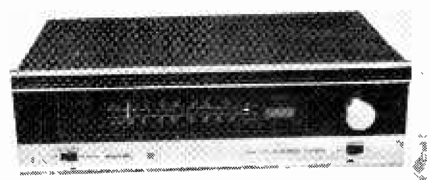
Radford FMT.3
(f.m., decoder optional)

Sensitivity: 2μV for 30dB quieting.
S/N ratio: 70dB.
Frequency response: 40-15000Hz ± 1dB.
Harmonic distortion: less than 0.6% at 75kHz deviation.
Detector bandwidth: 550kHz.
Aerial input: 75Ω and 300Ω.
A.M. suppression: 45dB.
Output voltage: 2V at 75kHz deviation.
Price: £60 plus p.t. (decoder fitted).
With decoder, channel separation is better than 37dB at 1kHz; carrier suppression better than 60dB; automatic stereo switching with mono override switch; switched a.f.c; tuning meter; interstation muting; f.e.t. r.f. amplifier. Radford Audio Ltd, Ashton Vale Road, Bristol BS3 2HZ.

Rogers Ravensbrook
(f.m. stereo)

Sensitivity: 5μV (usable).
I.F. rejection: 85dB.
Image rejection: 70dB.
A.M. suppression: 50dB.
Channel separation: better than 30dB at 1kHz.
Output voltage: 200mV for 30% modulation.
Price: £45 plus case.
Front end f.e.t; integrated circuit i.f. strip; switched a.f.c; automatic stereo switching with indicator; tuning meter; interstation muting. Rogers Developments (Electronics) Ltd, 4-14 Barmeston Road, Catford, London S.E.6.

Rotel 120-ST (a.m./f.m. stereo)



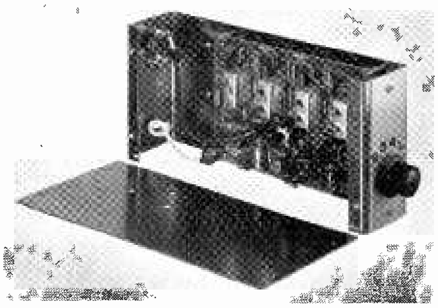
Sensitivity: 2.5μV for 20dB quieting.
Harmonic distortion: less than 1.5% at 1kHz
Channel separation: better than 35dB at 1kHz 100% modulation.
Output voltage: 1200mV at 100% modulation.
Price: £49 10s.
Automatic stereo switching with stereo indicator; a.f.c; tuning meter. Above details refer to f.m. Frequency coverage on a.m.: 535-1605kHz (m.w.). Distributors: Pullin Photographic Rank Aldis, P.O. Box 70, Great West Road, Brentford, Middx.

Sony ST-5100
(a.m./f.m. stereo)

Sensitivity: 1.8μV for 20dB s/n ratio.
Selectivity: 70dB (I.H.F.).

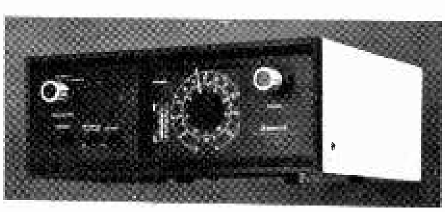
Image rejection: 90dB.
Spurious rejection: 100dB.
A.M. suppression: 50dB (I.H.F.).
Frequency response: 20Hz-15kHz ±1dB.
Aerial input: 300Ω balanced, 75Ω unbalanced.
Channel separation: better than 38dB at 400Hz.
Carrier suppression: 60dB.
Price: £99 15s.
Front end f.e.t.; i.f. filters; tuning indicators; a.f.c. control; muting switch. Above details refer to f.m. section. Frequency coverage on a.m.: 530-1605kHz (m.w.). Sony (U.K.) Ltd, 11 Ascot Road, Bedfont, Feltham, Middx.

S.N.S. FMT/9 (f.m. mono)



Sensitivity: 10μV for 30dB s/n ratio at 22.5kHz deviation.
Limiting: 10μV for full limiting.
Aerial input: 70-80Ω unbalanced.
Output voltage: variable 0-500mV.
Output impedance: 5kΩ.
Power supply: 9.1V d.c. or optional a.c. mains unit.
Price: on application.
This tuner is available either for single programme reception or in switched version for up to four preset stations. It has a crystal-controlled oscillator and in the four-station version the above sensitivity figures are slightly impaired. Designed specially for radio distribution systems. S.N.S. Communications Ltd, 851 Ringwood Road, West Howe, Bournemouth BH11 8LN.

Sansui TU555 (a.m./f.m. stereo)



Sensitivity: 2μV for 20dB quieting.
Image rejection: better than 50dB.
I.F. rejection: better than 60dB.
Channel separation: better than 35dB.
Aerial input: 300Ω balanced, 75Ω unbalanced.
Output voltage: more than 1.5V.
Output impedance: greater than 10kΩ.
Price: £77 9s 2d.
Signal strength meter and stereo indicator; stabilized oscillator supply with no a.f.c; special noise cancelling circuit; f.e.t. front end. Above details refer to f.m. section. Frequency coverage on a.m.: 535-1605kHz (m.w.). U.K. importers: Brush Clevite Co. Ltd, Thornhill, Southampton SO9 1QX, Hants.

Sugden R21 (f.m. stereo)

Sensitivity: 2μV for 30dB quieting.
Frequency response: 30Hz-15kHz ±1dB.
I.F. rejection: 80dB.
Image rejection: 70dB.
A.M. suppression: 50dB.
Aerial input: 75Ω or 300Ω.
Output voltage: 0.5V.
Channel separation: better than 30dB.
Price: approximately £67.
Front end f.e.t.; variable-capacitance diode tuning with four preset station buttons in addition to continuous tuning; switchable stereo filter giving low noise with reduced separation; special low-pass filter with 55kHz cut-off; pilot tone and sub-carrier filters. Differently styled version (R51) available. J. E. Sugden & Co. Ltd, Bradford Road, Cleckheaton, Yorks.

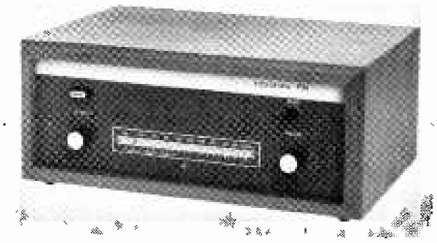
Sinclair System 3000

This f.m. tuner will be released in late autumn. The only information available at the time of going to press is that sensitivity is expected to be 5-10μV for 40dB quieting. The circuit will incorporate a phase-lock loop discriminator, variable-capacitance diode tuning and an i.c. stereo decoder. Sinclair Radionics Ltd, 22 Newmarket Road, Cambridge CB5 8DU.

Teleton STQ-201X (f.m. stereo)

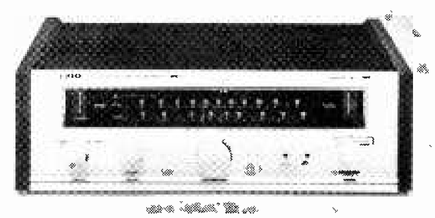
Sensitivity: 2.5μV for 20dB quieting.
I.F. bandwidth: 240kHz (3dB).
Image rejection: better than 50dB.
S/N ratio: 45dB for 1mV input.
Aerial input: 300Ω balanced.
Channel separation: 30dB at 1kHz.
Price: £36.
Stereo indicator; a.f.c.; tuning meter. Teleton Electro (U.K.) Co. Ltd, Teleton House, Robjohns Road, Widford, Chelmsford, Essex.

Tripletone FM Mk2 (f.m. stereo)



Sensitivity: 2-3μV for 20dB quieting.
Bandwidth: 210kHz.
Aerial input: 70-80Ω unbalanced.
Output voltage: 0.1V.
Channel separation: better than 30dB at 1kHz.
Price: £37 19s 10d (teak case), £35 1s 3d (chassis only).
Front end f.e.t.; a.f.c. (range 400kHz); automatic stereo switching and indicator. A mono version is available with provision for a plug-in decoder. The Tripletone Manufacturing Co. Ltd, 138 Kingston Road, Wimbledon, London S.W.19.

Trio KT-7000 (a.m./f.m. stereo)



Sensitivity: 1.5μV (usable).
Frequency response: 20-15000Hz −2dB.
Harmonic distortion: less than 0.3% (mono) at 400Hz, 100% modulation.
S/N ratio: better than 70dB at 30μV input.
Image rejection: better than 100dB at 100MHz.
I.F. rejection: better than 100dB.
Channel separation: better than 35dB at 1kHz.
Carrier suppression: better than 50dB.
Aerial input: 300Ω balanced, 75Ω unbalanced.
Output voltage: 1.5V at 400Hz, 100% modulation.
Output impedance: 7000Ω.
Price: £125.
Three-f.e.ts front end; crystal filter and integrated circuit i.f. stages; signal strength meter and tuning meter; automatic stereo switching and indicator; interstation muting; step-type output level control. Above details refer to f.m. section. Frequency coverage on a.m.: 540-1600kHz (m.w.). U.K. agents: Lasky's Radio Ltd, 3-15 Cavell Street, Tower Hamlets, London E.1.

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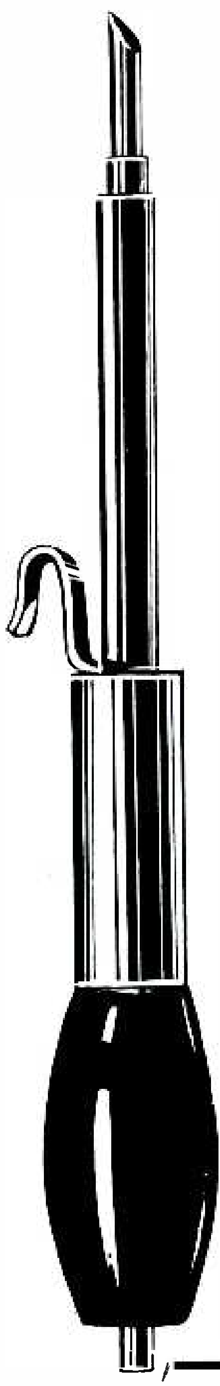

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