Wireless World

Editor:
H. W. BARNARD

Technical Editor:
T. E. IVALL, M.I.E.R.E.

Deputy Editor:
B. S. CRANK

Assistant Editor:
G. B. SHORTER, B.Sc.

Drawing Office:
L. DARRAH

Production:
D. R. BRAY

Advertisements:
G. BENTON ROWELL *(Manager)*
Phone 01-261 8339
G. J. STICHBURY
Phone 01-261 8037
A. PETTERS *(Classified Advertisements)*
Phone 01-261 8508 or 01-928 4597

Electronics—Education or Vocational Training?

The fact that electronics has found a foothold in the academic world, and people are coming out of universities and colleges with qualifications in the subject, does not mean that all is now settled. We can't just sit back and serenely watch it being taught year after year as a neat compartment of knowledge alongside mathematics, English, geography, economics and all the rest. There is a great deal of confusion in people's minds, both inside and outside the educational world, about the role of electronics as an academic subject and how it should be taught. "Is the subject suitable for education? Is it an exercise of the mind? Is it better left till education is completed? Is it not sufficiently attractive to ensure a voluntary attention to it? Is it a convenient subject for examination?" These questions were not, in fact, asked about electronics, but were written in an anonymous pamphlet attacking the introduction of history as an academic subject at Oxford University in the middle of the 19th century. If such a subject as history, with all the respectability of apparent uselessness, could be criticized so recently, it's no wonder that the 20th century phenomenon of electronics, still erupting new ideas and techniques, should be questioned in similar terms.

It is, in fact, because electronics is still developing that the confusion prevails in people's minds. First there is the circumstance that, being an industry, it needs recruits and so has instituted various schemes of training, with the help of educational organizations. Vocational training is not education — that is, in purpose — and this has led some people to believe that the subject of the vocational training cannot also be a subject for education. (History in the early 19th century was considered mainly as vocational training for soldiers, statesmen and lawyers.) The confusion is added to by managers in the industry who seem to expect the universities, in particular, to produce fully trained recruits for them, and are slightly offended when they discover that electronics graduates are not what they had in mind. So often has one heard the demand "What we want from the universities is . . . ", as if industry's requirements were the arbiter in education. The cynical tag that physics is useless electronics is not always intended as a joke.

A further difficulty resulting from the rapid growth of electronics is how to categorize the subject within the existing framework of human knowledge. Is it a part of physics (e.g. physical electronics) or is it a branch of electrical engineering, just as electrical engineering was at one time classified as a branch of civil engineering? Southampton University responded to this question boldly by establishing the first autonomous department of electronics, but most others have cautiously introduced the subject under such titles as Department of Electrical and Electronic Engineering. A new, though perhaps more fundamental, categorization is to see electronics as reaching into numerous fields of human activity as a technology of information processing, in communication between man and man, man and machine, and between machine and machine. Professor Gambling of Southampton University suggests that 'information engineering' would be an appropriate name for this.

One thing, at least, is clear. Electronics is not yet acceptable as an A-level subject in the schools. After considering an experimental period of teaching electronics at the Colchester Royal Grammar School — a scheme initiated and supervised by Professor Chaplin and some of his colleagues at nearby Essex University — the Schools Council for the Curriculum and Examinations has decided that the subject is 'too narrow'. It's a pity the Council is not willing to explain the criteria and methods of comparison by which this decision was reached.

Publication Date
We regret that, as a result of production problems, this issue is a fortnight late appearing. In addition we expect that the June issue will be published about one week late.

Trace Quadrupler for D.C. 'scopes

by D. Bollen

The quadrupler unit can be added to a single-beam scope to give four independent Y traces, or a pair of XY traces, without sacrificing sensitivity or d.c. coupling. All traces can be positioned anywhere on the screen without interaction. The principle of the quadrupler is to sample four inputs, by means of linear transmission gates switched by a JK flip-flop ring-counter. Inputs are sampled one at a time (quad mode) or two at a time (dual-pair mode), at switching frequencies extending to 2MHz.

Fig. 1 shows two methods of sampling four Y inputs one at a time. In Fig. 1(a), a free-running oscillator triggers the ring-counter and causes each gate to open in turn. Provided that the sampling frequency is not harmonically related to the timebase frequency, the display will appear to be four continuous input waveforms. The quality of this chopped display depends on a clean switching waveform with fast rise and fall times and a minimum of under- or overshoot contributed by the 'scope. However, it is difficult to avoid some trace thickening at high 'scope sensitivities, and a screen glow due to the finite rise and fall times of the switching waveform at sampling frequencies in excess of 100kHz.

A much better display at fast sampling rates can be achieved if the traces appear alternately, triggered by timebase flyback, see Fig. 1(b). Here the unwanted transients occur only at the edges of the screen, so there is no trace thickening or glow. The limitations with timebase triggered sampling are break-up of the display when switching and propagation delays exceed flyback time, and a display flicker below 5ms/cm, with average persistence screen coatings, due to the natural time division of four.

The block diagram of the quadrupler is given in Fig. 2. Input signals are applied, via frequency compensated attenuators, to four identical non-inverting pre-amplifiers, each having a gain of two, an f.e.t. input, a low output impedance, and variable d.c. output shifting. The four-position switch S_1 allows a sync. signal to be taken from any pre-amp. output, prior to gating. Gates are opened by the ring counter in a sequence selected by S_2, in response to clock pulses derived from either a square-wave oscillator with six switch-selected chop frequencies, or from a timebase triggered Schmitt. The wide choice of chopping frequencies ensures freedom from harmonic locking with the timebase.

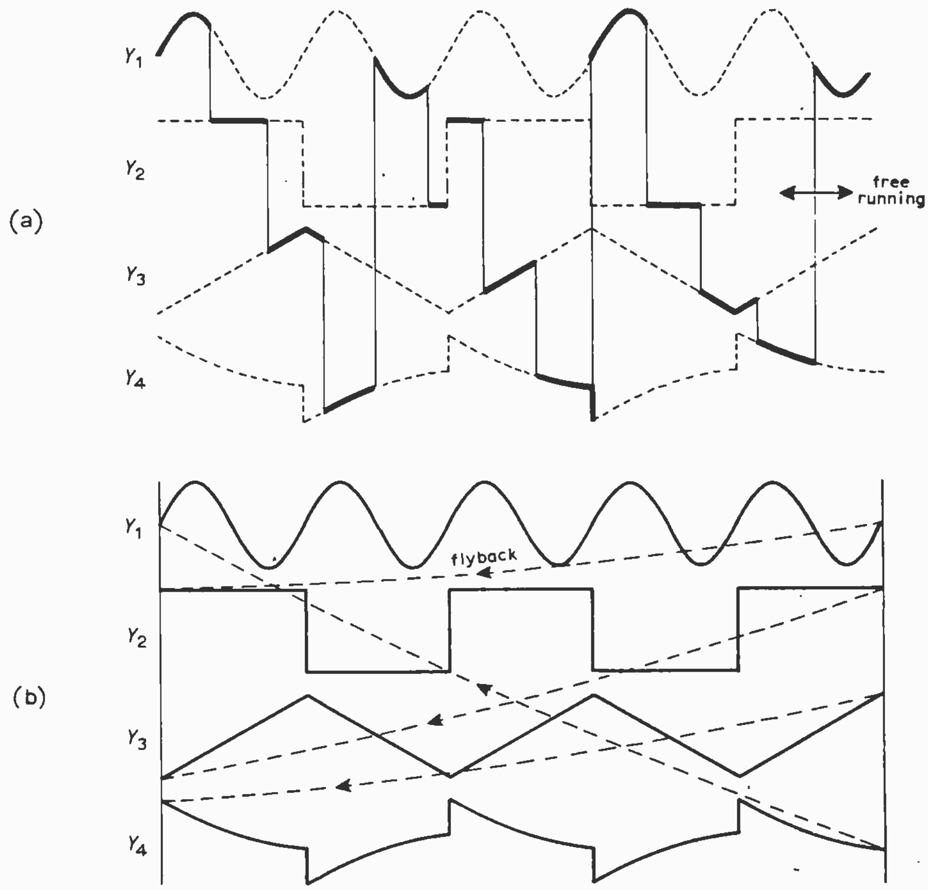
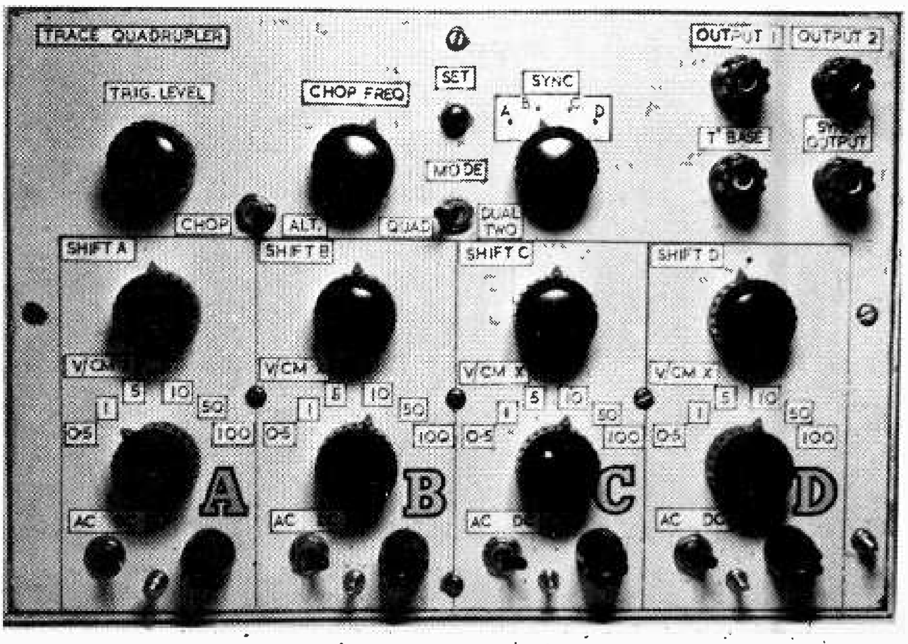


Fig. 1. How a four-trace display is built up, (a) in the chopped trace mode, and (b) in the alternate trace mode.



Gating

The basic circuit of the quadrupler four-diode gate is given in Fig. 3. When the gate pulse goes positive all four diodes conduct, and R_1 and R_2 are connected to the signal path. The input does not actually drive the gate output, but merely allows R_1 and R_2 to 'pull up' and 'pull down' the output terminal. Thus, when $R_1 = R_2 = R_3$ the maximum amplitude the gate will handle will be half the gate pulse amplitude, beyond which there will be clipping of signal peaks. With the gate open a signal passes from input to output with virtually no shift of d.c. level and very little attenuation.

All diodes are rendered non-conducting when the gate pulse goes negative, thus closing the gate. Each diode can then be represented by a small capacitor of about 3pF, in series-parallel with input, output, R_1 and R_2. This small value of capacitance accounts for the intermodulation between channels at high input frequencies, typically 10% at 8MHz. Fig. 4 gives the display sequences for quad and dual-pair operation.

Ring counter

The action of the ring counter in circuit Fig. 5 is to transfer a pulse, or a group of pulses, from left to right along outputs A to D when triggered by clock pulses.

A brief description of *JK* flip-flops will help to explain the chain of events in the ring-counter. Each *JK* element consists of inputs *J* and *K* feeding a master bistable, which in turn feeds a slave bistable having outputs Q and $\bar{Q}$. On the leading edge of a positive-going clock pulse, master and slave are first disconnected from each other, then information present at *J* and *K* terminals is entered into the master. As the clock pulse trailing edge arrives, *J* and *K* terminals are first isolated from the master then the previous information is transferred from master to slave, and appears at Q and $\bar{Q}$.

In the ring counter of Fig. 5, any *JK* element will assume the 'state' of the *JK* in front of it at the termination of a clock pulse. Meanwhile the preceding *JK* may well have changed state in response to the *JK* in front of it.

The action of preset and clear inputs is to override clock pulses and hold the slave bistable in a known condition. With preset earthed Q will be held on, and $\bar{Q}$ is held on by the clear input. In the quadrupler ring counter, clock pulses are continuous and the mode of operation is first selected by S_1 and then implemented by pressing S_2. If S_1 is placed in the quad position while the ring counter is running, all the outputs A to D

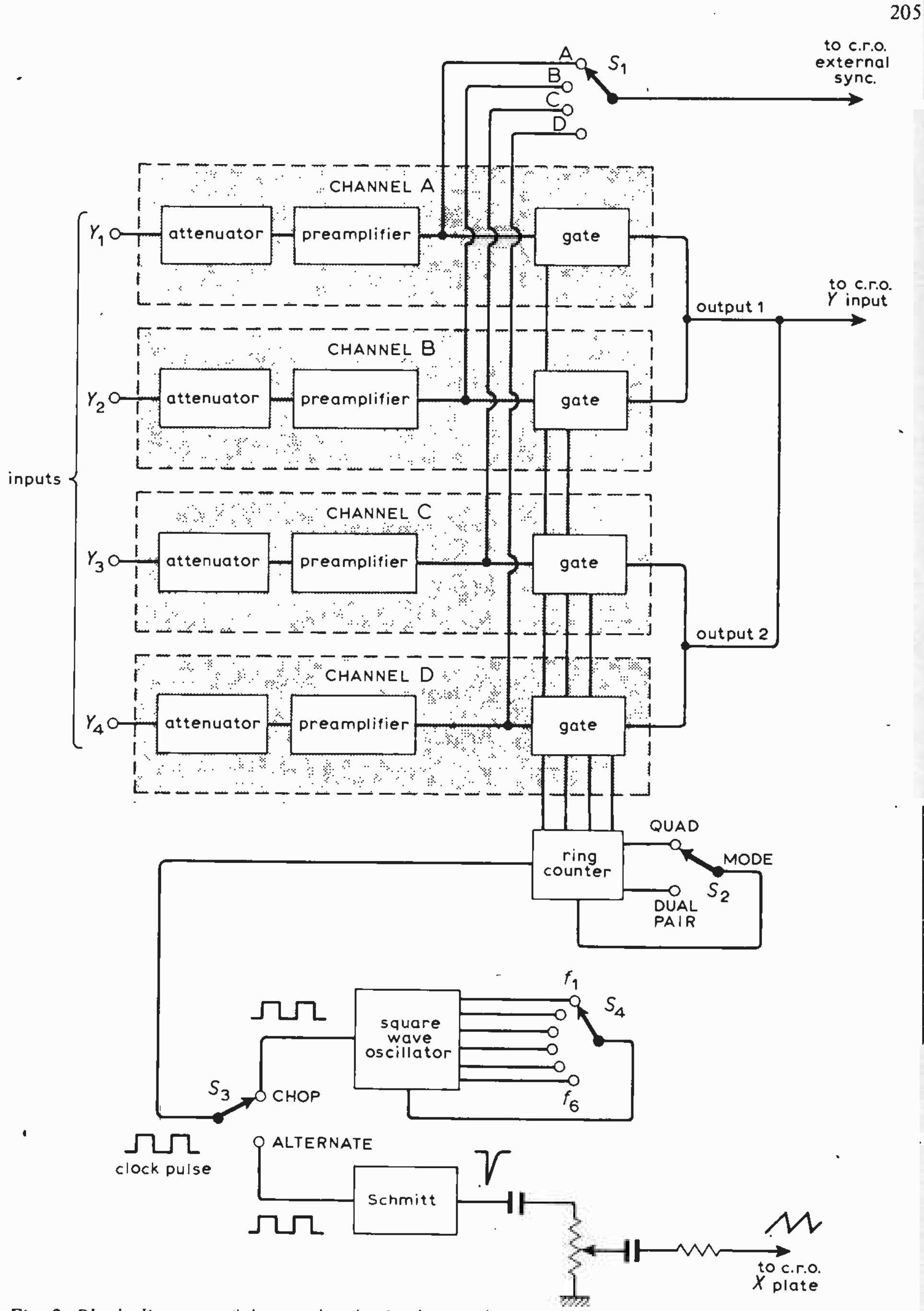


Fig. 2. Block diagram of the quadrupler in the quad mode.

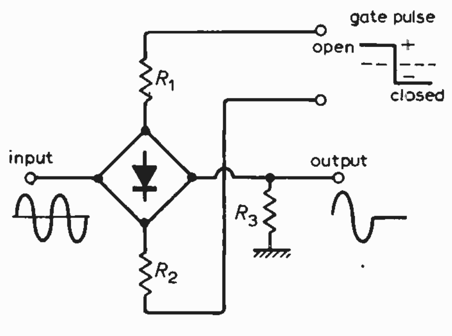
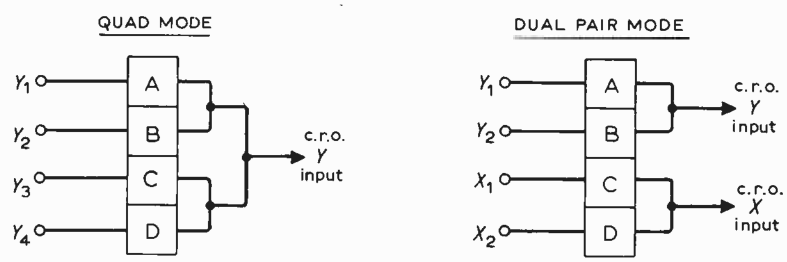


Fig. 3. Linear transmission gate. The input signal is controlled by the gate pulse.



sequence	gate open	display
1	A	Y_1
2	B	Y_2
3	C	Y_3
4	D	Y_4

sequence	gate open	display
1	A,C	Y_1, X_1
2	B,D	Y_2, X_2

Fig. 4. Gate switching sequences in the quad and dual-pair mode.

will freeze as soon as S_2 is pressed, with A 'on' (about 4V) and B, C, and D 'off' (about 0.2V). As S_2 is released, the first clock pulse transfers the 'on' condition from A to B, and thence to C and D on receipt of further clock pulses. If S_1 had been set to dual pair, pressing S_2 would have caused A and C to turn on, with B and D off. After releasing S_2, the first clock pulse would then shift 'on' from A and C to B and D, and back again with the second clock pulse.

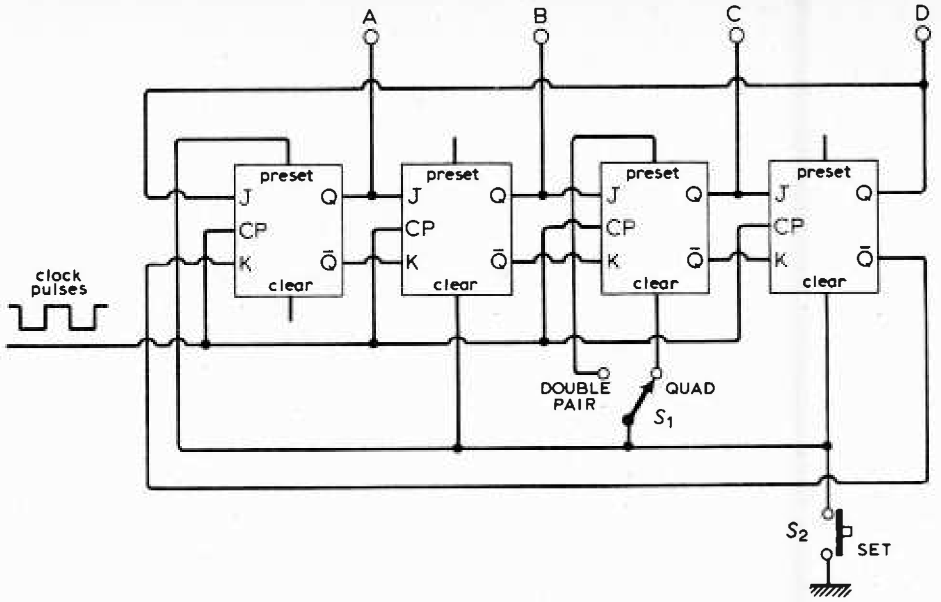


Fig. 5. The ring counter showing the set and reset arrangements.

Pre-amplifier

In the following circuit diagrams, quadrupled components are suffixed with the letters A, B, C or D, to correspond with the channel in which they are used. Components common to all channels have no suffix.

The pre-amplifier circuit of Fig. 6 has input attenuators designed around an impedance of 3MΩ shunted by 10pF, plus another 5pF from circuit strays. Maximum error is +3.3% −3% using 2% resistors selected on the basis of ±5% preferred value increments. In choosing attenuator division ratios consideration was given to achieving the widest possible range with only six steps, by omitting the 2 from a 1, 2, 5 sequence. With 'scope sensitivity set to 100mV/cm, each quadrupler input will, for example, offer 50mV, 100mV, 500mV, 1V, 5V, and 10V/cm.

The amplifier section of Fig. 6 uses parallel derived, series injected d.c. negative feedback, controlled by R_{39A}, to hold the gain down to two and reduce output impedance to less than 100 ohms. Capacitive shunting of the feedback path (C_{13} and C_{14} assisted by C_{15}) maintains gain up to 5MHz.

Ideally the pre-amplifier shift controls R_{40} should permit adjustment of Tr_2 collector voltage without influencing gain, but in practice there is some attenuation of signals with shift voltages greater than −4V, but this is apparent only with 'scopes having 10cm vertical scale at 1V/cm.

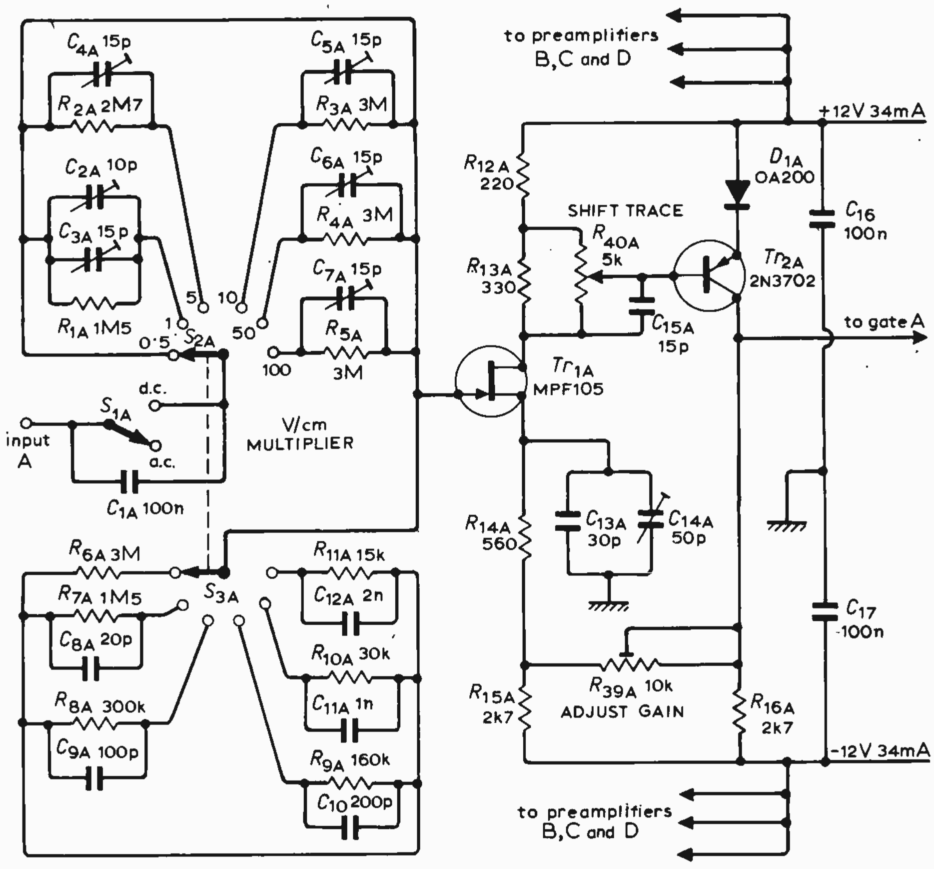


Fig. 6. One of the four input pre-amplifiers.

Quad-gate

It was noted earlier that gate pulse amplitude should exceed maximum signal amplitude if clipping is to be avoided. Also, a large gate pulse will tend to give cleaner switching. In the circuit of Fig. 7, the diode gates are switched by transistor pairs Tr_3 and Tr_4, with the base of each transistor driven from ring counter Q and $\bar{Q}$ outputs, thus achieving a gate pulse of more than twice the maximum signal amplitude.

As the need for fast gate opening and closing is mainly dictated by minimum timebase flyback time when working with alternate traces, 300ns for the quadrupler gates should be adequate with timebase sweeps of 50ns/cm or more. However, this gating time can be halved, if so desired, by omitting i.c. output protection resistors R_{25} and R_{26} from the circuit of Fig. 8.

Ring-counter and clock

In Fig. 8, flip-flop ground connections are taken to the −12V rail, instead of to earth, so that d.c. coupling can exist between ring counter outputs and the gate switching transistors, thus allowing the quadrupler to be used with long-persistence tubes at very slow sampling rates. Therefore, if resistors R_{25} and R_{26} are removed to improve gate switching times, accidental shorting to earth of ring counter outputs will almost certainly result in catastrophic failure of i.cs; a point to be borne in mind. Transistor Tr_5 in Fig. 8 supplies a roughly stabilized +5V to the ring counter, relative to the −12V rail.

If the quadrupler ring counter was susceptible to random noise pulses it would tend to assume an unwanted mode during use, and switch the wrong gate at the wrong time. The series 74N family demands a clock pulse with rise and fall times of less than 150ns for good noise immunity, and clock pulses should also originate from a low impedance source with a logic 0 of less than 0.4V and a logic 1 of more than 2.4V. The above conditions are satisfied by the clocking circuits of Fig. 8. Complementary emitter followers Tr_6 and Tr_7 ensure a low impedance, and provide a logic 0 and 1 of 0.2V and 4.5V respectively, when overdriven by the square-wave oscillator and Schmitt trigger. The 2N3702 used for Tr_7 saturates at around 0.2V.

The somewhat unusual square-wave

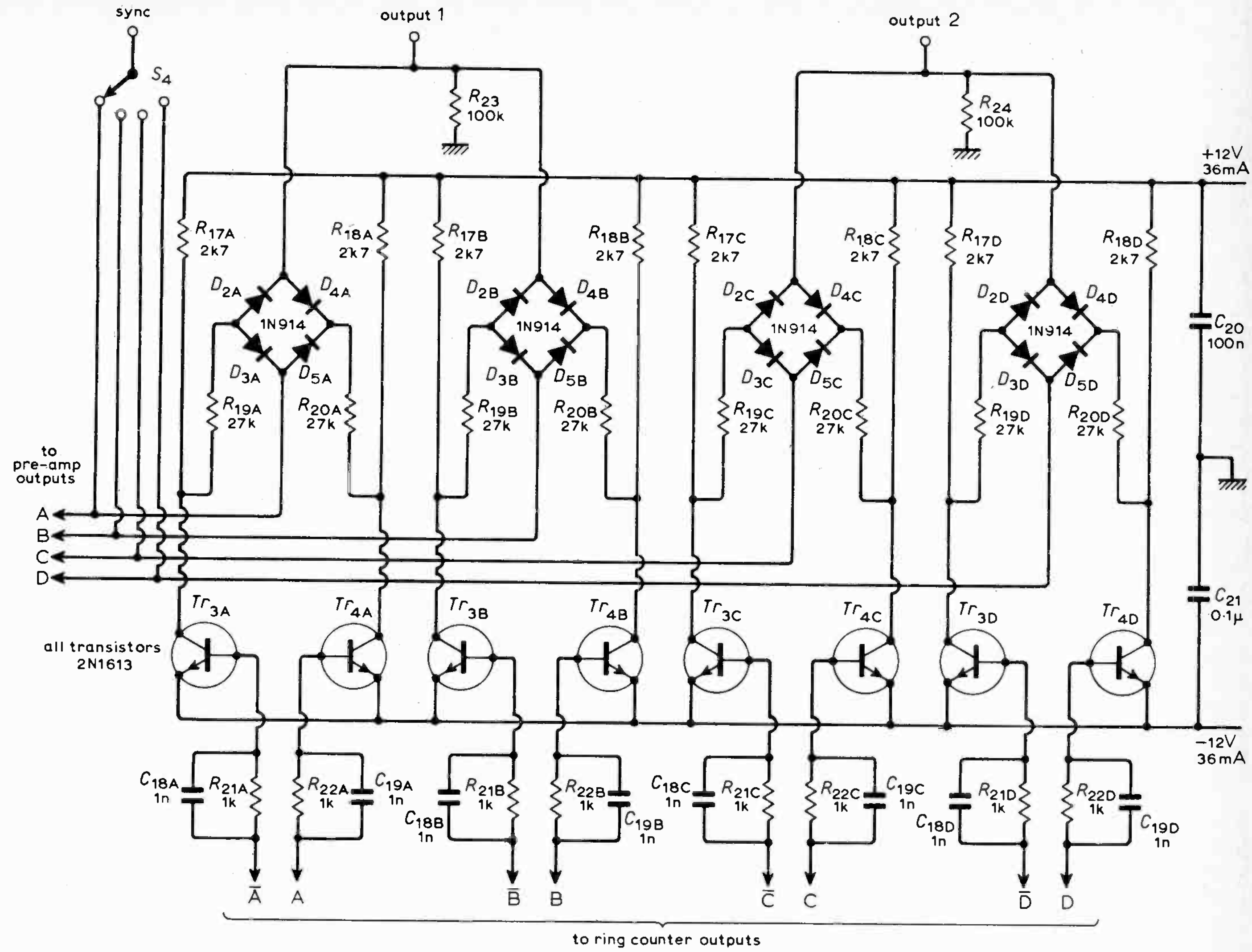


Fig. 7. The four linear transmission gates with the components for interfacing with the counter.

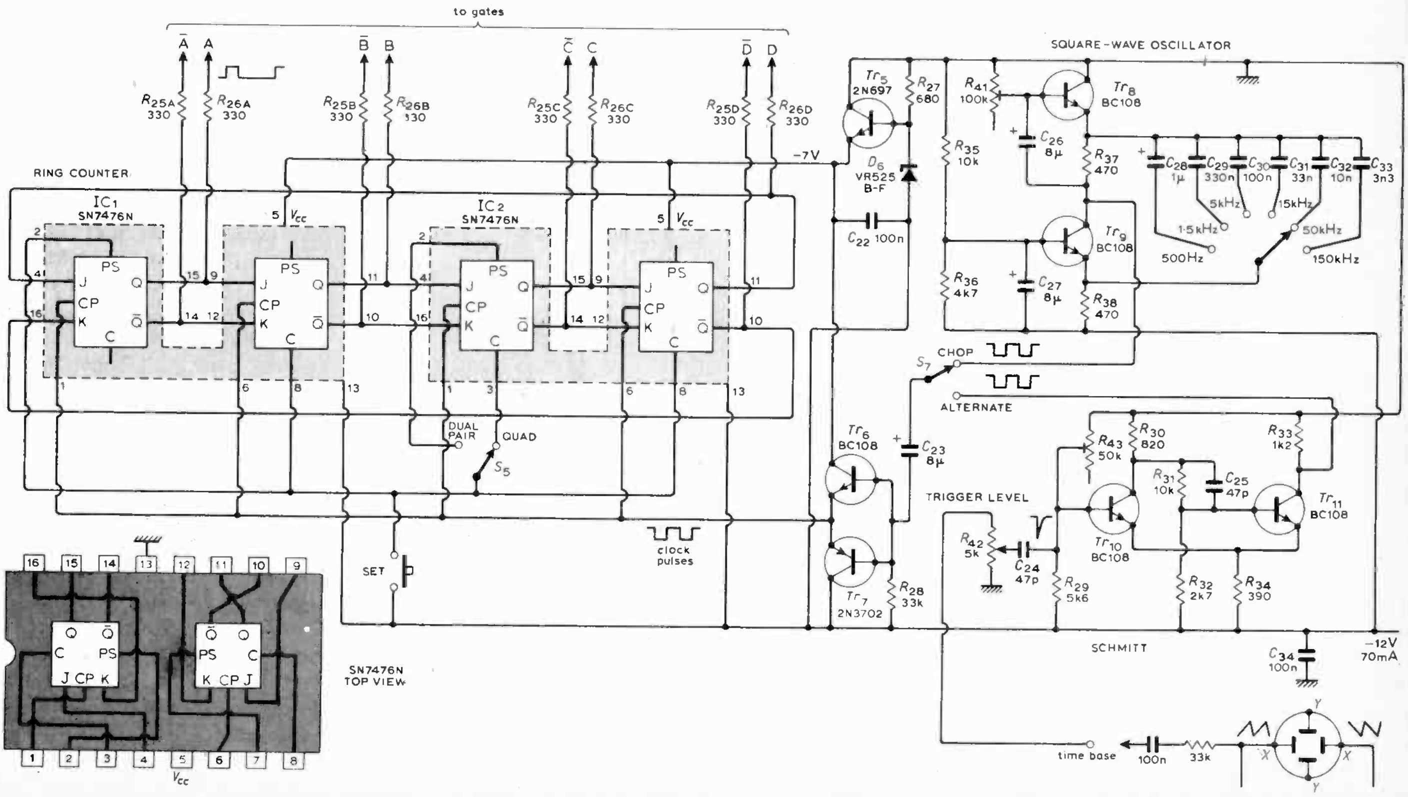
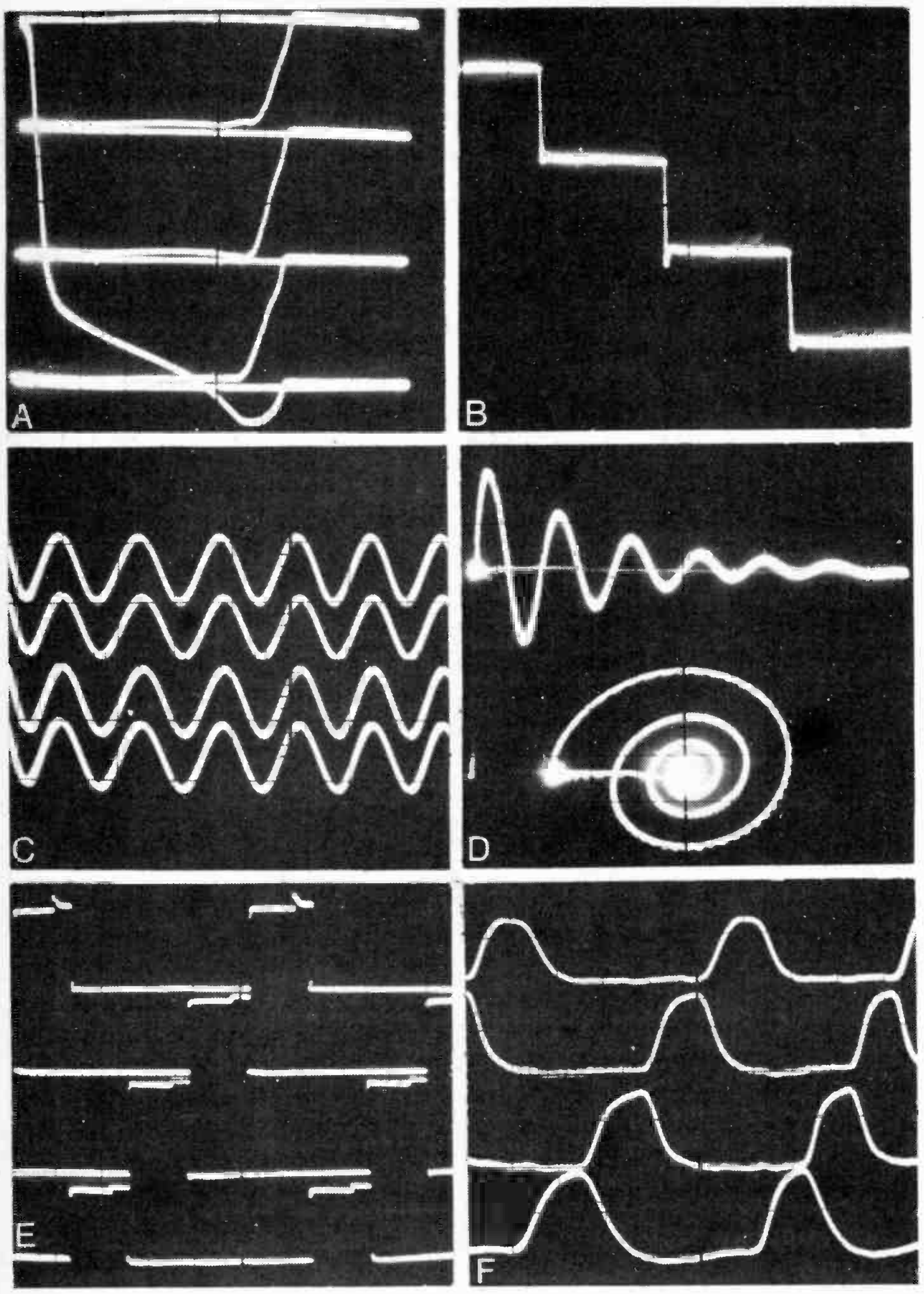


Fig. 8. Ring counter and clock pulse circuits.



(A) Alternate trace switching at a timebase rate of 100ns/cm with flyback blanking removed. (B) Chopping between channels at 150kHz; 'scope sensitivity 500mV/cm. (C) Showing the good phase relationship between identical 8MHz inputs after careful adjustment of the timebase controls, the chop frequency was only 500Hz. (D) Example of dual pair operation showing analogue computer simulations of a damped oscillatory system; the top trace is a decay curve, the bottom trace shows the limit cycle $Y = d^2X/dt^2$, $X = dx/dt$. (E) The b.c.d. outputs from an i.c. shift register in the quad mode; clock frequency = 10kHz, timebase 80μs/cm, chop frequency 500Hz. (F) the same shift register at a clock frequency of 8MHz, timebase 100ns/cm; although near the limit of the c.r.o's pulse response the quadrupler maintains a good relationship between pulse turn on and off times.

oscillator of Fig. 8 offers the following advantages over a conventional astable multivibrator; approximately equal rise and fall times of less than 150ns, and frequency selection by single capacitors. Potentiometer R_{41} establishes the mark-space ratio.

Looking next at the Schmitt, this is designed to operate beyond 2MHz, with capacitor C_{24} differentiating the timebase sawtooth to provide a steep negative-going pulse from the flyback edge. An output from the 'scope to the trigger control R_{42} is most conveniently taken from one of the X plates via a resistor of about 33kΩ in series with a 100nF isolating capacitor, but it might be preferable with transistor 'scopes to omit the resistor and tap off the sawtooth from the low-voltage timebase circuit.

Construction

Signal paths and ring counter output leads should be kept as short as possible, as should the connection from the timebase input socket to the Schmitt. It is, of course, essential to screen pre-amplifier inputs and attenuators against hum and pulse transients. The circuits of Figs. 6, 7, and 8 can be assembled on separate panels and positioned close to their controls and switches.

Capacitors C_{16}, C_{17}, C_{20}, C_{21}, C_{22}, and C_{34} are ceramic discs, and, apart from polyesters C_1 and C_{29}–C_{32} and electrolytics, all remaining fixed capacitors are polystyrene. Trimmer capacitors used in the attenuators and pre-amplifiers should be either ceramic or air-spaced, *not* mica compression.

All resistors, excluding the 2% metal oxide or high-stability used in the attenuators, are 10% 0.5W carbon. Pre-sets R_{39}, R_{41} and R_{43} can be sub-miniature skeleton types.

Power supply

The quadrupler should be run from a well stabilized 12–0–12V power supply capable of 200mA output. In the absence of such a supply, two 12V dry batteries will give tolerable results when shunted by 2500μF capacitors, for initial testing at reduced accuracy and increased drift.

Alignment

An audio, r.f., and square-wave generator will aid quick and accurate alignment of the quadrupler. Failing such test instruments simple oscillators can be made up in breadboard form to yield sine wave outputs of 1kHz and 5MHz, and a square wave of 1kHz; 5kΩ carbon potentiometers can be used as oscillator output attenuators, with the slider output shunted to earth by a 100Ω resistor in the case of the 5MHz oscillator.

First, check that the quadrupler functions correctly in the quad and dual pair modes with chopped and alternate traces. Ignore for the time being the amount of trace shift given by $R_{40A\text{-}D}$ as this is dependent on pre-amp. gain.

Connect the quadrupler to the oscilloscope as shown in Fig. 9. Inject a 1kHz sine wave signal into the 'scope and adjust controls for a display of 4cm peak to peak. Set the quadrupler attenuators to 1V/cm and switch the 'scope input to the quadrupler outputs. Adjust $R_{39A\text{-}D}$ for four identical waveforms of 4cm peak-to-peak. Superimpose the waveforms to check gain uniformity, and ensure that $R_{40A\text{-}D}$ will deflect their traces off the screen when the 'scope is set to 1V/cm.

Now set quadrupler attenuators to 0.5V/cm and inject a 5MHz signal from a low impedance source into the 'scope input. Adjust 'scope sensitivity for a display of 2cm peak-to-peak. Switch this same signal through the quadrupler and trim $C_{14A\text{-}D}$ for four identical waveforms of 4cm peak-to-peak. If the 'scope has insufficient bandwidth to display an undistorted 5MHz waveform of 4cm, either reduce signal amplitude or use a lower frequency.

The final test uses a square-wave generator to align attenuator capacitors for optimum pulse response. With quadrupler attenuators at 0.5V/cm apply a 1kHz square wave of about 1V to inputs A–D and adjust 'scope sensitivity for a display of about 4cm peak-to-peak. Assuming that the 'scope itself is correctly aligned, there should be virtually no over or undershoot, see Fig. 9. Now switch quadrupler attenuators to 1V/cm and increase signal amplitude to again give a display of 4cm peak-to-peak. Adjust $C_{3A\text{-}D}$ to obtain a correct square-wave response. The same process is then

Each channel	
input impedance	3MΩ and 15pF
frequency response	Flat d.c.-5MHz, −3dB at 8MHz
rise time	⩽ 10ns
intermodulation between channels	< 0.15% at 100kHz < 4.5% at 4MHz < 10% at 8MHz
maximum output	±5V into 100kΩ (quad mode) or into 50kΩ (dual pair mode)
trace shift	±6V
attenuator error	< ±3.3%
attenuator range	V/cm times 0.5, 1.5, 10, 50, 100
scope input drift	3mV/°C after warm-up
interchannel drift	0.5mV/°C
trace 'noise'	about 2mV, depending on 'scope and layout
Sampling	
chop frequencies	0.5, 1.5, 5, 15, 50, 150kHz
maximum alternate trace frequency	⩾ 2MHz, depending on flyback time

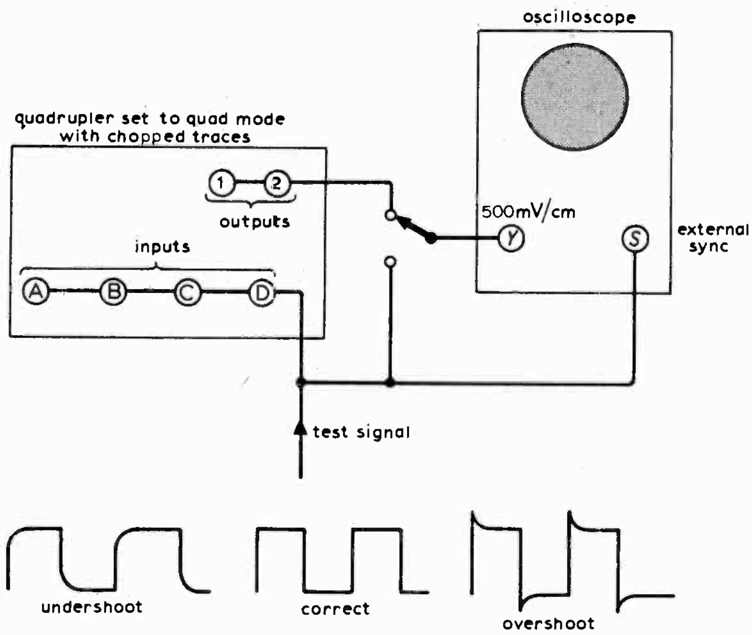


Fig. 9. Quadrupler alignment test circuit. Waveforms show the effect of attenuator trimming capacitor adjustment.

repeated for the remaining attenuator positions, by trimming appropriate capacitors ($C_{4A\text{-}D}$–$C_{7A\text{-}D}$). It will be necessary to increase 'scope sensitivity for an adequate display when the highest attenuator ranges are being trimmed.

The quadrupler in use

There is little to be gained in trying to operate the quadrupler with 'scope sensitivities of less than 50mV/cm as drift and trace noise will then prove troublesome. At the other end of the scale, severe waveform distortion can result if the 'scope is set higher than 1V/cm. With 'scope sensitivities confined to 100mV–1V/cm, quadrupler attenuators will still cater for inputs of from 50mV–100V/cm, and this should be more than adequate for most purposes.

Pre-amplifiers will begin to overload when input voltages exceed five times the V/cm attenuator setting and catastrophic failure of f.e.ts may result as voltages approach fifty times V/cm. Care must therefore be exercised in the choice of low attenuator settings when dealing with inputs of more than 20V amplitude.

With all electronic switch trace multipliers there can be some doubt that a correct phase relationship exists between independent input signals. Alternate trace working is more convenient to operate over a wide range of timebase speeds, and gives a better quality display, but it carries with it the penalty of 'phase slipping' at low audio frequencies and high r.f. Chopped trace working, on the other hand, ensures perfect phasing as long as the chop frequency exceeds that of the timebase, thus covering all signals up to a few hundred kilohertz.

Careful adjustment of timebase stability and trigger controls will usually minimize phase errors between high-frequency signals to within a few degrees. The technique is to first link quadrupler inputs together and inject a common signal, then the timebase is adjusted for a 'slip free' display while deriving a sync. signal from one of the pre-amplifier outputs.

Announcements

The Council of Industrial Design has changed its name to the Design Council. This follows the Department of Trade and Industry's request to the Council of Industrial Design that it should, in collaboration with the Council of Engineering Institutions, increase its activities in the field of engineering design.

The Electronics Division of the Institution of Electrical Engineers are organizing a residential vacation school on **M.O.S. Circuit Design**, to be held at the University of Edinburgh from 25-29 September. Further details can be obtained from the Divisional Secretary (Electronics), I.E.E., Savoy Place, London WC2R OBL.

The University of Essex have announced the receipt of Research Grants totalling £50,054. The North East Metropolitan Regional Hospital Board have extended the existing grant by £1,598 for a study of techniques based on direct patient-computer interaction for the assessment and treatment of communication disorders in neurological patients. A sum of £1,400 from the Post Office will finance a study on measurement of cross-polarization from the Sirio satellite. The Science Research Council have supplied £16,546 for an investigation into the generation and transmission of microsonic waves in solids.

Jameson Equipment (Trading) Ltd. have moved from London into premises at Abbey Manufacturing Estate, Woodside Place, Alperton, Wembley, Middx. HAO 1XA. Tel: 01-902 1114.

Thorn Radio Valves and Tubes Ltd and Thorn Colour Tubes Ltd have moved to Mollison Avenue, Brimsdown, Enfeld, Middx EN3 7NS. Tel: 01-804 1201.

Dale Electronics Ltd and Hamilton-Dale have moved from London to Dale House, Wharf Road, Frimley Green, Camberley, Surrey. Tel: Deepcut 5094.

Two contracts, worth £11M, for **military mobile radio equipment** have been announced by British Communications Corporation (a member of the Racal Group). The £6M order from the Ministry of Defence is for v.h.f./f.m. manpacks for the Clansman military communications project. The other contract is for radio equipment for use in Chieftain tanks of the Iranian Imperial Forces.

The Sound Powered Telephone Co. Ltd, of Tollesbury, Essex, has acquired from Pye-TMC the full manufacturing and marketing rights for the complete range of **sound powered telephones** for civil, marine and industrial use.

CIG International Capital Corporation have acquired the interests of the Top Rank Television Division of Rank Audio Visual Ltd. The acquisition has been made through a CIG subsidiary company Television Systems & Research Ltd.

Dana Laboratories Inc, Bilton Way, Dallow Road, Luton, Beds, have acquired control of E.I.P. Inc, of California, the company responsible for the first production y.i.g.-tuned frequency counter.

Aerial Facilities Ltd, of 20 Boston Place, London NW1 6HY, is a newly formed company which will provide a consultancy service on the siting of mobile communications aerials and who will also set up systems for mobile radio users using 'common aerial' working.

Martron Associates Ltd, 81 Station Road, Marlow, Bucks, has been appointed exclusive U.K. and European marketing outlet for the range of **digital frequency counters** manufactured by Stanley Laboratories, of Luton, Beds.

A contract valued at over £1M has been awarded to **Cossor Data Systems** by H.M.S.O. to supply nearly 500 computer terminals with associated equipment to the Royal Air Force for use with a new inventory management system.

Wireless World, May 1972

Linearity of the Transistor Amplifier

Effect of base-emitter junction

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

The relation between the base and collector currents of a bipolar transistor is substantially linear, whereas the relation between the base-emitter voltage and the collector current is exponential. Because of this, it is often said that the transistor is a current-operated device and should be driven at its base by a high-resistance source. This is, however, a practice which is becoming less and less adopted.

Most people now tend to think in terms of voltage drive and seem rarely to bother about the exponential voltage-current relation. Statements are sometimes made to the effect that this must cause severe waveform distortion, but in practice this does not seem to occur. The reason is that an emitter resistor is nowadays nearly always included for thermal stability and if this is not bypassed at signal frequencies, the resulting negative feedback linearizes the characteristic to a very high degree.

The question arises as to how much feedback is needed. The writer knows of no published information about this, but it is surprising if no one has in fact investigated the matter.

The essentials of a typical earthed-emitter circuit are shown in Fig. 1. The whole of the emitter current flows through the base-emitter junctions and the relation between the base-emitter voltage V_{BE} and the emitter current I_E is given by the basic diode equation

$$I_E = I_S[\exp.(KV_{BE})-1] \qquad (1)$$

where I_S is the reverse saturation current. This relation usually holds well over about three decades of current.

The practical difficulty about using equation (1) is that I_S is rarely known. Of course, if I_E is known for a given value of V_{BE}, I_S can be calculated. This can be rather troublesome because KV_{BE} is large and I_S is very small indeed. It is, however, unnecessary. If we let I_{OE} be the current corresponding to V_{OBE}, then

$$\begin{aligned} I_E &= I_{OE}+\Delta I_E \\ &= I_S[\exp.\ K(V_{OBE}+\Delta V_{BE})-1] \\ &= I_S \exp.\ KV_{OBE} \exp.\ K\Delta V_{BE} \\ &= I_{OE} \exp.\ K\Delta V_{BE} \end{aligned}$$

since the unity term is negligibly small.

We have therefore,

$$\Delta V_{BE} = \frac{1}{K}\log_e\left(1+\frac{\Delta I_E}{I_{OE}}\right)$$

$$= 0.06\log_{10}\left(1+\frac{\Delta I_E}{I_{OE}}\right) \qquad (2)$$

since $1/K = 0.026$V at room temperature and the conversion factor from Naperian to common logarithms is 2.3.

Now the circuit equation is

$$V_B = V_{BE}+I_E R_E \qquad (3)$$

$$V_{OB}+\Delta V_B = 0.06\log_{10}\left(1+\frac{\Delta I_E}{I_{OE}}\right) + (I_{OE}+\Delta I_E)R_E + V_{OBE}$$

whence $$\Delta V_B = 0.06\log_{10}\left(1+\frac{\Delta I_E}{I_{OE}}\right) + I_{OE}R_E\frac{\Delta I_E}{I_{OE}} \qquad (4)$$

We have thus got rid of the awkward I_S term. The equation relates the change of base voltage from the quiescent value which produces the quiescent emitter current I_{OE} to the fractional change of emitter current $\Delta I_E/I_{OE}$, which may be positive or negative but can never exceed unity. The remaining term $I_{OE}R_E$ is the one which affects linearity. It represents the emitter voltage above $-V_{CC}$ for an unbypassed emitter resistor which carries no current but I_E.

Fig. 2 shows in curve A a plot of ΔV_B against $\Delta I_E/I_{OE}$ when $R_E = 0$. This is the basic voltage-current relation for a diode and is highly non-linear. Curve B shows the relation when $I_{OE}R_E = 0.5$ volt. It is an almost perfect straight line. Reading from the curve for $\Delta V_B = \pm 0.42$V, $\Delta I_E/I_{OE}$ is 0.8 and -0.76, which represents second-harmonic distortion of only about 2.5%. Distortion can be further reduced by increasing $I_{OE}R_E$ but at the expense of gain.

It should be noted that the results apply to any transistor in the region where the base-emitter diode follows the fundamental diode equation. This is most small-signal transistors at current up to about 10mA. The peak signal output (current) for low distortion is about 75% of the quiescent current.

Of course, the collector resistance must not be too large in relation to the supply voltage, otherwise bottoming will occur with severe distortion. This is, however, a

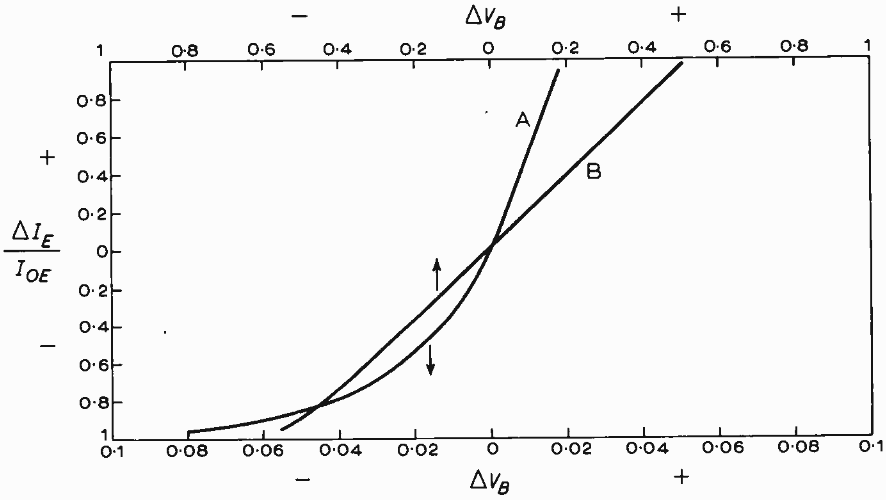


Fig. 2. Input/output curves for the stage of Fig. 1. Curve A is for $R_E = 0$ and curve B for $I_{OE}R_E = 0.5$ volt.

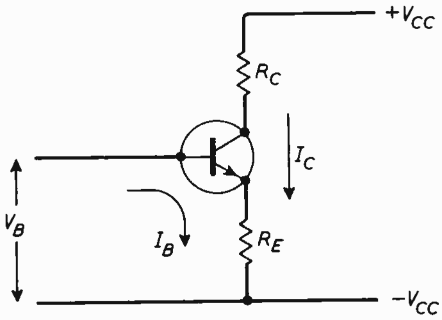


Fig. 1. Basic circuit of common-emitter stage.

quite independent mechanism which need not be considered here.

It is worth noting that an external base resistance R_B also exercises a linearizing effect. It is equivalent to an emitter resistor of value R_B/h_{fe} and, if large enough, gives constant-current drive. It is thus clear that the so-called constant-current drive and linearization by an emitter resistance are fundamentally the same. They both operate by making the actual base-emitter voltage changes very small compared with the signal input voltage applied to the transistor stage as a whole.

When two transistors are in push-pull, much of the basic distortion is balanced out. The true push-pull case is less easily computed than that of a phase-splitter; that is, the case of a push-pull pair with an input to only one transistor. Ignoring the small loss in the coupling resistor R_T, Fig. 3, the two collector signal currents are equal and opposite and so the input for a given current change ΔI_C in Tr_1 is ΔV_{B1} minus the input, ΔV_{B2}, for the same magnitude of current change in Tr_2. As the latter current change is negative, so is ΔV_{B2}, and the input is really the sum of the magnitudes. Curve A of Fig. 4 shows the resultant for $R_E = 0$ and is much more linear than curve A of Fig. 2. Curve B shows the result when $I_{OE}R_E = 0.1$V. This is not as good as curve B of Fig. 2, but that is for $I_{OE}R_E = 0.5$V. In all cases, for the same value of $I_{OE}R_E$, push-pull operation gives better linearity.

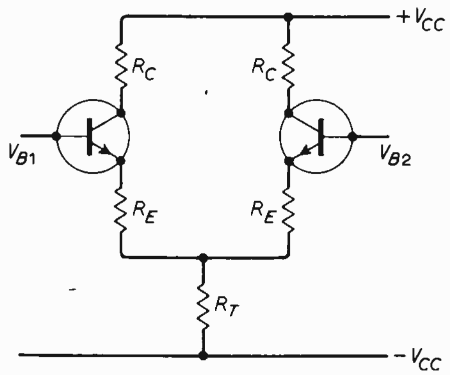
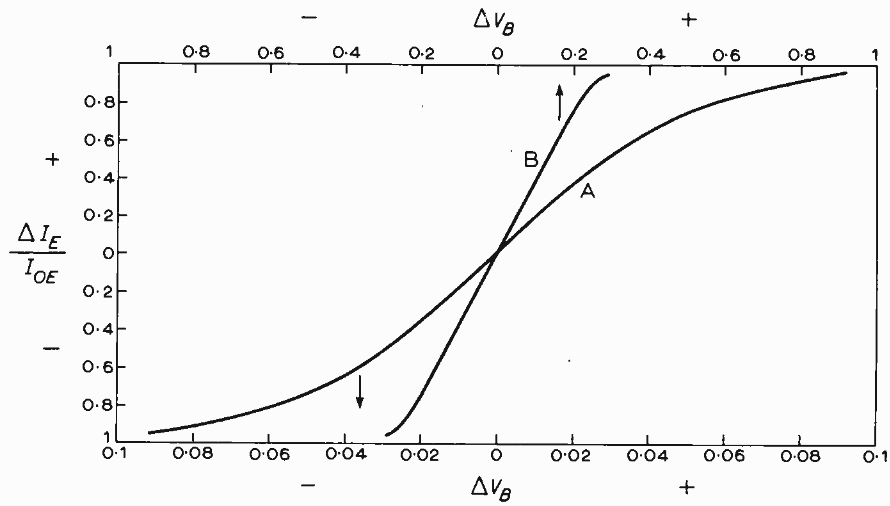


Fig. 3. Basic circuit of a push-pull stage.

Fig. 4. Performance of push-pull stage, curve A with $R_E = 0$ and curve B with $I_{OE}R_E = 0.1V$. The curves are for an input to one base only, the other being held at a fixed potential.



Tunable Phonon Sources

Hypersonic mechanical waves from sources which are 'essentially monochromatic', and can be tuned over the frequency range 150 to 280GHz, have been generated experimentally at Bell Labs in the U.S.A. Powers of up to hundreds of milliwatts are claimed. Hitherto it has been difficult to sustain coherent 'monochromatic' vibrations in a material and the energy has had the nature of wide-band noise, or just heat. One possible use of this new type of mechanical wave source is in phonon spectroscopy at previously inaccessible frequencies (the thermal frequencies). By this method new avenues of study may be opened into the electronic properties of materials, impurities and defects in crystals and the interaction of acoustic vibrations within a solid. Already the method has been used to observe the ground state splitting of vanadium impurity ions in sapphire. (The phonon is the quantum mechanical 'packet of energy' of mechanical waves, analogous to the photon of electromagnetic waves.)

The phonon generating device consists of a heater, in the form of an alloyed metal film of copper and nickel (Constantan), mounted near to a thin superconducting film of tin but separated from it by a thin layer of electrical insulator. The tin film is attached to the material being studied by phonon spectroscopy, and is sealed inside a cryostat (low temperature chamber), which is used to lower the operating temperature to 1.3°K.

The heater is turned on and off to create heat pulses. Each pulse of heat forces phonons into the superconducting film. In a superconductor the loss of resistance results from electrons becoming bound together in pairs. The phonons from the heater break these electron pairs apart, but when the electrons recombine they emit another phonon of a particular energy which is equal to the binding energy of the electron pair. The breaking and recombination of pairs of electrons generates a narrow band of high-frequency phonons.

In order to tune the phonon source, a magnetic field is applied, parallel to the thin film superconductor. The energy gap of the superconducting electrons decreases uniformly as the magnetic field is increased, and because of this the phonon frequencies can be changed by varying the magnetic field. In superconducting tin, for example, phonons can be tuned over the 150 to 280GHz range mentioned above. Bell Labs say they expect to be able to produce phonon frequencies from about 50GHz to 1THz by experimenting with different superconducting materials and tuning them. They think they may be able to generate phonon powers of more than one watt by this technique.

To detect the phonons the researchers use a second superconductor, operating in a complementary way to the one in the phonon generator. They also make use of a property of an antimony doped germanium crystal: the antimony can be tuned by squeezing the crystal so that its energy gap corresponds to that of the superconduction energy gap. The crystal can be made to absorb phonons of only a particular frequency, and this is detected through a decrease in the amplitude of the observed signal.

News of the Month

Could MADGE replace ILS?

The Government is considering whether to adopt a radio interferometer type of aircraft landing aid for use on military airfields. Called MADGE (Microwave Aircraft Digital Guidance Equipment), it has been designed as an approach aid for helicopters and fixed wing aircraft and is the result of a private research programme at Mullard Research Laboratories. Last year it was the winner of a NATO international competition for guidance systems and is now being developed by the M.E.L. Equipment Company (a Mullard subsidiary) for the Ministry of Defence. Mullard and M.E.L. naturally hope that it will become a NATO standard system, after its success so far, but this will depend on its first being officially adopted by the British Government. Meanwhile Mullard are continuing to improve the system — notably to give better elevation information at low approach angles of the order of 1° — and are making no secret of their longer term hope that MADGE might prove a viable replacement for the well known ILS (Instrument Landing System) used in civil aviation throughout the world.

In the MADGE system a 5GHz beacon in the aircraft transmits a pulsed signal to three angle measuring receivers on the ground which accurately measure the direction of arrival of the signal by means of the multiple lobes of aerial interference patterns. Two receivers, in boxes of horizontal format, measure azimuth angles, one in the 'approach' sector and the other in the 'overshoot' sector. The third receiver, in a vertical-format box, measures elevation angle in the 'approach' sector only. The receivers derive angular information on each transmitted pulse in the form of parallel digital codes. These codes are fed to a data link transmitter (housed in the 'approach' azimuth receiver box) where the angle data is first encoded in serial form and then transmitted back to the aircraft as an amplitude modulated pulse train. In the aircraft a receiver decodes the angle data and compares it with azimuth and elevation angles which have been set as a desired approach path by the pilot. Error signals are derived in analogue form and these pass to conventional crossed-pointer indicators as used in ILS — or could be fed to an autopilot. The distance between the aircraft and the landing site is derived digitally in the aircraft from the go-and-return time taken by the transmitted pulse, and is presented on an analogue indicator. Velocity is derived by differentiating the analogue value representing distance.

Aircraft approach angles in azimuth can be up to $\pm 45°$ about a central datum line (likewise for overshoot), while approach angles in elevation can be between 1° and 25°. The range of the azimuth and distance measuring equipment is 27km and that of the elevation measuring equipment 18km.

It will be noted that this system differs from ILS in that the approach information is derived on the ground — which, incidentally, raises the possibility of using the system for air traffic control — and that it provides a wide range of approach angles for the pilot to select from (ILS giving only a single glide path at an elevation angle of 3°). MADGE is claimed also to have a greater immunity from the effect of ground obstacles, such as buildings, trees and vehicles, on the aerial radiation patterns and hence on the accuracy of measurement. Another way in which it differs from ILS is that the ground equipment is small, portable and battery operated, consisting of three boxes each about 4ft 6in long and 20kg in weight. The aircraft equipment weighs 7.9kg.

British Rail centralize freight control

A contract valued at over £26,000 has been awarded to International Aeradio Limited by British Rail for the manufacture and installation of a communications data control centre. It will form an important part of the new British Rail computer and freight management system which is to start operations in 1973.

The system, known as TOPS (Total Operations Processing System) represents an investment by British Rail of around £10M and is based on a similar system developed over a period of eight years by the Southern Pacific Transportation Corporation of the United States. The computer equipment will be a dual IBM 370/165 to be installed close to the British Rail headquarters at Marylebone and the communications equipment will be installed in a control centre close to the computer hall.

The object of TOPS will be to increase the efficiency of handling of the British Rail goods fleet of 300,000 wagons and it is planned that this fleet will eventually be reduced to 175,000 wagons by 1976 while the freight tonnage is estimated to increase from the present 200 million tons to 220 million tons. The net annual return forecast from this computer-controlled system is estimated at £5M by 1978.

International Aeradio's contribution to the programme involves the manufacture of a comprehensive set of data signal processing, switching and monitoring equipments which will be situated between the IBM computers and the 430 remote data terminals which are to be located at the numerous British Rail freight terminals and area headquarters. The design of the data control centre has been carried out by a joint British Rail/IAL engineering team.

The complete data communications network, which utilizes the British Rail national telecommunications network, will be controlled from the IAL monitoring console and will enable the British Rail controllers to monitor the performance of each of the low-speed and medium-speed data paths. Where circuit degradation occurs the system enables data path re-routing to be carried out both within the main control centre and at the remote terminals.

American boost for European EVR

When the Columbia Broadcasting System (whose CBS Laboratories originated EVR — Electronic Video Recording*), withdrew from the international EVR Partnership set up to develop the system commercially, many people saw this as a severe blow to the future of the Partnership. With several rival systems of video recording becoming available, it seemed as if the inventor had no faith in what he had produced. Now another American company, Motorola, has stepped in to provide commercial support — though not as a member of the Partnership, which is owned by the European companies ICI and Ciba-Geigy.

Motorola has been manufacturing EVR players, and producing and distributing cassette programmes for them, for some time in the U.S.A. Their new move, after 're-evaluating our EVR posture', is to start designing, manufacturing and exporting EVR players for operating on European television standards. In the U.K. they have made agreements with two firms to provide outlets for the products: Environmental Visual Systems Ltd, of London, will be responsible for purchasing, marketing and selling the players, and Telefusion Ltd, of Blackpool, for rental, installation and servicing. The rental element in this arrangement is seen by Motorola as particularly important as no such outlet exists in the U.S.A., and, since the players are quite expensive (about $800 in the U.S.A. and $1000 in Europe), renting is likely to prove more attractive to potential users than outright purchasing. The first players are expected to arrive in Britain in August.

Environmental Visual Systems Ltd will also be marketing EVR programmes, in the form of cassettes for the players. These programmes will come from various sources, including the Motorola Teleprogram Center in the U.S.A., and some will be initiated or promoted (though not physically produced) by E.V.S. Ltd. The programmes, whatever their original form (e.g. conventional film or magnetic video tape), will be converted to cassette form by the EVR Partnership at their Basildon, Essex, processing plant.

Customers for the players and programmes are expected to be mainly professional organizations. Motorola see little possibility of a domestic market in the immediate future.

Mini-computer price cut

The U.K. market for large computers is about £300M a year at present and is expanding at a rate of around 20% per annum. Small computers, while enjoying only about 7% of this market, are increasing their sales at between 40 and 45% a year and should catch up with the larger computer market by 1980. At present the small computer scene is dominated by the Americans and the Digital Equipment Corporation (of PDP-8 fame) are said to hold about 40% of the world market.

Very soon we will see small computers emerge from Japan and a price cutting war will ensue which at present has parallels in the desk calculator and the digital integrated circuit market places.

One of Britain's small computer manufacturers, Computer Technology Ltd who make the Modular One machine, have increased their share capital by £300,000 to £1.7M so that they can step-up production and have cut their prices by an average of 25%.

Computer Technology are gambling that these measures will enable them to greatly increase production so that the price can be held down. They will be helped if discussions with I.C.L., which are currently in progress, come to fruition.

I.C.L. may be using Modular One computers as 'front-ends' for their own large machines. The result could be a steady flow of orders for Computer Technology.

Computer Technology claim that with the price reductions 'the Modular One computer is the best buy in the world in terms of computing power per pound sterling'.

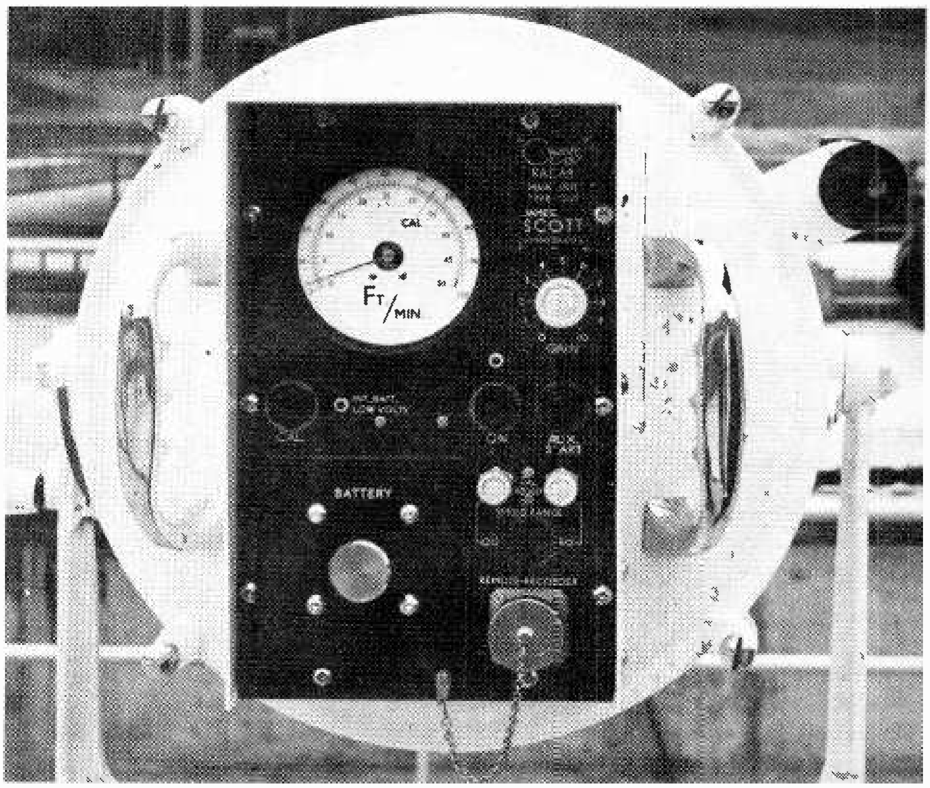
Machine speech combats machine noise

Anyone who has tried to make himself heard above the noise of machinery in a factory will be aware that it requires a special technique of pitching or projecting the voice rather than just shouting. A recent experiment at a telephone equipment factory of Western Electric, Oklahoma City, U.S.A., suggests that maybe a machine-made voice is really needed to get through machine-made noise. Wiring instructions helpful to production-line workers were produced in the form of synthetic speech constructed from digital data stored in a computer. The audio signals were recorded on a cassette tape recorder and played back through an ear-piece to a female 'wireman', who could stop and start the recorder by means of a foot pedal. In commenting on the experiment this wireman remarked that the 'caricatured' nature of the synthetic speech made it easier to understand than ordinary human speech when in competition with typical factory noise.

The synthetic speech was produced by equipment and programmes devised by engineers at Bell Labs, New Jersey. They used an existing technique for synthesizing speech from stored digital data representing the characteristic formants in human speech. In this technique, individually spoken words are analyzed for their characteristic formants, and stored in a DDP-516 digital computer. When a particular 'answer back' word sequence is needed, the required formants are accessed and linked together. Rules of speech are applied so that each word will be used with voice pitch and pauses that sound natural for its position in the sequence. Finally, the formant, pitch, and timing data are sent to a digital filter which simulates the resonances of the human vocal tract. The filter output is converted from digital to analogue form to produce synthetic speech.

In the wiring instruction experiment simplified duration and pitch rules could be used because the utterances were essentially short phrases. With such

Two of Britain's largest tankers, the Hudson Venture and the Hudson Friendship, have adopted a doppler system called 'Cee-Wave' for monitoring approach speed when docking. The equipment, pictured, manufactured by James Scott (Electronic Engineering) Ltd is mounted amidships on the appropriate side and gives a continuous indication of closing velocity from 0-100ft/min at a range of up to 1000ft. Read-out is on two large moving-coil meters; one is fitted on the instrument itself and another on a remote display unit mounted on the bridge.



*See "Colour Electronic Video Recording", *Wireless World*, August 1970 phrases, pitch and duration data are less critical than they would be in the recital of complete sentences. The wiring instruction information was generated originally as a pack of computer punched cards. The cards were simply put into the card reader of the computer, and there the spoken wiring instructions were synthesized.

The large storage capacity and flexibility of a computed system are not really needed for the limited vocabulary of the wiring instructions, but the high speed of the system allows frequent changes to be made in the wiring specifications.

Printed word plus programme on a.m.

Barry Research of California, U.S.A., have introduced a teleprinter system which makes it possible for a.m. transmitters to broadcast teleprinter data at 60 to 100 words per minute at the same time as programme material without the quality of the programme material being degraded. At the transmitting end a teleprinter is used to produce input paper tapes containing the message to be printed by the receiving stations. A converter produces a sub-carrier which is phase modulated by the information on the paper tape; the sub-carrier then phase modulates the a.m. carrier to be transmitted (the carrier is also amplitude modulated with the programme material).

A standard broadcast receiver is used for reception. The coded subcarrier is extracted from the receiver's i.f. output and demodulated in a separate unit which drives a teleprinter.

The development means that it would be possible to set up a national (or even international?) hard copy news service using existing transmitting stations.

Large liquid crystal display

Often in control engineering a display is called on to show some static information (system block diagram) and some dynamic information (varying quantities or symbols) as well. Brown Boveri have developed a large liquid crystal display which uses a photographic transparency to depict the static information. The dynamic information employs the 'dynamic scattering' liquid crystal effect and the characters and symbols are addressed matrix fashion. The matrix is fed with both a.c. and d.c. pulses which result in fully energized elements having a high optical output while energized elements have their threshold voltages artificially increased thereby increasing the ratio of reflectivity between 'on' and 'off' elements.

Physics Exhibition and LABEX

The management of U.T.P. Exhibitions Ltd and the council of the Institute of Physics have entered into an agreement concerning LABEX International 1973 and The Physics Exhibition. Next year, the two exhibitions will be held simultaneously at Earls Court, London (9-13 April). LABEX will occupy the ground floor and the Physics Exhibition will occupy the first floor but passage from one to the other will be possible.

Each exhibition will retain its traditional features, including for example, the usual programmes of lectures, films and discussion meetings. Where appropriate, the two organizers have agreed to co-ordinate their programmes. The agreement applies to the 1973 exhibitions only, but if the venture succeeds a continuing biennial event on similar lines is envisaged for the future.

ESRO council president re-elected

At its first meeting this year (in Paris in March) the Council of the European Space Research Organization unanimously re-elected its president, Professor G. Puppi of Italy, who agreed to serve until the end of 1972. In addition the Council re-elected J. van Eesbeek of Belgium and elected General L. de Azcarraga of Spain as its two vice-presidents.

Communication 72, programme

The provisional programme for the Communication 72 three-day conference, jointly organized by *Electronics Weekly* and *Wireless World*, to be held in Brighton this year is detailed below. The conference is being run in the mornings with three parallel sessions with one Highlight paper, presented by an eminent personality, each day. Delegates will be free to visit the associated exhibition in the afternoons. In the evenings there will be discussion groups and a conference banquet.

June 13, Session 1 — MOBILE RADIO

9.00 'Control aspects of v.h.f./u.h.f. communication systems' by C. G. Gibbs (Cossor Communications).

9.45 'Personal radio for the seventies' by A. M. Jones (Pye Telecommunications).

11.00 'Selective calling as applied to mobile and portable radiotelephones' by D. A. Clare (Pye Telecommunications).

11.45 'Test equipment for mobile radiotelephone systems' by P. MacNamara (Telecommunications Ltd).

June 13, Session 2 — DATA COMMUNICATIONS

9.00 'Integrated circuits in low-speed modems' Motorola Applications Laboratory.

9.45 'A new group delay measuring set for audio circuits' by A. Barlucchi and N. Montefusco (Siemens).

11.00 Highlight paper — 'Data forecasting in the Post Office' by R. Gadd (Post Office).

June 13, Session 3 — POINT-TO-POINT COMMUNICATIONS

9.00 'A comparative study of the use of the 890-960MHz band for point-to-point communications' by Dr. R. D. Egan (Granger Associates).

9.45 'Recent advances in the design and performance of h.f. communication equipment' by G. J. Lomer (Racal Communications).

11.00 'Developments in all-solid-state broadband h.f. transmitters' by P. Dillon (Redifon Telecommunications).

11.45 'HF communication and the spectral purity of frequency synthesisers' by Dr. Grabe and Herr Gerhold (Rohde & Schwarz).

June 14, Session 4 — COMMUNICATIONS IN TRANSPORT

9.00 'Developments in maritime communications' by C. E. Secker (Post Office).

9.45 'RADIAX radiating coaxial cable' by J. R. Avery, Dr. K. R. Slinn and D. A. W. Whitson (Andrew Antenna Systems).

11.00 'Railway communication systems' by M. Millett (Nelson Tansley).

11.45 'Tunnel communications for London Transport underground v.h.f. radio scheme' by J. Elliot MacDonald (Storno).

June 14, Session 5 — MOBILE RADIO

9.00 'Computer derived radio network surveys' by Dr. J. Durkin (M.P.T.), M.A.P. Longy of PMA Consultants, and C. E. Dadson (Joint Radio Committee of N.P.I.).

9.45 'Multiple fixed station sites in the private mobile radio service' by B. Armstrong (Pye Telecommunications).

11.00 Highlight paper — 'The problem of area coverage in relation to mobile radio' by J. P. Titheradge, Directorate of Telecommunications, Home Office.

June 14, Session 6 — TEST EQUIPMENT

9.00 'Automatic test equipment in r.f. communications' by E. R. Valentine (Honeywell).

9.30 'The use of A.T.E. for testing communication systems and sub-assemblies' by A. G. Hayes and D. McCloud (British Aircraft Corporation).

11.00 'A new synthesised signal generator' by K. R. Thrower (Racal Instruments).

11.45 'An equipment for measuring the a.m. and f.m. noise spectra of c.w. and pulsed signals at microwave frequencies' by J. G. Humphries (Com-Rad Electronic Equipment).

June 15, Session 7 — MOBILE DATA COMMUNICATIONS

9.00 'Mobile fascimile communications' (Muirhead).

9.45 'Pinpoint — a vehicle location and monitoring system' by R. W. Gibson (Mullard Research Laboratories).

11.00 'High-speed mobile radio data interrogation and retrieval' by A. K. Sharpe (Pye Telecommunications).

11.45 'A specialised real-time management information system' by J. Kuykendall (General Telephone and Electronics International Systems).

June 15, Session 8 — MILITARY RADIO COMMUNICATION

9.00 'Manpack and military vehicle radio equipment for forward areas' by E. Ribchester *et al* (British Communications Corporation).

9.45 'Routeing problems in mobile trunk networks' (Signals Research and Development Establishment).

11.00 Highlight paper — 'Evaluation of Communications Equipment for military use' by General Sawyer, Signal Officer-in-Chief of the Army.

June 15, Session 9 — NEW TECHNIQUES

9.00 'Computerized receiver frequency control' by M. J. Crisp (Watkins-Johnson International).

9.45 'Software Solutions to store and forward message switching problems' by R. I. Clark (Plessey Electronics).

11.00 'A high efficiency d.s.b. transmitter' by V. Petrovic (University College, Swansea).

11.30 'Demodulation techniques for diminished carrier a.m.-derived systems' by R. J. Holbeche (University College, Swansea).

The Transmission-line Loudspeaker Enclosure

A re-examination of the general principle and a suggested new method of construction

*by A. R. Bailey**, Ph.D., M.Sc., M.I.E.E.

Since the wool-filled transmission-line loudspeaker enclosure was first described† there has been a steadily increasing interest in its use.

The basic transmission-line enclosure is shown in Fig. 1. Radiation from the back of the driver cone flows down a pipe filled with a low-density sound-absorbing material. Fibrous absorbents such as loose wool, cotton wool and kapok can be used; sound absorption decreasing as the frequency goes down. In general it is very difficult to obtain good absorption if the path length is less than one-quarter wavelength of the sound in free-space; at 30Hz this corresponds to a path length of about 9ft.

If the pipe length is less than 9ft, sound at and below 30Hz will emerge from the open end of the pipe. Due to time delay in the pipe, the sound will not start to cancel the radiation from the front of the cone until the effective pipe-length is less than one-sixth of a wavelength. It is therefore possible to use the radiation from the open end of the pipe to reinforce that from the front of the loudspeaker cone at low frequencies.

The effect of the wool filling in the pipe is to slow down the wave relative to its velocity in free air. This reduction factor is between 0.7 and 0.8 for the recommended packing density, so the system will operate down to a somewhat lower frequency than would otherwise be expected.

* University of Bradford

†Bailey, A. R., 'Non-resonant Loudspeaker Enclosure', *Wireless World*, Oct. 1965.

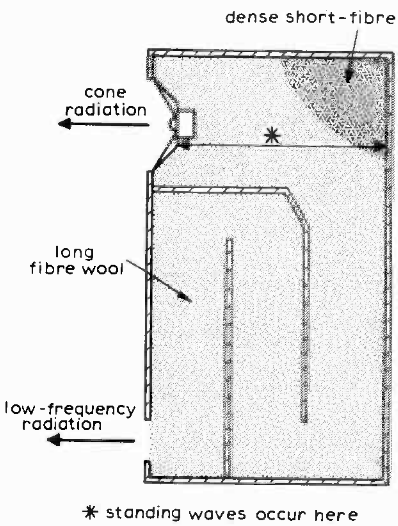


Fig. 1. Original transmission-line cabinet design.

The folding in the original cabinet design caused sound coloration due to reflections at the bends — particularly the first one at the back of the cabinet. The degree of coloration introduced by this first reflection (which, incidentally, is present in all plain box-shaped cabinets) was quite serious with the high crossover frequency of 1500Hz. Certainly the reproduction without it sounded as if an echo had been removed. The reasons for this were investigated.

In a simple closed box, as shown in Fig. 2, a sound impulse generated by the cone will have two components — the direct radiated pulse from the front of the cone and that propagated back into the cabinet. If this latter is assumed to be a plane wave, i.e. sound travelling parallel to the cabinet sides, it will strike the back wall and bounce back to the cone still as an impulse. Some of this energy will radiate through the cone to the outside and the remainder will be reflected back into the cabinet for re-reflection. The net result is a succession of steadily weakening pulses being radiated from the cabinet. The acoustic output will therefore be as shown in Fig. 3.

If the reflection from the back wall of the cabinet is changed so that it is gradual rather than abrupt, then the reflected wave will not be a unit impulse from an initiating unit impulse, but a pulse whose length and shape will depend on the nature of the reflection. This may be more readily appreciated by referring to Fig. 4. Here, the back wall of the cabinet is triangular. Sound radiated from the back of the cone is successively subjected to reflection, the first reflection occurring due to sound from the edge of the cone, and the last being due to that from the centre. As the path lengths for these reflections are very different, the sound in the cabinet will not emerge as a unit impulse but as a much lower amplitude long pulse — more like a continuous low level sound. This is subjectively far less noticeable to the ear than a series of decaying impulses. The two effects are compared in Fig. 5.

Obviously, it would be far better to

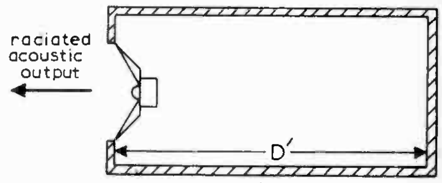


Fig. 2. Unlagged totally enclosed cabinet.

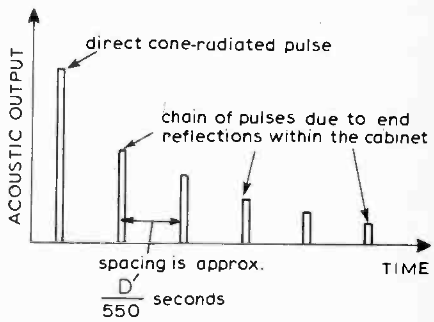


Fig. 3. Pulse-response (idealized) of cabinet shown in Fig. 2.

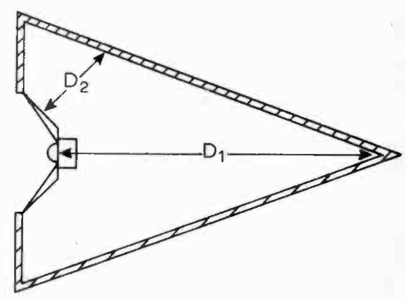


Fig. 4. Triangular cabinet showing different reflecting path lengths.

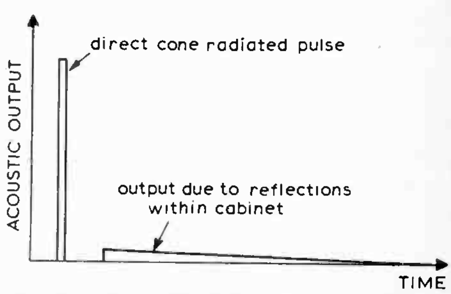


Fig. 5. Approximated output for triangular cabinet excited with impulse from loudspeaker.

abs sound completely on the back wall of the cabinet. Here we face the impossible, but it is imperative to make the best use of absorbent within the cabinet.

Towards a solution

The basic requirement in using a forward facing loudspeaker unit is to guide the sound into the vertical direction without producing bad reflections. In addition the system must be fairly simple, to keep woodworking costs low, and also mechanically strong, to avoid significant panel resonances.

After experimenting with different cabinet shapes, the arrangement shown in Fig. 6 was arrived at. This has several advantages over the original design. First, the pipe is triangular in cross-section, thus giving less audible coloration due to reflections. Secondly, the woodwork is very simple; only two internal partitions are necessary. Thirdly, the front of the cabinet and the large partition are automatically braced so reducing panel resonance. Finally, sufficient area is available on the front of the cabinet to mount a mid-range unit in part of the line remote from the back of the bass driver when internal pressures are not too high. In practice, it has not been found necessary to use a separate enclosure for the mid-range unit.*

Details of the partition sizes are shown in Fig. 7. A three-speaker system is recommended, as there is at present a difficult 'gap' between known low-coloration bass units and tweeters.

Incidentally, it should be noted that many 'high-fidelity' drive units are that in name only. Frequency-response is only one aspect of performance, and the transient response is far more important overall. Pulse or step testing loudspeakers in a long matched acoustic transmission line is most illuminating. Some units will still be radiating appreciable acoustic power 50ms after the exciting pulse has disappeared!

The units specified are known to have good transient response and are available with a suitable crossover network. Such networks are very difficult to design and it is *not* sufficient to use a general-purpose crossover unit. Unfortunately, loudspeakers do not behave as pure resistance at all frequencies — often quite the contrary. Design of crossovers from this assumption is completely incorrect and it is not uncommon for correct inductor sizes in a network to be double that expected from simple theory. In addition the different phase-angles of speakers at the crossover frequency complicates matters even more, and bad design can lead to abnormally low impedance levels over some parts of the frequency range. In short, crossover networks *must* be designed to operate with the speaker units that they are to be used with or very peculiar results can be obtained.

As previously, long-fibre wool is recommended as the continuous acoustic absorbent that fills the whole of the transmission line pipe. The wool must be well teased out or it loses its effectiveness. Anchoring the wool is something of a problem as it can compact with transport or use over a period. Nails or dowels projecting from the partitions will serve, but make stuffing difficult. The best suggestion yet made is to use a 'Netlon' core for the wool, fibres being teased through it and left sticking out all round. Using front-mounting loudspeakers as specified, the two front pipes can be loaded through the speaker and port apertures, and the rear pipe is easily filled by removing the back of the cabinet. Alternatively, if the cabinet top is made removable all three pipes can be loaded from the top. A packing density of about ½lb per cubic foot is about right. Excess wool will cause back-pressures on the cone, and too little will cause pipe resonances in the low bass region.

What column resonance remains in the system can be reduced by putting 45-degree corner reflectors at the back of the speaker and also at each side of the first bend. These are not critical but should be so arranged that sound from the back of the cone will 'bounce' down the first pipe and then up the second, i.e. consider the sound to be light and the reflectors to be mirrors. The improvement is only about 1dB in frequency response when using reflectors, and as this improvement is only just detectable it may well be decided to omit them. If included, they should be made from ¾in chipboard or some similar material, and firmly fixed to the cabinet.

The port area is not critical as there is none of the tuning effect that occurs in the bass reflex enclosure. Changes in port area of two-to-one ratio produce no noticeable effect, but nevertheless it would be unwise to make the port much smaller than that given in the drawings as this is already considerably smaller in area than the pipe feeding it.

As mentioned previously, response curves should be treated with extreme caution as they represent only part of the performance of the speaker. Nevertheless the overall response curve should be as flat as is reasonably possible. The curve for the complete system when measured in an anechoic chamber is shown in Fig. 8.

In looking at this curve several points must be noted. First, the falling bass response shown will not occur in the same manner in a room. The presence of a floor will give a 3dB lift due to the absence of diffraction in the downward direction.

*However, there may be audible improvement if an absorbent filled enclosure is used behind the mid-range unit. ED.

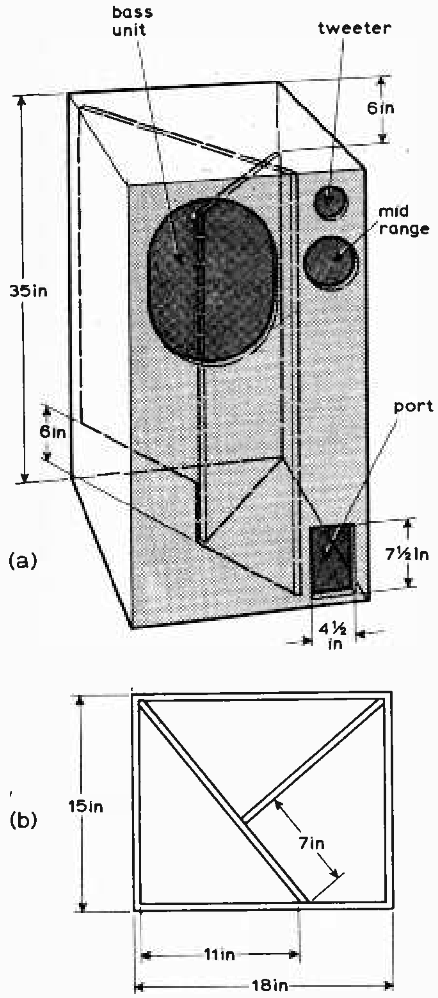


Fig. 6. Final design.

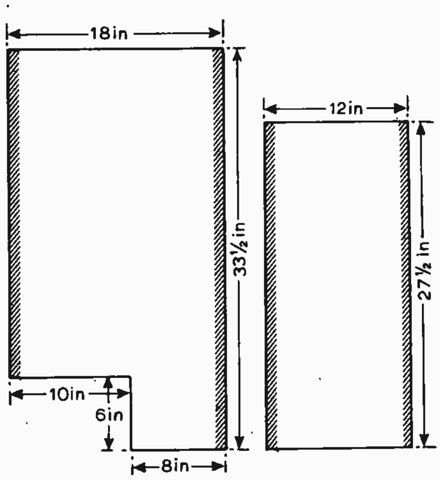


Fig. 7. Dimensions of internal partitions.

Recommended drive units and crossover networks

Bass driver	BD 25/1	£12.50
Mid-range unit	MD 9/2	£8.50
Tweeter	TD 3/2	£6.25
Crossover network	FN 10	£8.75

Radford Audio Ltd., Bristol BS3 2HZ.

Alternative units

Bass driver	B 139	£12.50
Mid-range	B 110	£7.80
Tweeter	T 27	£5.00
Crossover unit	DN 12	£4.50

K.E.F. Electronics Ltd., Tovil, Maidstone, Kent. (The performance with the K.E.F. units is improved if the Radford crossover is used in place of the K.E.F. crossover.) Long-fibre wool may be obtained from John W. Pennington (Dowley Gap) Ltd., Midland Wool Warehouses, Briggate, Windhill, Shipley, Yorks. The cost is 53p/lb including postage.

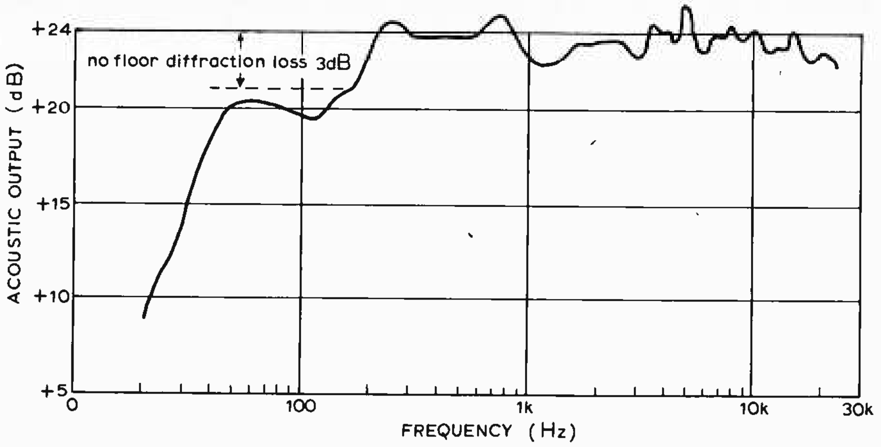


Fig. 8. Anechoic response curve of complete loudspeaker system.

Similarly, walls and ceilings add to the on-axis bass output. In fact a 'flat' response in an anechoic chamber will sound very bass heavy when the system is used in a normal room. Under normal room conditions the bass output of the system described is quite adequate, windows being easily rattled at 30Hz.

The smoothness of the overall curve is the most critical point, the odd decibel of gain or loss in overall response being far less important than a response with a smooth envelope. A jagged response curve means high Q components in the ouput and these are very noticeable on test. In fact a high Q resonance that lifts the overall response by only 0.2dB may ruin the reproduction of an otherwise excellent speaker. Transient response testing is the ideal answer, but interpretation of the results is very difficult at present except on a rather empirical basis.

The ultimate test is the ear, but one must always remember that personal prejudice can enter into things to a very large extent. For this reason the best test material is not music but such things as pure sine waves (for distortion) and applause, or better still white and pink noise (for transient response).

Regarding the use of pink noise, which incidentally is only white noise attenuated by 3dB per octave with increasing frequency, the following incident happened to the author. A pink noise generator had been built and was being tested with a speaker system. A most noticeable hum was produced and all attempts to find the source in the generator failed. It was finally discovered that the hum was not present in the noise, but was the fundamental resonance present in the speaker system. This speaker (which was of the unlagged reflex type) had previously been used in music tests, and several people had commented on the good performance, particularly in the bass region.

If the cabinet size is felt to be rather too large, then it is possible to scale all the dimensions given according to the diameter of the bass driver being used. This will result in a poorer bass performance and is not advised, but a very creditable performance is possible using an 8in driver unit and scaling all the dimensions by four fifths. This factor results from the recommended bass driver having the same effective cone area as a normal 10in unit of circular construction.

In conclusion it must be emphasized that only the system as described is in any way guaranteed. Readers can experiment, of course, but must be prepared to solve their own problems.

Sixty Years Ago

May 1912. *The Marconigraph* carried a long article devoted to the loss of the *Titanic* and the part wireless played in the rescue of the survivors. The *Titanic* sailed on April 10th and on April 15th struck an iceberg. The distress call was answered by the *Carpathia*.

The wireless installation on board the *Titanic* probably represented the state of the art at the time and was described in the article as follows.

'The wireless equipment of the *Titanic* was the most powerful possessed by any vessel of the mercantile marine. Its generating plant consisted of a 5-kW motor-generator set, yielding current at 300 volts 60 cycles. The motor of the set was fed at 110 volts d.c. from the ship's lighting circuit, normally supplied from steam-driven sets; while, in addition, an independent oil-engine set was installed on the top deck, and a battery of accumulators was also provided as a stand-by. The alternator of the motor-generator set was connected to the primary of an air-core transformer, and the condenser consisted of oil-immersed glass plates. To eliminate as far as possible the spark-gap and its consequent resistance, which as is well known, is the principle cause of the damping of the waves in the transmitting circuit, the ordinary Marconi rotary disc discharger was used. This is driven off the shaft of the motor-generator. The guaranteed working range of the equipment was 250 miles under any atmospheric conditions, but actually communication could be kept up to about 400 miles, while at night the range was often increased up to about 2,000 miles. The aerial was supported by two masts, 200ft. high, stepped 600ft. apart, and had a mean height of 170ft. It was of the twin T type, and was used for the double purpose of transmitting and receiving. The earth connection was made by insulated cable to convenient points on the hull of the vessel.

'The receiver was the Marconi standard magnetic travelling band detector used in conjunction with their multiple tuner, providing for the reception of all waves between 100 and 2,500 metres. The multiple tuner was calibrated to permit of the instrument being set to any prearranged wave-length, and further to be provided with a change switch to permit of instantaneous change of the circuit from a highly syntonized tuned condition to an untuned condition (for stand-by) especially devised for picking up incoming signals of widely different wave-lengths. By reason of its robust nature the magnetic detector could be employed permanently connected to the transmitting aerial, thus dispensing with all mechanical change over switching arrangements.'

Cheap, Stable Local Oscillator

Receivers for inexpensive communications systems — which might include future s.s.b. broadcast receivers — need a variable local oscillator which is both cheap and highly stable in frequency. Unfortunately the ordinary continuously tunable LC oscillator although cheap is not sufficiently stable, whereas the discretely stepped frequency synthesizer although highly stable is by no means cheap. A new type of variable oscillator, devised by M. J. Underhill of Mullard Research Laboratories, seems to be a useful compromise between these two extremes. The frequency of the oscillator is stabilized by an automatic control system which uses as a reference a fixed period of time, provided by an ultrasonic delay line. This delay line is in fact a standard mass-produced component for PAL television receivers and is therefore quite inexpensive. The new oscillator provides a number of equally spaced locking frequencies, each of which can be fine tuned by a voltage control to give full coverage of a frequency band. The stability of the oscillator is determined ultimately by the stability of the phase delay of the delay line used.

A block schematic of the oscillator is shown in Fig. 1. The PAL delay line gives a phase shift proportional to frequency. The phase shift changes by 2π radians for a frequency change of 15.625kHz, this value being given by the reciprocal of the delay time. Thus a phase detector measuring this phase can be used to lock the frequency of a voltage controlled oscillator to a number of frequencies spaced by 15.625kHz. In Fig. 1 a phase detector with a linear response for phase differences between $-\pi$ and $+\pi$ is used in the closed-loop control system. The feedback loop controls the oscillator frequency so that the error between the output voltage of the linear phase detector, V_p, and a variable reference voltage, V_r, is minimized. In this way V_r can be used to set the phase for which the loop will lock, and by varying the phase between $-\pi$ and $+\pi$ radians the locking frequency points can be swept over a 15.625kHz range. Because practical linear phase detectors cover a phase range of just less than 2π, an inverter and phase reversing switch can be used as shown to give two overlapping fine-tune frequency ranges. The switch can be operated either manually or automatically.

If an error amplifier with a finite gain is used and the range of the phase detector output voltage is arranged to be less than the range of the reference voltage, the reference voltage tuning control can also be used to jump the frequency in a controlled manner from one locking point to the next. If the negative limit of the fine tuning range is exceeded the oscillator jumps in the negative direction; it jumps in a positive direction when the positive limit is exceeded.

Another method of jumping the frequency suggested by the designer is to use frequency dividers between the voltage controlled oscillator and the phase detector inputs and to arrange switchable taps so that the phases at the inputs of the detector can be varied by increments of $\pm\pi/2$ radians or less. The sign of the phase increment variation then determines the direction of frequency jumping. This method can be utilized by means of digital switching.

The action of the feedback in the system is not only to lock the period (hence frequency) of the oscillator to the delay period of the delay line, but also to reduce the f.m. noise sidebands of the oscillator within a band of $\pm$ 2 or 3kHz. The spectral purity of the oscillator is said to be 'appreciably' better when locked than when free running. The Mullard DL14 PAL delay line used is claimed to give a temperature stability less than an order down on that of a quartz crystal. Over a 3 to 6MHz frequency range the insertion loss varies by less than 6dB, giving a basic 2:1 frequency range of operation.

The use of standard frequency dividers, either in the voltage controlled oscillator output or between the v.c.o. and the delay line, makes it possible for a system using the DL14 line to provide any frequency at the output.

A practical delay-stabilized oscillator constructed on this system uses standard digital and linear i.cs operating from a single 5V stabilized supply and contains no inductors.

One interesting point to which the designer draws attention is that the 15.625kHz frequency spacing of the locking points for a PAL delay line, when multiplied by 4/5, gives 12.5kHz, which is a channel spacing in mobile radio frequency allocations.

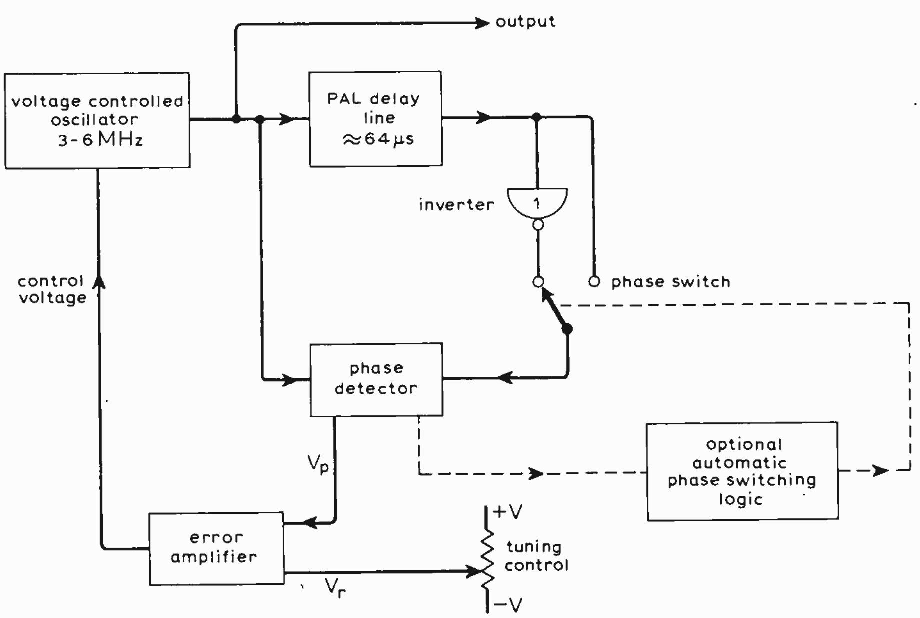


Fig. 1. Block diagram of the delay-stabilized oscillator.

Correction

Some errors occurred in the capacitor list in the article 'Hand Portable Transceiver' by D. A. Tong (April, p.160). Capacitors shown with the unit μ (micro) should be in picofarads except capacitors 48, 49, 57, 82, 85 which are in nanofarads. All electrolytics and capacitors 8, 10, 16, 31, 35, 50, 54, 62, 84, 88, 90 and 91 are as shown in the article. The charging current for the Deacs is 90mA.

Using Junction F.E.Ts

by L. Unsworth, B.Sc.

When read in conjunction with an earlier article[1] this contribution explains, in easy terms, how to design circuits with junction f.e.ts. The article contains nothing new but the equivalent circuits and mathematics have been reduced to their simplest possible form.

In the August 1971 issue of *Wireless World* J. Jenkins regrets that a lack of information on the theory and applications of f.e.ts has, to a certain extent, limited their use in circuit design. He then proceeds to help rectify the situation and I would now like to try and reinforce his efforts. Biasing is not covered since this has already been discussed in some detail.[1,2]

A junction f.e.t. (n-channel) consists of a bar of n-type silicon sandwiched between

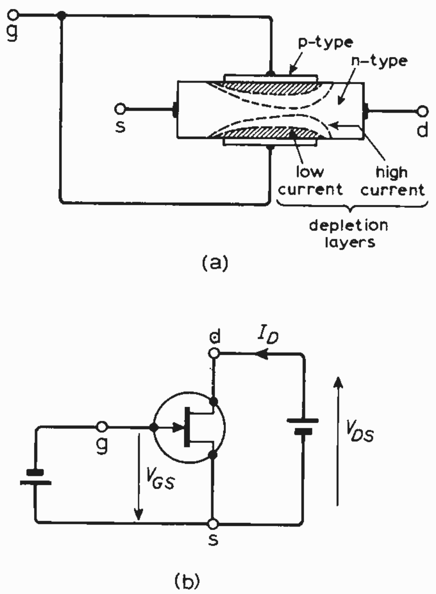


Fig. 1. (a) A bar of n-type silicon sandwiched between two p-type layers together form an n-channel f.e.t: (b) The polarity of power supply and bias voltages when using an n-channel device.

two regions of p-type silicon [Fig. 1(a)]. The n-type bar forms the drain and source regions and the two p-type regions form the gate. For low values of drain-source voltage, V_{DS}, when zero bias is applied between gate and source ($V_{GS} = 0$), the bar of n-type silicon acts as a resistor. As V_{DS} is increased, the drain current, I_D, increases and produces a potential difference along the bar (channel). This has the effect of reverse-biasing

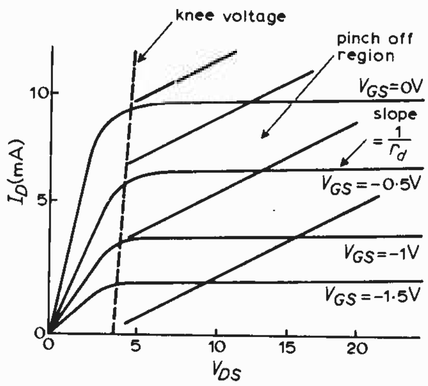


Fig. 2. A family of curves which describes the performance of a typical f.e.t.

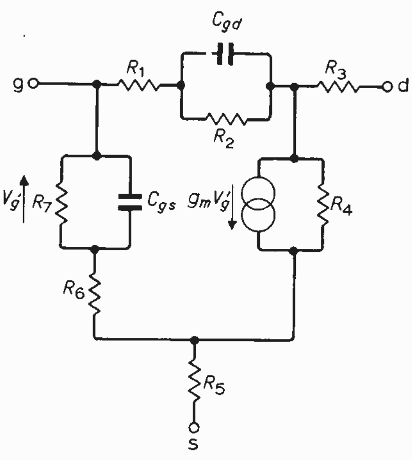


Fig. 3. General equivalent circuit.

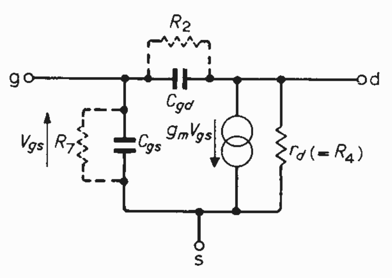


Fig. 4. The equivalent circuit for use at medium-frequencies.

the gate-channel junction and creates a depletion layer which extends into the bulk of the material reducing the effective cross-sectional area of the channel. The wider parts of the depletion layers occur in the region of greatest reverse-bias potential (near the drain), since the potential drops along the length of the channel. Increasing V_{DS} further results in a point being reached where no increase in drain current is produced because the depletion layers almost exactly balance the increase in V_{DS}. The device is then said to be 'pinched-off'. If the gate-source voltage, V_{GS}, is made negative, the device becomes 'pinched-off' at lower values of I_D. A family of curves similar to those in Fig. 2 can be plotted. The saturation value of I_D for $V_{GS} = 0V$ is given the symbol I_{DSS}. The value of V_{GS} (for a given V_{DS}) which will reduce the value of I_D to 1nA is the 'pinch-off' voltage V_P.

Since the gate is isolated from the channel by reverse-biased junctions, the gate current, under normal conditions, is very small (for some types <10pA). Also since there are no junctions obstructing the flow of I_D, the noise in the device tends to be limited to the thermal noise of the channel (at medium frequencies).

The equivalent circuit of a junction f.e.t. is given in Fig. 3. Resistors R_2 and R_7 represent the resistance of the reverse-biased junctions; R_1 and R_6 represent the losses in the device which become noticeable at very high frequencies; R_3 and R_5 represent the bulk and parasitic resistance between the active regions of the device and the terminals; and R_4 is the incremental output-resistance (r_d). Capacitors C_{gs} and C_{gd} represent the capacitance of the reverse-biased junctions.

For all but the highest frequencies the general equivalent circuit of Fig. 3 can be simplified as in Fig. 4. At all but the lowest frequencies R_2 and R_7 are negligible.

For low-frequency applications the equivalent circuit of Fig. 5 suffices. In practice the signal source resistance, R_s, would be in parallel with R_{gs} so that for normal values of R_s (<100MΩ) the effect of R_{dg} can be neglected since $R_{dg} \approx 10^{12}\Omega$ and will cause negligible feedback from drain to gate. The mutual conductance, gm, of the device and the gate-source signal voltage, V_{gs}, determine the value of the equivalent current generator.

Common source amplifier design

The circuit of a common source amplifier is given in Fig. 6(a) and the a.c. equivalent circuit is shown in Fig. 6(b). Resistance r_d is neglected since $R_L \ll r_d$. We are assuming a medium frequency design therefore C_{gs} and C_{gd} have been neglected. The output voltage, e_o, is determined conventionally by multiplying the signal current, gmV_{gs}, and the load resistance:

$$e_o = -gm V_{gs} R_L$$

and since:

$$V_{gs} = e_{in}$$

the voltage gain:

$$A = \frac{e_o}{e_{in}} = -gm R_L \quad (1)$$

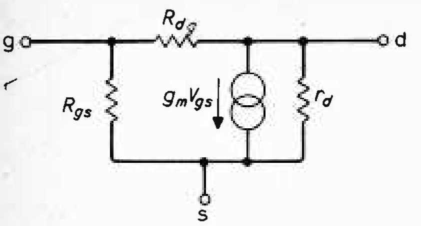


Fig. 5. For low-frequencies this equivalent circuit suffices (when junction capacitance can be ignored).

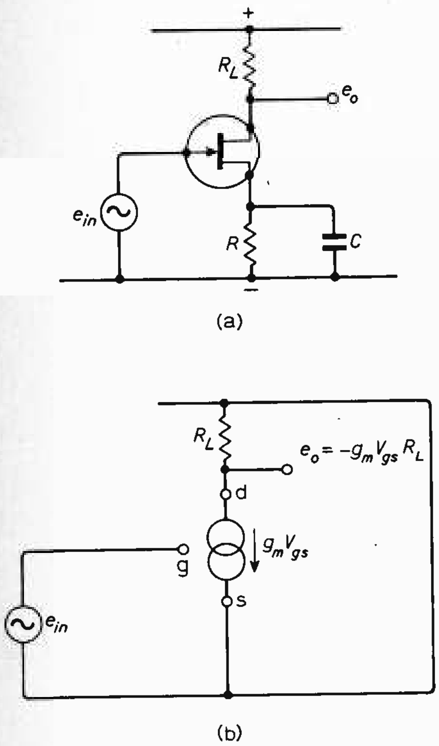


Fig. 6. (a) Common source amplifier and (b) its equivalent circuit.

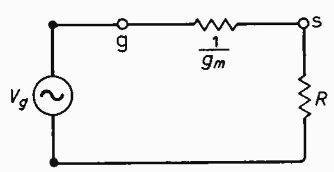


Fig. 7. At the source of the device the gm can be represented as a resistor with a value of 1/gm kΩ.

The negative sign simply indicates the 180° phase shift between input and output. In the case where R is not decoupled the following procedure may be adopted. Looking in at the source terminal, the gm of the device may be represented by a resistor with value $1/gm$ kΩ (if gm is in mA/V), between gate and source. At the source terminal this appears as in Fig. 7. The gate voltage appears across the resistors $1/gm$ and R. V_{gs} is now given by:

$$V_{gs} = \frac{1/gm}{(1/gm)+R} \times V_g$$

∴ the value of the current generator is:

$$gmV_{gs} = \frac{V_g}{(1/gm)+R}$$

The output voltage, e_o, is still given by $gmV_{gs}R_L$ and so is equal to:

$$\frac{-V_g}{(1/gm)+R} \times R_L$$

and $$A = \frac{e_o}{V_g} = \frac{-R_L}{(1/gm)+R} \qquad (2)$$

If equations 1 and 2 are compared, it can be seen that the effective gm of the device is given by:

$$gm' = \frac{1}{(1/gm)+R} \qquad (3)$$

Thus the gain of a common source f.e.t. stage with an undecoupled source resistor is given by:

$$A = -gm'R_L \qquad (4)$$

In order to decouple effectively, the reactance of C should be approximately one-tenth that of the parallel combination of $1/gm$ and R.

When higher frequencies are approached, the effects of C_{gd} and C_{gs} must be considered (Fig. 8). If the signal source resistance, R_s was zero then the upper frequency limit would be determined by the losses in the f.e.t., but in practical circumstances the limit is set by R_s, C_{gs} and C_{gd}. The signal voltage across C_{gd} at low frequencies, i.e. when $V_{gs} = e_{in}$, is given by $(e_{in} + Ae_{in})$ since the gate swings positive by an amount e_{in} while the drain swings negative by an amount Ae_{in}. Therefore the current through C_{gd}:

$$i_f = \frac{e_{in}(1+A)}{X_{C_{gd}}} \qquad (5)$$

where $X_{C_{gd}}$ is the reactance of C_{gd}.

The current through C_{gs}:

$$i = \frac{e_{in}}{X_{C_{gs}}}.$$

If the gate current is assumed to be zero, which is a reasonable assumption:

$$i_{in} = i_f + i$$
$$= \frac{e_{in}(1+A)}{X_{C_{gd}}} + \frac{e_{in}}{X_{C_{gs}}}$$
$$\therefore\ Y_{in} = \frac{i_{in}}{e_{in}} = \frac{1+A}{X_{C_{gd}}} + \frac{1}{X_{C_{gs}}}.$$

An equivalent circuit of the input circuit at high frequencies is given in Fig. 9.

The input signal is thus attenuated at high frequencies by a capacitance given by $C_{in} = C_{gs}+(1+A)C_{gd}$ so that the gate-source signal voltage is given by:

$$V_{gs} = \frac{e_{in}}{1+j(f/f_0)} \text{ where } f_0 = \frac{1}{2\pi C_{in}R_s}.$$

Therefore the stage amplification is now:

$$\frac{A}{1+(f/f_0)^2}.$$

Worked example

The f.e.t. of Fig. 10 has the following parameters: C_{gd} = 10pF, C_{gs} = 20pF, gm = 1.5mA/V and r_d = 50kΩ. If R_L = 50kΩ find the mid-band gain and 3dB cut-off frequency when R_S = 25kΩ. The stage gain:

$$A = -gm\frac{(R_L \cdot r_d)}{R_L + r_d}$$
$$= -1.5 \times 25 = -37.5.$$

The effective input capacitance:

$$C_{in} = C_{gs}+(1+A)C_{gd}$$
$$= 20+(38.5\times 10) = 405\text{pF}.$$
$$\therefore\ f_0 = \frac{1}{2\pi R_S \cdot C_{in}} = 16\text{kHz}.$$

Ways of reducing input capacitance: it is evident that the biggest contribution to C_{in} is the term $(1+A)C_{gd}$. In order to reduce its value, the voltage swing at the drain must be reduced [see Fig. 8(b) and equation (5)]. However, this implies an undesirable reduction in stage gain. Other possibilities consist of neutralizing the effect of C_{gd} by feeding back a current equal in magnitude but opposite in sense to i_f, or perhaps by tuning C_{gd} out by connecting an inductor in parallel with it. The last two methods are unsatisfactory because firstly they are only effective over a very narrow band width and secondly C_{gd} is bias-dependent and it becomes very difficult to maintain effective neutralization.

The first method would, initially, appear to be unsatisfactory also, since voltage gain is sacrificed. However, if a bipolar transistor is combined with the f.e.t., this disadvantage can be overcome. From Fig. 11(a) and (b) the voltage gain of the f.e.t. stage is given by:

$$A = -gm_1r_e \text{ (for } r_e \ll R_d \ll r_d)$$

where gm_1 = mutual conductance of Tr_1

r_e = emitter 'resistance' of Tr_2.

Voltage gain, A, will be much less than one because r_e is small and, since $r_e \ll R_d$, the

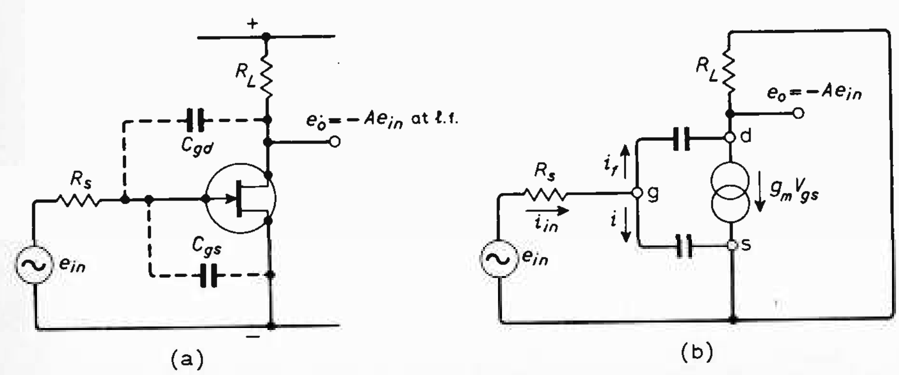


Fig. 8. The junction capacitances C_{gd} and C_{gs} limit the frequency response of the device.

current $gm\,V_{gs}$ flows almost entirely through r_e. Tr_2 is a common base stage and, since this current will then flow through R_L, the output voltage will then be given by:

$$e_o = gm_1 V_{gs} R_L.$$

The overall mid-band gain (when $e_{in} = V_{gs}$) will therefore be:

$$\frac{e_o}{e_{in}} = gm_1 R_L.$$

The 3dB cut-off frequency will be determined by $(C_{gs}+C_{gd})$ and the source resistance.

Common drain amplifier

A basic common drain amplifier (source follower) is given in Fig. 12(a) while Fig. 12(b) and (c) show the input and output equivalent circuits. From Fig. 12(c):

$$e_o = \frac{R_L}{(1/gm)+R_L} V_g$$

$$\therefore \frac{e_o}{V_g} = \frac{gmR_L}{1+gmR_L}.$$

At low frequency, $e_{in} = V_g$ and for $R_s \ll R_{in}$, the stage amplification

$$A = \frac{e_o}{e_{in}} = \frac{gmR_L}{1+gmR_L}$$

$$\approx 1 \text{ for } gmR_L \gg 1.$$

Improved circuit using bootstrapped drain: Resistor R_d (Fig. 13) is chosen so that V_{DS} does not fall below the knee voltage.

Let A = amplification of the f.e.t. stage

A' = amplification of the bipolar stage

then since $e_o = AA'e_{in}$ the voltage swing across C_{gd} is reduced to $(1-AA')\,e_{in}$ and hence the input capacitance will be:

$$C_{in} = (1-A)C_{gs}+(1-AA')C_{gd} \quad (6)$$

Worked example

The f.e.t. of Fig. 14 has the following parameters: gm = 2mA/V, C_{gd} = 5pF, C_{gs} = 10pF. The bipolar transistor has a β of 100. Calculate the stage gain and the effective input capacitance.

Let A = voltage gain of the f.e.t. stage

A' = voltage gain of the bipolar stage

$$A = \frac{gm_1 R_L}{1+gm_1 R_L} = \frac{2\times 5}{1+(2\times 5)} = \frac{10}{11}.$$

The emitter 'resistance' of Tr_2 is given by:

$$r_e = \frac{25}{i_e(\text{mA})} = \frac{25}{1.5}\,\Omega$$

$$\therefore gm_2 = \frac{1}{r_e} = \frac{1.5}{25}\text{ A/V} = 60\text{mA/V}$$

$$\therefore A' = \frac{60\times 10}{1+(60\times 10)} = \frac{600}{601} \approx 1$$

$$\therefore \text{the total gain} = \frac{10}{11}\times 1 = \frac{10}{11}$$

From (6) $$C_{in} = \left(1-\frac{10}{11}\right)\times 10+\left(1-\frac{10}{11}\right)5$$

$$= 1.35\text{pF}.$$

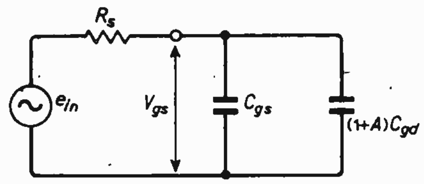


Fig. 9. The input signal to a common source amplifier is attenuated at high frequencies by the term $C_{gs}+(1+A)C_{gd}$.

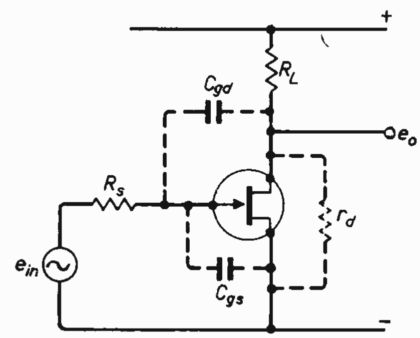


Fig. 10. The circuit used as an example in the text.

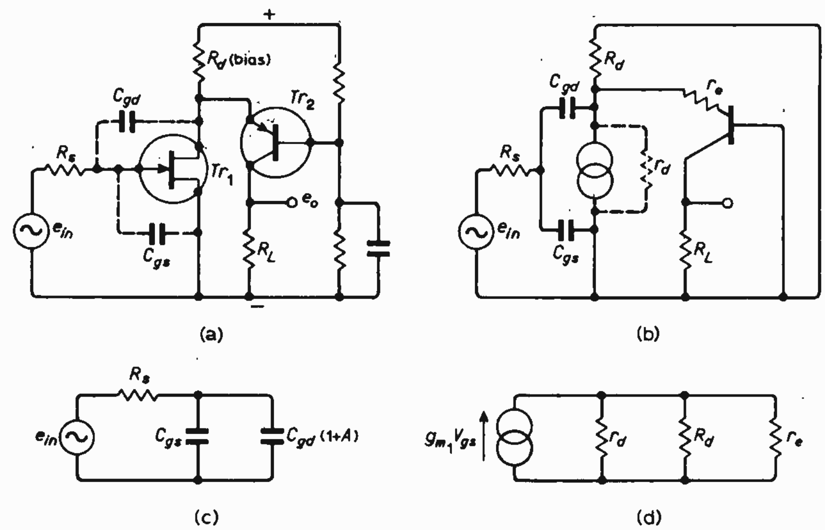


Fig. 11. Adding a bipolar transistor to reduce the term $(1+A)C_{gd}$. (a) The circuit; (b) equivalent circuit; (c) the input circuit and (d) the output circuit.

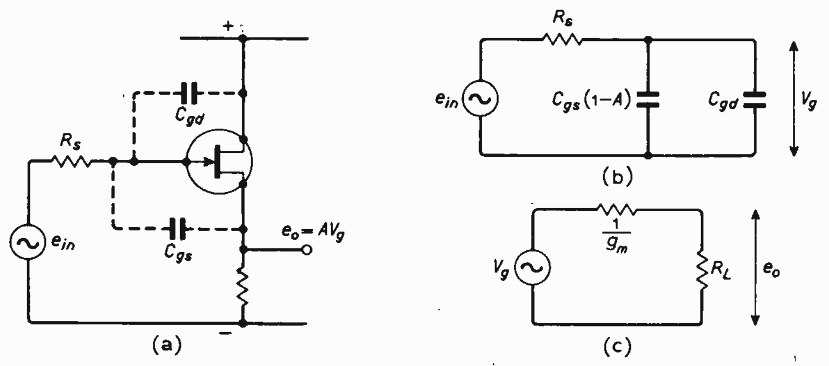


Fig. 12. (a) Common drain amplifier; (b) input equivalent circuit; (c) output equivalent circuit.

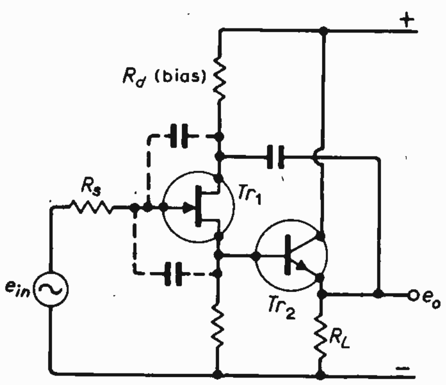


Fig. 13. Improved circuit using bootstrapped drain.

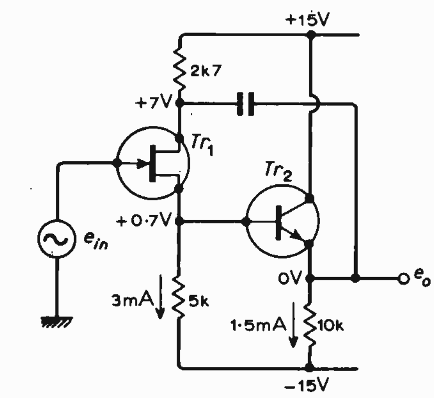


Fig. 14. A worked example of Fig. 13.

The circuit of Fig. 16 illustrates another example of how a bipolar transistor can improve the effective parameters of the f.e.t. First, however, may I remind readers of the simple low-frequency equivalent circuit of the bipolar transistor. In common emitter mode, the transistor may be represented as in Fig. 15. Assuming $\beta r_e \gg R_d$, then:

$$I = gm_1 V_{gs} + gm_2 V_{be}$$
$$= gm_1 V_{gs} + gm_2 R_d i_d$$
$$= gm_1 V_{gs} + gm_2 R_d gm_1 V_{gs} \quad (7)$$
$$\therefore e_o = gm_1 V_{gs} R_L + gm_1 gm_2 R_d R_L V_{gs} \quad (8)$$

Now:

$$V_{gs} = e_{in} - e_o \quad \therefore e_{in} = V_{gs} + e_o$$
$$e_{in} = V_{gs} + gm_1 V_{gs} R_L + gm_1 gm_2 R_d R_L V_{gs} \quad (9)$$

From (8) and (9):

$$\frac{e_o}{e_{in}} = \frac{gm_1 V_{gs} R_L + gm_1 gm_2 R_L R_d V_{gs}}{V_{gs} + gm_1 V_{gs} R_L + gm_1 gm_2 R_L R_d V_{gs}}$$

$$\therefore A = \frac{(gm_1 R_L + gm_1 gm_2 R_L R_d)}{1 + (gm_1 R_L + gm_1 gm_2 R_L R_d)} \quad (10)$$

At the output we have the conditions illustrated in Fig. 17. R_o is the output resistance of the circuit. Fig. 17(b) shows the condition for $R_L = 0$.

Now:

$$R_o = \frac{Ae_{in}}{I_{sc}} = \frac{Ae_{in}}{gm_1 V_{gs} + gm_1 gm_2 R_d V_{gs}}$$

(from (7)). Also $V_{gs} = e_{in} - e_o$, but $e_o = 0$.

$$\therefore R_o = \frac{Ae_{in}}{gm_1 e_{in} + gm_1 gm_2 R_d e_{in}}$$

and assuming $A = 1$:

$$R_o = \frac{1}{gm_1(1 + gm_2 R_d)} \quad (11)$$

The equation of an f.e.t.

For many practical devices, the drain current in the pinch-off region is related to the gate-voltage by:

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

where V_{gs} = gate-source bias voltage

V_p = pinch-off voltage.

Example

In the circuit of Fig. 16 the f.e.t. equation is given by:

$$I_d = 10\left(1 - \frac{V_{gs}}{4}\right)^2 \text{mA} \quad (12)$$

Assuming $I_b \ll I_d$ and given $V_{be} = 600$mV, $C_{gd} = 3$pF, $C_{gs} = 5$pF, $R_d = 600\Omega$ and

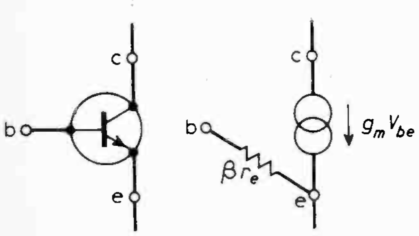


Fig. 15. Bipolar transistor and equivalent circuit.

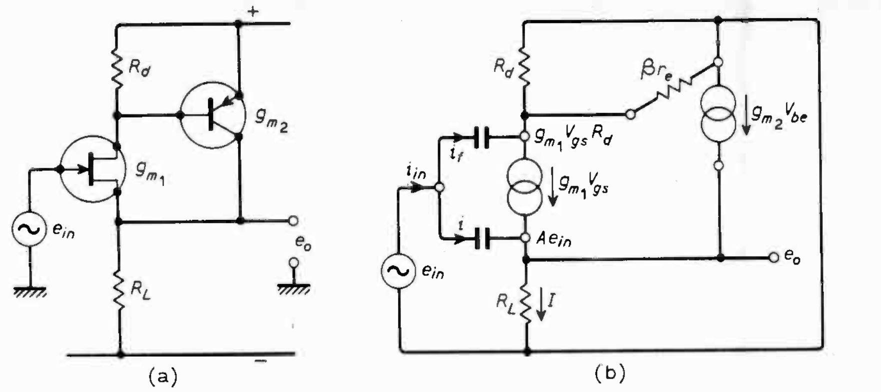


Fig. 16. Another circuit which shows how a bipolar transistor can be used to improve the basic f.e.t. circuit.

$R_L = 10\text{k}\Omega$, calculate (i) I_d, (ii) V_{gs}, (iii) I_c, (iv) voltage amplification, (v) R_o, (vi) C_{in}.

i. $$I_d = \frac{0.6}{600} = 1\text{mA}.$$

ii. From (12):

$$1 = 10\left(1 - \frac{V_{gs}}{4}\right)^2$$
$$\therefore V_{gs}^2 - 8V_{gs} + 14.4 = 0$$

whence $V_{gs} = 5.26$V or 2.74V.

Since $V_p = 4$V it must be assumed that $V_{gs} = 2.74$V.

iii. $$V_g = 0\text{V}$$

$\therefore$ the voltage across the source resistor $= 9 + 2.74 = 11.74$V

$$\therefore I_c = \frac{11.74}{10} - I_d = 0.174\text{mA}.$$

iv. gm_1 = mutual conductance of the f.e.t.

= rate of change of I_d with V_{gs}

$$\therefore gm_1 = \frac{dI_d}{dV_{gs}} = 20\left(1 - \frac{V_{gs}}{4}\right) \times \frac{-1}{4}$$
$$= -20\left(1 - \frac{2.74}{4}\right) \times \frac{1}{4}$$
$$= -1.57\text{mA/V}$$

(i.e. I_d increases as V_{gs} decreases)

gm_2 = mutual conductance of the bipolar transistor $= \frac{1}{r_e}$.

Now:

$$r_e = \frac{25}{I_c(\text{mA})}, \text{ assuming } I_c = I_e$$

$$\therefore gm_2 = \frac{0.174}{25} A/V = 6.95\text{mA/V}$$

$\therefore$ from (10):

$$A = \frac{(1.57 \times 10) + (1.57 \times 6.95 \times 10 \times 0.6)}{1 + (1.57 \times 10) + (1.57 \times 6.95 \times 10 \times 0.6)}$$
$$= 0.985.$$

v. From (11):

$$R_o = \frac{1}{1.57(1 + 6.95 \times 0.6)} = 124\Omega.$$

vi. From Fig. 16(b):

$$i_f = \frac{e_{in} + gm_1 V_{gs} R_d}{X_{C_{gd}}}.$$

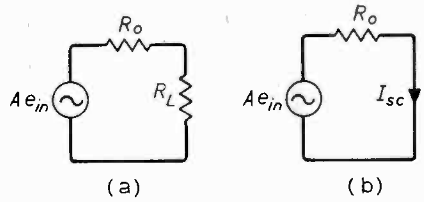


Fig. 17. Output equivalent circuits; (a) with load and (b) short circuit conditions.

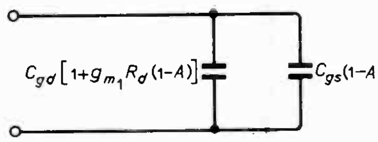


Fig. 18. Input equivalent circuit of Fig. 16.

Now:

$$V_{gs} = e_{in} - e_o = e_{in} - Ae_{in}$$
$$\therefore i_f = \frac{e_{in} + gm_1 R_d(1-A)e_{in}}{X_{C_{gd}}}$$

and $$i = \frac{e_{in}(1-A)}{X_{C_{gs}}}$$

$$i_{in} = i_f + i \text{ (assuming } i_g = 0)$$
$$\therefore Y_{in} = \frac{i_{in}}{e_{in}} = \frac{1 + gm_1 R_d(1-A)}{X_{C_{gd}}} + \frac{(1-A)}{X_{C_{gs}}}.$$

The input thus appears as in Fig. 18.

$$\therefore C_{in} = C_{gd}(1 + 1.57 \times 0.6 \times 0.015) + C_{gs} \times 0.015$$
$$\therefore C_{in} = 8.125\text{pF}.$$

As I pointed out at the beginning, the purpose of this article is to try and throw a little more light on the subject of f.e.ts and I purposely began with basic ideas in order to try and achieve this aim. The mathematics and equivalent circuits used were kept as simple as possible in order that the article should communicate with the maximum number of readers. It is hoped that any doubts or misunderstandings concerning basic f.e.t. circuits will thus have been dispelled.

References

1. 'Ten Practical F.E.T. Source-follower Circuits', J. O. M. Jenkins, *Wireless World*, August 1971, pp. 366–7.
2. 'Dual-trace Oscilloscope Unit', W. T. Cocking, *Wireless World*, September 1971, pp. 421–424.

Letters to the Editor

The Editor does not necessarily endorse opinions expressed by his correspondents

Amateur record?

Communication records still occur, although some are more unusual than others. I would like to record one which may, perhaps, never be broken. It is certainly unusual.

In the morning of 2nd March, when conditions were poor, perfectly clear phone contact was established between G. R. M. Garratt (G5CS, Esher, Surrey), E. J. Allan (GM5NW, Dundee, Scotland), myself (G6DW, Dorking, Surrey) and W. A. S. Butement (VK3AD, Melbourne, Australia). In VK3AD's shack were G. S. Samways (VK3OG), R. H. Cunningham (VK3ML), A. Holst (VK3OH) and L. Windom (W8GZ) who all spoke during the contact.

GM5NW was licensed in 1922 and G5CS, myself, and VK3AD in 1923. VK3AD's friends were licensed in 1922, 1928, 1914 and 1919 respectively—some in the United Kingdom.

The eight persons therefore, who spoke together over the air on this occasion, had been licensed for a combined period (excluding the war) of 402 years.

D. H. Johnson,
Dorking,
Surrey.

Dr W. A. S. Butement, who was a member of Watson-Watt's original radar team, went to Australia in 1949 as chief scientist in the Department of Supply. L. Windom was the originator of the off-centre-fed aerial bearing his name.—Ed.

Baxandall speaker

Readers who built the Baxandall loudspeaker design, as described in the August and September 1968 issues, may be interested to know that it is possible to obtain an extended bass response from this little speaker. As pointed out by Mr Baxandall, the low-frequency response of the speaker is similar to that of a second-order high-pass filter, as the equivalent circuit is identical to such a filter. By choosing the Q-value of the circuit at 0.78 the two poles of this filter can be made to coincide with two of the poles of a fourth-order ± 1dB Tchebyshev high-pass filter. (Fig. 1.)

The interesting feature is that this filter has a cut-off frequency of 0.52 times the resonant frequency of the original second-order system. Applying this idea to the Baxandall speaker simply means adding a filter section to the amplifier. As the speaker resonates at about 100Hz, this section is part of a fourth-order Tchebyshev filter with a 52Hz cut-off frequency (Fig. 2). This filter section can be designed using Mr Bronzite's tables (March 1970). Only the first section, for which design figures are given under D_1, is needed (Fig. 3).

The improvement is in fact brought about by administering a control amount of bass lift to a maximum of some 9dB at 55Hz. At normal listening levels, using a 10-W amplifier, no trouble with overload was experienced even on programme material with fairly heavy bass content. The improvement in the bass response was considerable. Thus it would seem that an improvement of a whole octave can be obtained, making this scheme well worth while to the owners of this speaker.

T. Magchielse,
Delft,
Holland.

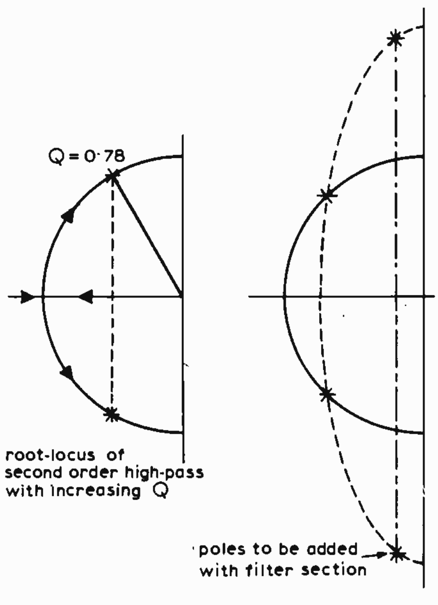


Fig. 1.

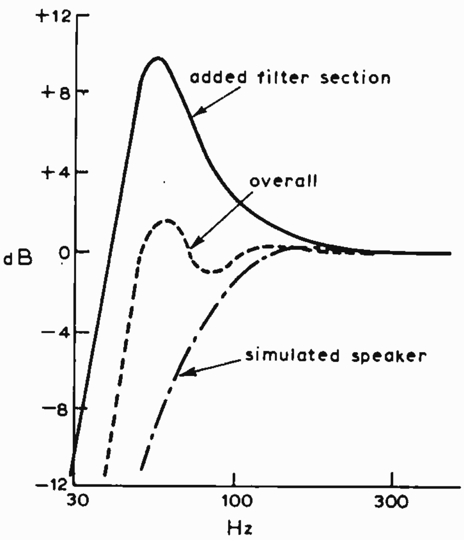


Fig. 2.

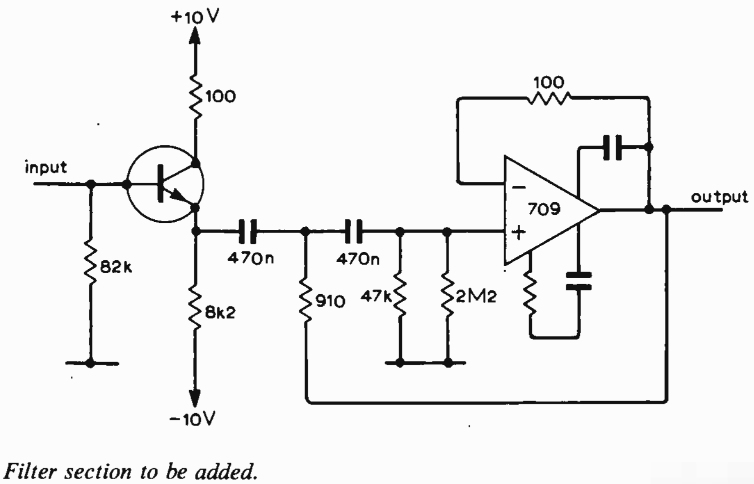


Fig. 3. Filter section to be added.

White-noise generator

There are a few points raised by the article 'White-noise Generator' by H. R. Beastall in your March issue, which invite comment. The author does not mention that his circuit is capable of generating a noise spectral density of great precision. The binary noise signal at the input terminals of his low-pass filter has a mean intensity given by the following expression:

$$S(f) = 2E^2T_s\left[\frac{\sin(\pi f T_s)}{\pi f T_s}\right]^2 \text{volts}^2 \text{ per Hz},$$

where $2E$ is the peak-to-peak amplitude and T_s is the period of the 900kHz clock. If the low-pass filter has unit amplification, the output noise intensity will also be $S(f)$ and the mode of generation defines it more precisely than does the author's suggested method of measuring it via $e_s^2/\Delta f$. Further, the filter should have a slightly rising characteristic if white noise is required[1,2].

The Appendix dealing with noise figure may mislead readers into thinking that a measurement of this kind, with 600Ω source impedance, has some particular significance. The estimate of the complete equivalent noise circuit of amplifiers is more important, particularly in this frequency range where source impedances of almost any value might be encountered[3]. Finally, I think the arithmetic is wrong. $4kTR_s\Delta f$ with the author's values is 4.8×10^{-13} volts squared per Hz.

H. Sutcliffe,
Professor of Electronic Engineering,
University of Salford.

References

1. H. Sutcliffe and G. H. Tomlinson: 'A Low-frequency Gaussian White-noise Generator', *Int. J. Control*, 1968, Vol. 8, No. 5, pp. 457–471.
2. H. Sutcliffe and K. F. Knott: 'Standard L.F. Noise Sources using Digital Techniques and their Application to the Measurement of Noise Spectra' *Radio and Electronic Engineer*, Vol. 40, No. 3, Sept. 1970, pp. 132–136.
3. E. A. Faulkner: 'Optimum Design of Low-noise Amplifiers', *Electronics Letters*, Vol. 2, No. 11, Nov. 1966, pp.426–7.

The author replies:

The great precision mentioned by Professor Sutcliffe involves great precision in measuring E and T_s. This can be compared with measuring the output voltage and the noise bandwidth. Obviously T_s can be measured digitally to a much greater accuracy than can the noise bandwidth. It is worth noting that consideration was given at the design stage to including a mean intensity meter in the instrument. This was to be a full-wave rectifier instrument driven, as Professor Sutcliffe suggests, from the binary noise signal at the input to the low-pass filter. The reason for this approach is that the meter current would be steady so that the meter 'jitter' obtained with most noise measurements is eliminated. Since the basic requirement was for a low-cost, simple instrument this idea was abandoned.

With regard to the slightly rising characteristic for the filter, I feel that Professor Sutcliffe is really splitting hairs. The envelope of the voltage spectrum follows a $(\sin x)/x$ form as indicated in Fig. 3 of the original article. The clock frequency is nominally 900kHz and the filter cut-off frequency is 50kHz.

Now $$\frac{\sin\left(\frac{50\pi}{900}\right)}{\left(\frac{50\pi}{900}\right)} = 0.995$$

so that the fall in amplitude is only 0.5% or 0.04dB and can be ignored in the present application.

Estimation of the equivalent noise circuit of an amplifier is important at the design stage and the instrument described is suitable for this. However, measurement of the noise figure of an amplifier with 600Ω source impedance has a particular significance in production testing components for 600-ohm matched communication systems. The noise generator was designed to make such measurements as part of the course-work leading to a Higher National Diploma in electrical and electronic engineering.

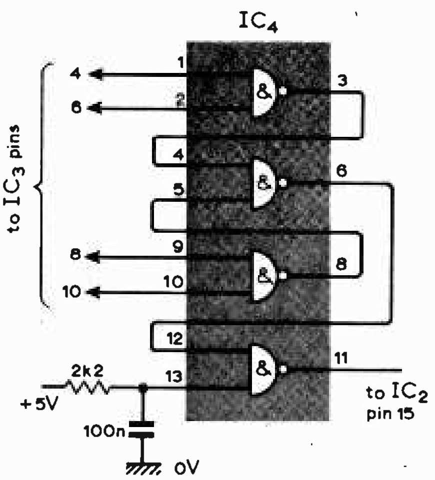
There is no mistake in the arithmetic. 4.8×10^{-13} volts squared per Hz = 0.48×10^{-6} millivolts squared per Hz. Since e_s was specified in millivolts, $4kTR_s\Delta f$ must be in millivolts squared.

Finally, I have had a number of enquiries regarding the SS–6–1032 shift register i.c. This is made by General Instrument Microelectronics. A plastic packaged version, the SS–7–1032, is now available. This is electrically identical but cheaper. Both devices are available from S.D.S. (Portsmouth) Ltd, Hilsea Industrial Estate, Portsmouth, Hants, PO3 5JW.

H. R. Beastall.

The digital white-noise generator in the March issue by H. R. Beastall is an excellent design, but I would like to suggest some improvements and simplifications.

The shift register with feedback, forming a chain code generator, has the disadvantage that a stable state in which all stages contain a logical '0' exists. Therefore if on switching on all stages initially contain a '0', the chain code generator is, so to speak, stuck in a groove and a string of '0s' will be generated. There are two standard methods of avoiding this:



Nicholas Lewis suggests that IC_4 should be rewired as shown.

(1) By means of an all '0' inhibitor, in which a decoder detects the all '0' condition and ensures that a '1' then appears in the first stage, but this is not possible in this case because only some of the outputs of the register stages are accessible.

(2) A '1' may be injected into one stage when switching on. For this an extra two-input NAND gate is required, but fortunately the exclusive OR gate can be rearranged to use one less gate. The circuit is shown below.

On switching on, the 100nF capacitor holds the output of its NAND gate at a '1', thus filling the shift register up with '1s'. Once the capacitor has charged up the circuit functions normally. The NAND gate may oscillate when the voltage across the capacitor reaches about 1.2V, but this should have no ill effects.

The removal of IC_5, coupled with the changing of R_1 to 500Ω wired to +5V instead of 0V would seem a worthwhile simplification. The filter is now driven from a source impedance of not greater than 125Ω, which should have a negligible effect on the filter characteristics. Also, for applications not requiring the greatest precision, IC_6 can be replaced by a transistor (e.g. BC107-8-9) wired as shown.

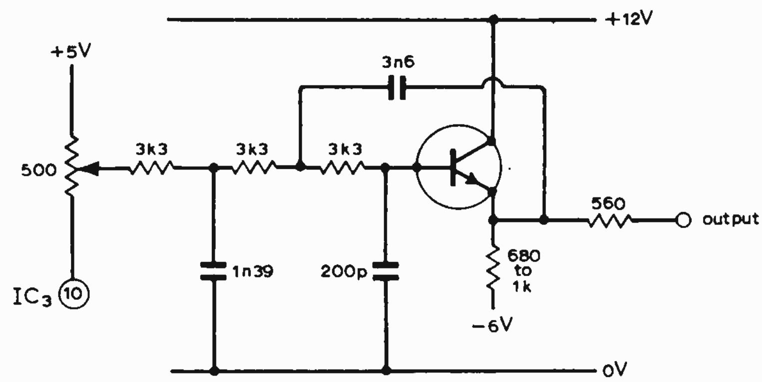
Finally, it is good practice to wire a 100nF capacitor across pins 7 and 14 of IC_3 or IC_4.

Nicholas Lewis (aged 17),
Chichester,
Sussex.

The author replies:

It is true that maximal-length sequence generators can exist indefinitely in the 'all zero' state. The chances of this happening are about one part in 2×10^9. The prototype noise generator has never started in this state. If a prospective constructor wishes to be sure, the scheme suggested by Mr Lewis can be used. Incidentally, during development, a badly placed test prod accidentally held one of the logic states constant for more than 31 clock periods and the generator was forced into the 'all zero' state. This can be very disconcerting as the output disappears and one is led to believe that the i.c. is damaged. Switching the supply, however, soon restores the output.

The output impedance of a t.t.l. gate is 150Ω in the '1' state and 16Ω in the '0' state. If IC_5 is removed and R_1 is changed to 500Ω the filter can be driven from a source



If great precision is not required Nicholas Lewis shows how a BC107 can replace IC_6.

impedance of about 162Ω. This is 5% of the value of the input resistor to the filter and is thus barely acceptable.

Use of a direct-coupled emitter follower for the filter amplifier will introduce considerable d.c. offset into the output. Furthermore, this will vary with the setting of R_1. The offset could be removed with a capacitor at the output. However, an original requirement was that the output spectrum should be flat within 1dB over the frequency range 20Hz to 20kHz and this would require an unpolarized capacitor of about 30μF.
H. R. Beastall.

Bootstrapping

Quite a lot has been written recently in your columns with regard to simple circuits using 'bootstrap' techniques: (P. M. Quilter Apr. '71, H. P. Walker, May and June '71, and various letters to the editor), and it may interest readers to hear of problems and solutions we experienced with similar circuits for our studio recording mixer.

We found that with a split collector load, the resistor to the collector, having positive feedback applied through it, was prone to hash (resistive microphony) due to the much increased impedance, and that some form of high-pass capacitance was needed to shunt unrequired frequencies to earth. This was particularly necessary with Mr Quilter's tone-control circuit, where long

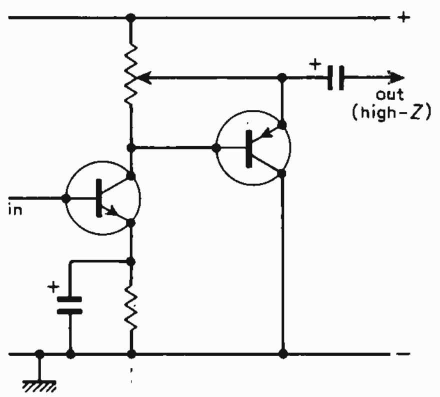


Fig. 1.

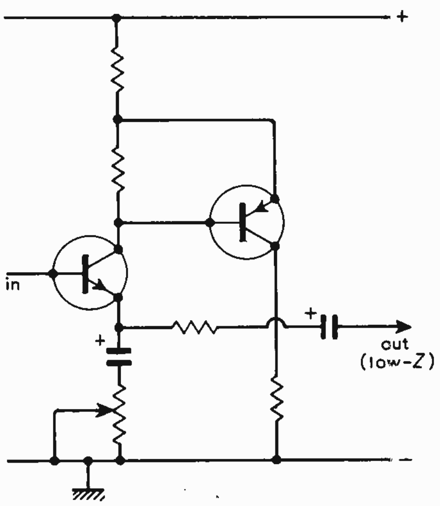


Fig. 2.

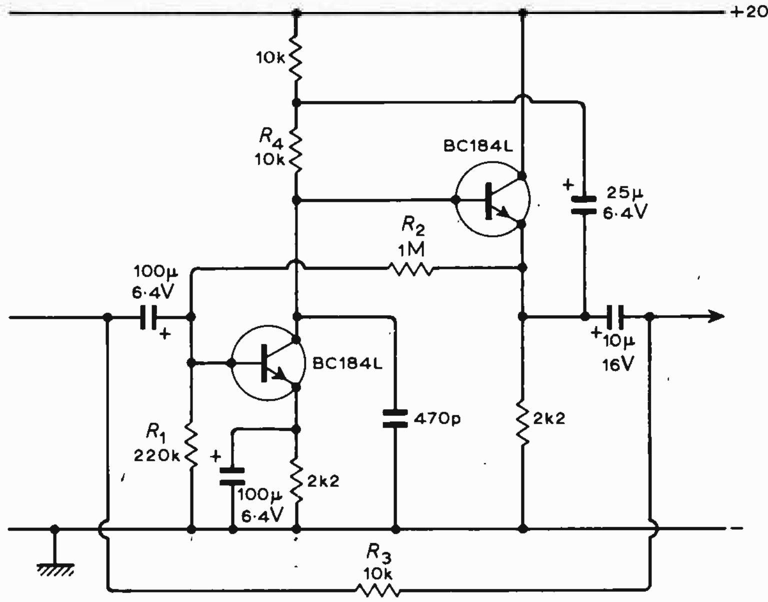


Fig. 3.

leads were used to the controls. Even a 470pF capacitor between input and output did not fully cure the hash pick-up, and so we hung a few hundred pF from Tr_1 collector to earth, with good results.

In our early experiments we managed to get a collosal power gain by a.c./d.c. bootstrapping, as in Fig. 1, by varying the ratio between the two halves of a split collector load; but we found associated problems. The output, used as a positive feedback bootstrap element, increased its own impedance, and would have required feeding into an even higher impedance to avoid heavy loading. The circuit also became unstable. We modified this circuit to Fig. 2, but found, strangely enough, that we could obtain a much greater s/n ratio without the bootstrapping, namely, with the emitter of the second transistor direct to rail. As a good s/n ratio was more important to us than enormous gains from one amplifying transistor, our experiments changed direction.

Readers with a thirst for large signal-to-noise ratios, well over the 100dB mark, may be interested in a modification to Mr Quilter's tone-control circuit, making it suitable for virtual earth applications. The circuit is in Fig. 3. There are two feedback paths. First, R_1 and R_2 provide d.c. feedback and also d.c. stability. These two values are kept high in order to keep noise at a minimum. Secondly, R_3 provides a.c. feedback, and is made not less than 10kΩ so that it does not have a shunting effect on R_4. It is also decoupled from d.c. by the input and output capacitors.

As one of your readers recently remarked, this whole subject of bootstrapping certainly needs more exploring and experimentation.
C. R. Cathles,
New Age Entertainments Ltd,
Epsom, Surrey.

The ASP circuit

I should like to congratulate Mr van der Walt (Letters, January '72) on his anti-symmetrical pair (ASP) circuit. It is both elegant and economical in components. My own recent measurements show that when built into a feedback tone control and compared to Mr Quilter's (April '71 issue), the two compare very favourably, with a slight advantage for Mr van der Walt's circuit because fewer components are needed. Distortion is very low with both circuits and signal handling capacity is good (at least 6V p-p output signal with 15V h.t. rail).

For the ASP circuit the measured gain (with no external negative feedback) was 960 times using a BC109 (I_c = 0.5mA) and an MM2721 (I_c = 2.5mA). Calculated gain for these currents is g_m/h_{oe1} = 910; very close to the calculated value. There was no trace of instability during the test.

Therefore I find myself in complete disagreement with Mr Harper (Letters March '72).

Mr. Harper's computer analysis gave some extraordinary numerical values, and his experiments provided strange results yet he does not explain why his analysis gives such different answers from the theory put forward by Mr van der Walt. It would be useful to know in what way the mathematical model used in the computer analysis differed from the one provided by Mr van der Walt. One remark particularly of Mr Harper's concerning Tr_2 deserves comment. This is an emitter follower, pure and simple, with a bootstrapped resistor R_2. Bootstrapping does not constitute positive feedback, its only effect is to raise the apparent value of R_2, so providing a higher

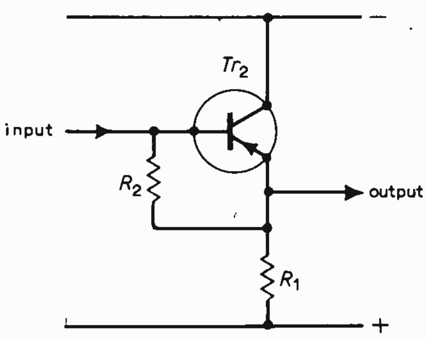


Fig. 4

collector load than if R_2 were connected to the positive rail.

Emitter followers do not usually oscillate on their own and the presence of Tr_1 makes no difference here. Oscillation when occurring in circuits of this kind is often due to inadequate h.t. decoupling at h.f. or some spurious feedback path.

Wow and flutter meter

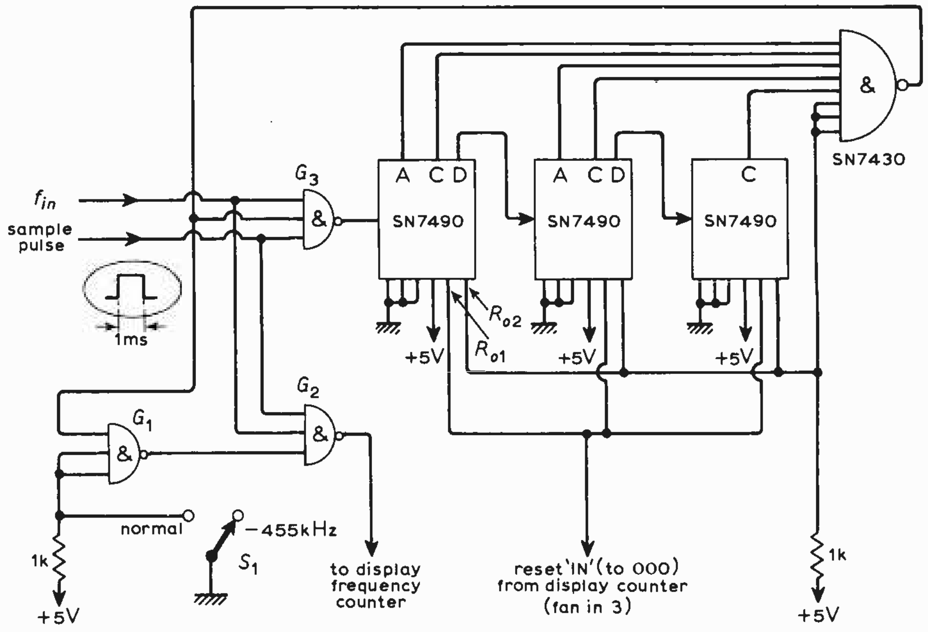
Referring now to the wow and flutter meter design by Mr Ockleshaw in December 1971. Pin 9 of IC_3 (MC4024P) is shown open circuit, but should be earthed. It is the earth rail for the second multivibrator, and though not used, it should be earthed to ensure substrate earthing and proper isolation. The 2.5kΩ variable resistor should be wire-wound.

Should the second multivibrator be needed, e.g. for providing a 3kHz test tone for tape recorder flutter measurements, the pin connections are as follows:

pin 8: output signal.
pin 9: earthed (as above).
pins 10, 11: connect a 68nF capacitor in series with a 1kΩ variable wire-wound resistor and join direct to pins 10, 11. The 1kΩ variable resistor is then a fine frequency control giving a range of frequency of 2 to 3.6kHz.
pin 12: input control voltage—connect to ground.
pin 13: feed via switch from the +4.5V supply lead.

The oscillator may be correctly set for frequency once the phase-locked loop has been set up (from an oscillator or a test record) by connecting the oscillator output from pin 8 directly to the input of the wow and flutter meter.

B. J. C. Burrows,
Ewelme,
Oxford.



D. J. Airey's circuit.

Displaying frequency digitally

Two mistakes are apparent in the circuit diagram accompanying the letter from J. A. Titus in the March issue, demonstrating an arrangement for subtracting 455kHz from a counter input.

First, one input of the counter sensing gate at the top should be connected to the C output of the third 7490 (The present arrangement counts to 255). Secondly, an inverter is required between the output of this NAND gate and the gates it feeds, otherwise the subtract 455 counter will never commence operation.

A useful extrapolation of this interesting technique would be to use cascaded programmable dividers of the 74190/1 type whose parallel entry facility enables division by any integer up to 10 or 15.

The subtraction counter could then be programmed to deduct any desired frequency before the main counter operated.

H. F. Lewis,
London W.5.

With reference to the letter from J. A. Titus on displaying frequency digitally in the March issue. I would like to point out an error in the circuit diagram, it will only invert the input signal.

I propose that the circuit should be as shown. This will only subtract 455kHz from the input frequency if the least significant decade of the frequency counter is equivalent to 1kHz, i.e. a sampling pulse width of 1ms.

To reduce the propagation delay of the sampling gating the NAND gate G_2 could form the sample gate of the frequency counter. With the switch as shown the experimenter has a direct read out of receiver frequency and a frequency counter for other uses.

D. J. Airey,
York.

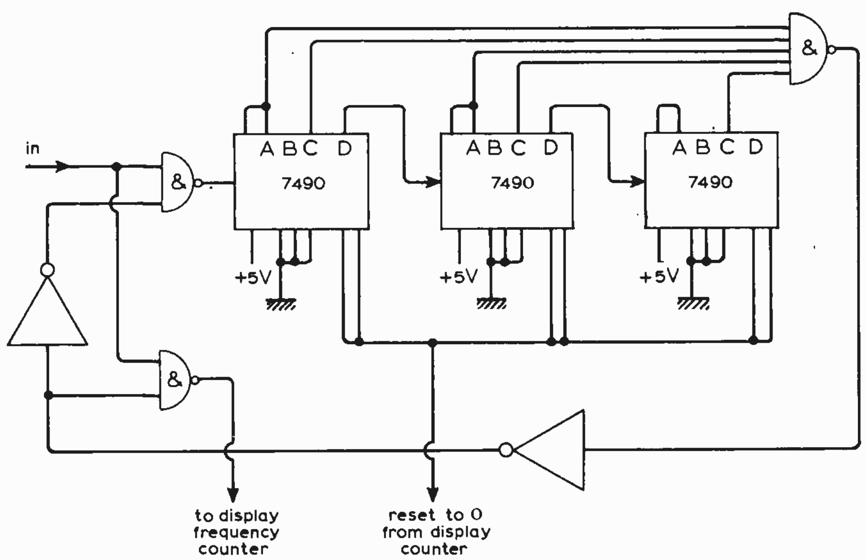


Diagram accompanying H. F. Lewis' letter.

Becker 100-W amplifier

Readers contemplating the building of the 100-W amplifier described in the February issue by Mr Becker may be interested in the following points:

(1) It is advisable to fuse *both* output lines since, if only one line is fused, someone will eventually short the unfused line to chassis (Parkinson's Law). The result is expensive.

(2) Parasitic trouble at the higher power levels has been experienced with this design (built from a Powertran kit). It was cured by coupling Tr_7 and Tr_8 collectors with a capacitor of approx. 4000pF. The capacitor can be conveniently mounted, on tags, between the collector tab mounting bolts.

(3) It may be advantageous to glue the biasing transistors Tr_9 and Tr_{10} to the output heat sink, rather than to the driver cases. With a kit containing 'matched' transistors the quiescent current, with the biasing transistors glued to the drivers, varied from 160mA at ambient temperature to over 400mA at 50°C.

A. G. Jones,
Porthcurno,
Penzance.

Sonex 72

Comment on the recent Skyway hotel audio exhibition

While Great Britain has always had a good reputation for the technical standard of her audio equipment the Sonex exhibition gave very little evidence of any real commercial vitality except in the manufacture of loudspeakers.

What is looked for is not a proliferation of 'budget' equipment but the emergence of reliable high-performance audio units. We readily admit the excellence of many manufactured products that have been showing for years and years. The shock lies in the contrast between these items (and even more recent ones) and the staggering variety of items of imported equipment.

Even with the recent currency revaluation the Japanese maintain a firm hold on the amplifier and tuner market in Britain, and there is a growing likelihood that they will dominate the cassette and reel-to-reel tape markets here.

Although some British loudspeakers are capable of very high-quality reproduction over the range 60Hz-15kHz only a few of the larger and expensive models (Cambridge Audio, IMF, Radford, Celestion) provide deep bass (down to 40Hz and below) with realistic balance. British direct radiator drive units are rather weak in power-handling capability and it is consequently rare for 100dB sound pressure level to be reached without system stress becoming apparent. Obviously, as signal bandwidth and dynamic range both increase, with ever improving recording techniques, wide-range low-distortion speakers capable of accurately reproducing large signal peaks (up to 120dB) will be demanded.

Such speakers have been available for some time in America, and a recent exhibition at the U.S.A. Trade Centre suggests that models will be finding their way into Britain before very long. Of course J. B. Lansing systems have been available here for the last two years.

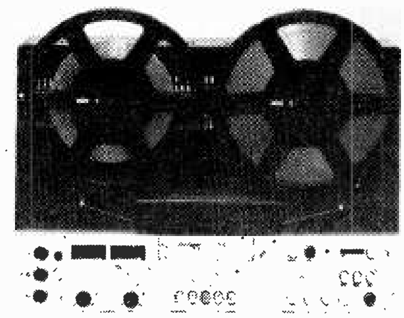
Progress seems to have been made toward a practical true four-channel disc. Quadraphonic disc releases so far made have been matrixed records, where four signals are linearly combined into two in one of a variety of ways. One of the problems in a disc system with groove modulation extending to 40 or 45kHz is the number of playings one can get before crosstalk and signal-to-noise ratio are degraded unacceptably. Development of the Shibata stylus by Matsushita has helped which, because of its curvature at the points of contact is less, claims to reduce wear through lowering pressure on each wall (see photograph on p.244), but the life still was not good enough. As we have reported earlier, RCA have been collaborating with Matsushita in developing their disc system and now RCA say discs can last for 100 playings with an ordinary stereo pickup using a conical stylus at 5g, while maintaining satisfactory crosstalk and signal-to-noise ratio. One contributory factor has been the use of a new record compound involving a multiple resin system containing anti static, lubricating stabilizer and other lubricants, which have double the life. Some changes have also been made in signal processing — more details later. RCA say they will have a disc available this month and will issue regular recordings in Autumn — for the same price as conventional discs. Demodulators and cartridges for the system come a bit expensive though.

Meanwhile the first U.K. company to issue matrixed quadrophonic records is Pye Records, beating both EMI and CBS to the post — because Sony SQ decoders have not come off the line when expected. Pye uses the Sansui QS system for which equipment is also available through Vernitron. QS system properties have been discussed in the January and February issues, but in the table on p. 56 we omitted to say that in mono playback any centre back sounds will be suppressed — as of course applies to SQ as well.

Once again we must draw attention to the poor sound isolation afforded by the Skyway hotel rooms. At times it was impossible to hear the programme material above the aircraft rumble. This was particularly bad on the top two floors even with the windows shut.

Magnetic cartridges

Decca's Mark 5 ffss head, called the London, weighs only 4g, has a very low sensitivity to stray magnetic fields and has a relatively high signal output — $7\frac{1}{2}$mV at 5cm/s. The 'secret' behind the design is the use of a new magnetic material introduced by Mullard, and the tempering of the armature by immersion in liquid nitrogen. All Decca cartridges give a remarkable openness, particularly on strings.



The Braun TG1000 tape recorder has an electronic speed change, three heads (two or four tracks optional) and tape tension is photoelectrically controlled.

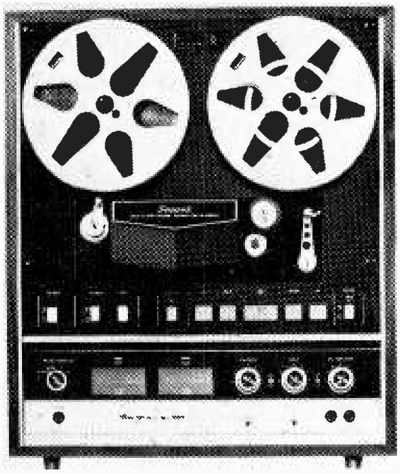
Shure introduced the M75ED type 2 cartridge in a programme which included an EMI quadraphonic record. The disc was decoded by a Sony SQ1000 and signals amplified and replayed over Rogers equipment. The M75E type 2 cartridge can be fitted with a new stylus to bring it up to the new standard.

The SP15 cartridge from Bang & Olufsen will fit a standard $\frac{1}{2}$in mount and weighs 5.5g. Performance is specified as 20Hz-30kHz $\pm$2.5dB with compliance of 30 $\times$ 10^{-6}cm/dyne and channel separation $<$20dB at 500Hz-10kHz. Channel difference is $<$1.5dB. Tracking pressure 1-1.5g.

Tape equipment

Acoustico Enterprises showed the Braun TG1000 tape recorder having signal-to-noise ratio at $7\frac{1}{2}$ i.p.s. of greater than 60dB and electronic speed change ($7\frac{1}{2}$, $3\frac{3}{4}$, and $1\frac{7}{8}$ i.p.s.). Tape tension is controlled photoelectrically. The unit has three heads (two or four tracks optional) and several extra facilities. Frequency response is 20Hz to 25kHz with wow and flutter quoted as less than 0.05%. Recommended retail price is over £200.

Sansui's SD5000 tape deck employs large-core recording and playback heads (20Hz-20kHz at $7\frac{1}{2}$ i.p.s.) and provides 60dB signal-to-noise ratio. Also included are auto reverse and remote control capability, cue button, and pause switch. Price £328.



Sansui's new SD5000 tape deck incorporating a delay relay circuit for control of tape tension.

The J.V.C. CD1667 cassette deck (from Denham & Morley) employs a noise reduction system different from the Dolby B circuit, and performance with chromium tape is claimed to be 30Hz-16kHz±3dB. Wow and flutter are given as 0.15% r.m.s. Price is £121 inc. p.t.

Four-channel record replay system

The J.V.C. discrete channel record (CD-4 system) is to be marketed by RCA to the extent that eventually all RCA's new recordings will be compatible with both stereo and discrete four-channel equipment. Denham & Morley are offering an introductory package costing about £100 comprising a four-channel demodulator, a 4MD1X four-channel cartridge fitted with the Shibata stylus and two CD4 records.

Loudspeakers

The DM2 loudspeaker from B & W, priced at £52 plus tax, employs three units. The bass driver is of Bextrene and incorporates a mechanical crossover to a 2 inch central dome. Electrical crossover is at 3.5kHz to a Celestion HF1300 and at 9kHz to a super tweeter. A unique feature of this speaker is the 'eighth wave acoustic line' (patent applied for) which is claimed to provide reduced cone excursion between 30 and 60Hz with increased acoustic output.

Lowther have introduced a new folded horn — the Super Acousta. Two PM6 or PM7 units are fitted side by side and provide a translational efficiency of between 20 and 27%.

The Ditton 66 from Rola Celestion, priced at £99, incorporates three drivers and an auxiliary bass radiator. This system claims low-frequency response to 16Hz and employs a new mid-range unit and their recent 'super tweeter'.

K.E.F. introduced two new miniatures, the Coda and the Cantor which incorporate improved versions of the T27 tweeter and seem to replace the Cresta and Celeste.

Judging by the source of the claims put forward by the manufacturers — '. . . as good or better than bigger loudspeakers — even fifty times bigger' — Servo-Sound speakers ought to be quite stunning. The claim made is that by the use of a patented form of feedback box resonances have been eliminated. Clearly motional feedback *is* used and so is a large measure of low-frequency compensation resulting in surprising amounts of bass. Nevertheless shortcomings in the single drive unit rather spoil the recipe.

Books Received

Thickfilm Microwave Circuits — An evaluation of materials and processes for thickfilm striplines at microwave frequencies by E. Anderson and S. T. Oleson is a report issued by Elektronikcentralen (the Danish Research Centre for Applied Electronics). The report (in English) describes an investigation of the basic materials and processes necessary in the manufacturing of thickfilm circuits for frequencies in the GHz range. The r.f. properties of these circuits are a function of a large number of variables (e.g. the substrate materials, the pastes used for the screening of the conductors and the mesh size used in the screening process) and in order to evaluate the importance of some of these variables, about 1200 measurements have been made on 150 samples in the frequency range of 1-10GHz, and the results recorded.

After an introduction and a description of patterns and materials the sections of the book are: measuring methods, measurement results and conclusions. Finally, a chapter on processes includes a method of etching developed to improve the 'edge' properties of thickfilm conductors. Pp.64. Price D.kr.50. Elektronikcentralen, Venlighedsvej 4, DK-2970 Hørsholm, Denmark.

Long Distance Television by Roger W. Bunney is a booklet describing the requirements for good quality reception from weak signals due to large distance between the transmitting mast and the receiver. The field strength delivered to a receiving site depends upon a number of factors — the frequency of the transmitter, the transmitter's effective radiating power, the height of the transmitting mast and the intervening terrain between the transmitter and the receiving site. Thus at some distance beyond the optical horizon it would at first seem that distant signals would be too weak to be of any use in providing viewable pictures. However, it is certainly possible to obtain reception of such transmissions at very considerable distances, often at high strengths. The mechanism by which these signals are propagated is discussed and also included are details of the various television transmission systems at present in use throughout the world. Pp.36. Price 50p. Weston Publishing, c/o 58 Triconderoga Gardens, Weston, Southampton SO2 9HD.

Communication Systems Analysis by P. B. Johns and T. R. Rowbotham bridges the gap between the rather oversimplified approach of many basic textbooks and the highly mathematical and specialized literature on communication theory. The contents begin with an introduction to the basic mathematics involved with signal and noise analysis. Thermal noise is considered in detail and so are analogue modulation methods and interference, echo and a.m. to p.m. distortion in f.m. systems, non linear distortion and finally digital modulation systems, including techniques involved with sampling, pulse code modulation and delta modulation. Many worked examples are included throughout the text and there are problems and answers set on each chapter. The practical nature of this work provides the basis for an approach to communications in second and third year undergraduate courses and will also be found useful as a reference work for practising communication and systems design engineers. Pp.207. Price £3.40 (cased), £2.20 (limp). The Butterworth Group, 88 Kingsway, London WC2B 6AB.

Newnes Radio Engineers Pocket Book: 14th edition has been completely revised and obsolete information dropped. This has made room for a considerable amount of new material relating to recent developments in the fields of radio and electronics. Apart from 'standard' information, tables and information are given which, although not directly relating to electronics, may be needed by the designer, hobbyist, student or serviceman. Pp.188. Price £1.20. The Butterworth Group, 88 Kingsway, London WC2B 6AB.

Conferences and Exhibitions

LONDON

May 8-12 — Olympia
Instruments, Electronics and Automation Exhibition
(Industrial Exhibitions, 9 Argyll St, London W1V 2HA)

May 9-19 — Earls Court
Mechanical Handling Exhibition
(ITF-Iliffe Exhibitions, Commonwealth House, New Oxford St, London WC1A 1PB)

May 17 & 18 — Imperial Hotel
Electronic Designfare
(A.P. Publications, Moreley House, Holborn Viaduct, London E.C.1)

May 21-25 — Bloomsbury Centre Hotel
Radio Show
(International Radio & Electrical Distributors Association, Broadway House, The Broadway, London S.W.19)

BOREHAMWOOD

May 4-12 — GEC-Elliott Plant
World of Automation
(M.D. East, GEC-Elliott Automation, Elstree Way, Borehamwood, WD6 1RX)

OVERSEAS

May 1-3 — Houston
Offshore Technology Conference
(I.E.E.E., 345 E. 47th St, New York, N.Y. 10017)

May 1-3 — Annapolis
Superconductivity
(I.E.E.E., 345 E. 47th St, New York, N.Y. 10017)

May 8-11 — Montreal
Quantum Electronics Conference
(Dorothy Edgar, Courtesy Associates, Suite 700, 1629 K Street, N.W., Washington, D.C. 20006)

May 21-24 — Pittsburgh
Underground Transmission
(I.E.E.E., 345 E. 47th St, New York, N.Y. 10017)

May 22-24 — Chicago
Microwave Symposium
(I.E.E.E., 345 E. 47th St, New York, N.Y. 10017)

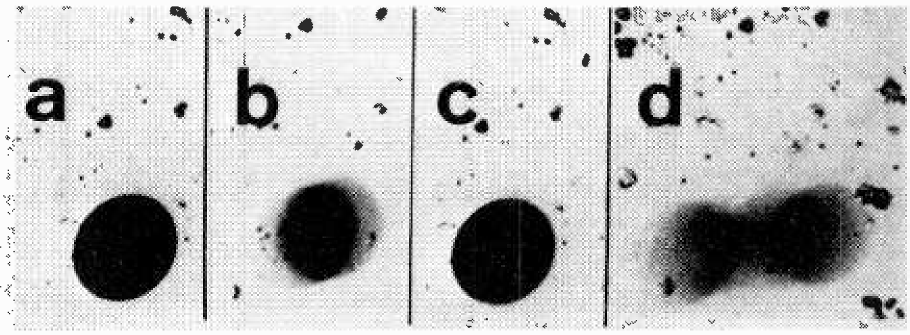
May 30-June 3 — Basle
International Wire Exhibition
(Mack Brooks Exhibitions Ltd, 7 London Rd, St. Albans, Herts.)

May 31–June 11
International Radio, TV and Audio Show
(Federation Nationale des Industrie Electroniques, 16 rue de Presles, Paris 15e)

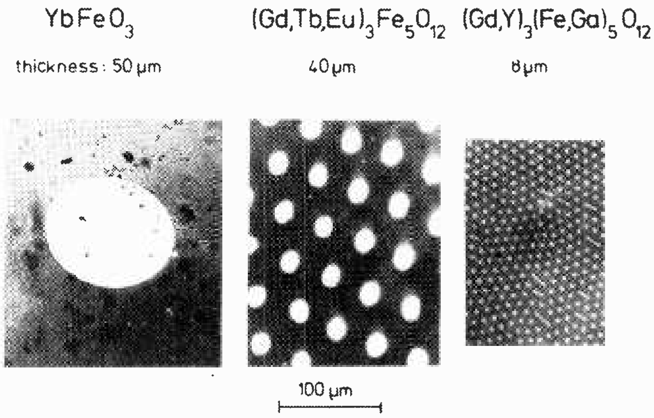
Notes from the Physics Exhibition

Magnetic bubbles for memories?

Interest in magnetic 'bubbles' lies in their possible use in an information store – the presence or absence of a bubble at a certain position representing a bit of information – and hence simple devices that generate, propagate, detect and annihilate bubbles are being studied. Bubbles are cylindrical domains which can be produced in thin sheets of certain ferromagnetic materials (our front cover shows a crystal in production) with a steady, magnetic field applied normal to the sheet. The domains are magnetized in the opposite direction to the bias field. Bubble size is related to material thickness, saturation magnetization, and applied field but with a rare-earth iron garnet 40-μm thick bubble diameter is around 15-μm. When no other fields are present a bubble is held in position because its wall is pinned to certain lattice defects in the crystal. This pinning force can be overcome with a field gradient greater than a certain threshold value resulting in a lateral force on the bubble. In bubble devices this gradient is set up by passing a signal current through wires on top of the crystal sheet. Philips Research Laboratories have found that the threshold can be lowered if an alternating auxiliary field is superimposed on the bias field. This means that either the signal current can be lowered or a faster displacement can be obtained. In the illustration, a sinusoidally varying gradient ('signal') field is applied in (b) giving the periodic displacement shown, but (d) shows the effect of applying an (higher frequency) auxiliary field – the displacement has the same periodicity, but a much greater displacement. (Photos at (a) and (c) are with bias field only and bias plus auxiliary field only.)



Cylindrical magnetic domain (a) can be displaced by application of a gradient field — sinsusoidally varying in this case — to give a lateral force, on the bubble (b). Sensitivity can be increased (d) by addition of an auxiliary field of higher frequency. Photographs are taken by shining polarized light through a thin slice of material and then through an analyser, the plane being rotated according to direction of magnetization (Faraday effect).

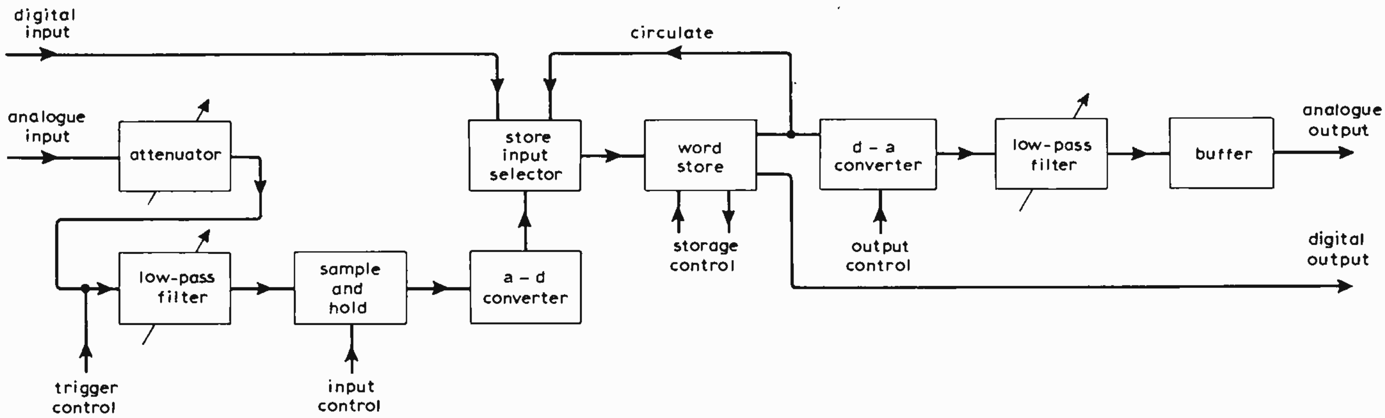


Magnetic bubbles in three different materials.

A simple bubble device to demonstrate detection can be made by overlaying patterns in a soft, easily magnetized material – Permalloy – and in gold for conductors. A bubble approaching the Permalloy changes the direction of magnetization of the Permalloy, which also changes its resistance. Annihilation of bubbles can be achieved by making the current pulse sufficiently high that the total field exceeds a bubble-collapse value. Bubble splitting and transport can be demonstrated with rotating fields and more elaborate Permalloy patterns used as a sort of guide.

Transient event recorders

Instruments designed to store, for subsequent recording and analysis, single events, such as shock waves, spoken words, electrical power line and switching transients and biological responses, were shown by both B & K Laboratories and Data Laboratories. The two instruments work on basically the same principle. An incoming waveform represented as a varying voltage is sampled at a pre-determined rate. The sampled values are converted into 8-bit binary numbers by an analogue-to-digital converter and successive numbers continuously clocked through a store (by shift register technique) and destroyed at the output. Thus the store always contains a series of numbers representing the waveform of the event over a certain interval of time — the interval being determined by the sampling rate and the capacity of the store. When the event ceases the digital information remains static in the store. To read out this stored information the pattern in space (cells of the store) must be converted into a serial pattern in time, and this is done by connecting the output of the store back to the input and continuously circulating the stored



In this transient-event recorder, an analogue waveform is sampled, converted into digital form and successive numbers clocked through a store. After the event, the stored data are circulated and read out in digital or analogue form.

numbers (see block diagram of the B & K instrument). The resulting time sequence of binary numbers at the output of the store is then either fed out directly to a digital recorder (e.g. tape punch, magnetic-tape data recorder) or passed to a digital-to-analogue converter which produces an analogue waveform for display on an oscilloscope or *x-y* plotter. For the analogue recording, the instrument at the same time generates a ramp voltage waveform which provides the timebase needed for the display.

Both instruments have a variety of controls and facilities to increase their flexibility of use. They differ mainly in two aspects. The B & K type 7502 Event Recorder can handle incoming transients with components up to 50kHz (to 25kHz, with internal low-pass filters ensuring that the sampling law is fulfilled) and can provide storage capacities ranging from 2,048 to 10,240 words. Data Laboratories type DL905 Transient Recorder can handle transients with components up to 3MHz and has a storage capacity of 1,024 words.

Liquid crystal displays

Currently available liquid crystal displays generally use the effect known as dynamic scattering. Devices have a capacitor-type construction with transparent electrodes separated by 20μm or so and with a nematic liquid crystal as dielectric. (Nematic, cholesteric and smectic refer to molecular orientation.) The thin film of material is in an ordered molecular state in the absence of a field and is transparent. When the electric field is greater than a certain value the liquid crystal undergoes a turbulence causing it to become diffusing. Forward scattering of incident light makes it difficult to get good contrast in an ambience-lit display and the magnitude of current means a relatively short d.c. life. But it was obvious from the Royal Radar Establishment stand that better things are possible and will almost certainly be commercially available before long.

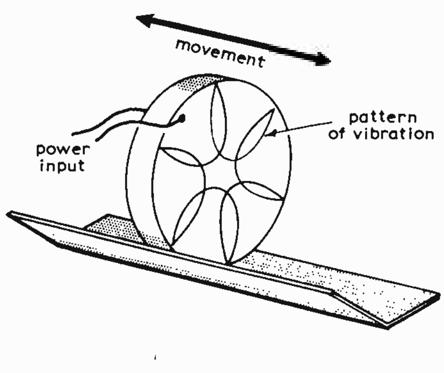
An alternative display device which uses no current and with the consequent promise of long life is the 'twisted' nematic display. The cell is made so that the liquid is twisted by 90° between plates, so turning the plane of polarization of light by 90°. An applied field destroys the twist. Making the two cell plates of polarizing material – planes parallel say – gives transmission with a field and darkness with no field.

Another display relied on 'cholesteric phase change'. Molecules of a cholesteric liquid crystal are left in complete disorder after an electric field has been applied and switched off again, resulting in turbidity lasting for several hours. When a field is applied the molecules line up parallel to the field and the material becomes nematic and transparent. When the field is switched off the display becomes opaque. This effect has shown rise and decay times of the order of tens of microseconds and with its sharp threshold and controlled persistence may be useful in low-voltage displays.

Use of cholesteric liquid crystals with induced conductivity (a nematic/cholesteric mixture) with molecules normal to the field leads to dynamic scattering but lasting for several weeks after the field is removed. Application of a field varying at 1kHZ clears the material. Typically 50V r.m.s. may be needed for 1 second. Further development may possibly reduce this requirement and produce a usable memory.

Ultrasonic linear motor

A piezoelectric transducer can be used as a linear motor by coupling surface wave energy to another surface. Decca Radar suggest that this may be useful – possibly for chart recorders and conveyor belts – and have made some preliminary investigations with barium titanate discs. At one of its resonant frequencies the vibration mode in the plane of the disc resembles the petals of a flower, and the snake-like motion of the periphery provides axial thrust when placed against a flat surface. In the demonstration a disc, 4cm in diameter and 1cm thick and driven with 3W of power at 70kHz, moved along an aluminium channel section at about 10cm/s. Change of direction was effected by turning the disc a few degrees about its axis.



Linear motion is achieved by coupling surface wave energy of a piezoelectric disc to another surface.

Liquid pump without moving parts

A phenomenon in polar liquids is the turbulent nature of liquid motion under the influence of intense electric fields. Charge carriers (unipolar) move under the influence of the electric field and 'pull' the liquid along with them. A prototype nitrobenzene pump has been developed at Grenoble University and was demonstrated by the University of Southampton. In the model, a charge injecting grid consists of a thin stainless steel sheet pierced with 1-mm holes placed opposite a collecting semi-permeable membrane. This membrane acts in such a way as to inhibit injection of charges of the opposite sign from the collector. Earthing the stainless steel electrode and applying a controllable negative potential to the collecting cation-type membrane produces positive charge injection and a flow of the liquid from the steel electrodes placed 3-mm apart and each having a surface area of 0.3cm^2 static heads of 85cm of nitrobenzene and fluid flows for power consumptions of the order of hundreds of milliwatts have been developed.

Laser Doppler velocimeter

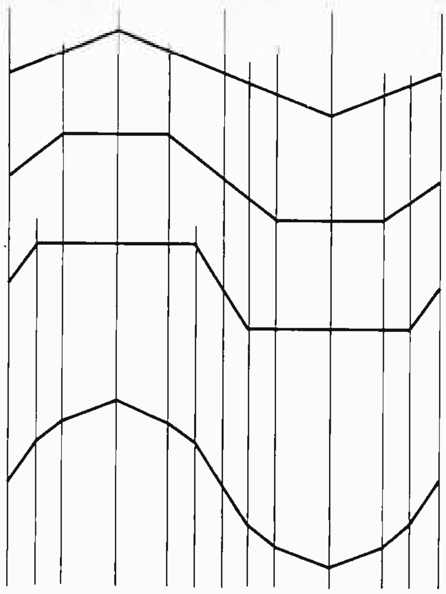
A non-contacting method of measuring the velocities of fluids or solids was demonstrated by the British Steel Corporation, for whom an instrument is being developed by Cambridge Consultants. It depends on the presence in the material of light scattering particles, such as minute air bubbles in liquids. Two beams of light derived from a laser light source of frequency *f*, are focused by a projection lens onto the material. With translucent fluids, the transmitted and forward-scattered light is collected by a receiving lens and focused onto a photo-detector. The scattered light has the frequency $(f + f_D)$, where f_D is the Doppler frequency. In the photo-detector the scattered and unscattered light is mixed to produce an electrical signal of frequency f_D equal to the frequency difference between the scattered and unscattered light.

For analysis of the electrical signal from the photo-detector a special tracking filter has been designed. It incorporates a 20:1 frequency range bandpass pre-amplifier with 18dB/octave roll-off at upper and lower frequency limits, which has automatic gain control to allow signals between 5mV and 1V to be normalized for processing for subsequent circuitry. The tracking filter section follows the pre-amplifier and is designed to function as a normal tracking filter with a bandwidth of 10% of the minimum frequency of the range selected, i.e. 100kHz on a 1MHz to 10MHz range. It has band edge shaping, giving 12dB/octave from centre frequency roll-off to improve selectivity. The filter also incorporates a tracking oscillator, tuned to the signal frequency, which can be used to provide continuous indication of the tuned frequency. During signal 'drop out' periods the frequency is held stable at the value determined during the last signal burst period. A variety of outputs is available from the filter: the direct tracking filter and tracking oscillator outputs, a burst presence gating signal and a squared version of the filter output gated by this signal, which can be used in conjunction with a fated counter to measure the signal frequency. The filter itself incorporates a counter, driven by these signals, which indicates the Doppler frequency on a 4-digit display. It is capable of tracking frequency variations due to turbulence at a 50-kHz rate on the 1 to 10MHz range, and on this range will automatically acquire and track any signal of frequency between 700kHz and 14MHz.

New techniques in function generation

A new technique in sinewave approximation is used in a function generator developed at University College, Swansea in collaboration with the Wayne Kerr Co. The method has an inherent discrimination against second harmonic distortion, is capable of excellent high-frequency performance (transistors are non-saturating) and the circuit shows no increase in distortion over wide variations in ambient temperature.

The new approach is based on approximating a sinewave by adding a triangular wave to a number of clipped or trapezoidal versions of it. The trapezoidal waveforms – each having different slopes and amplitude – are each formed in differential pairs fed from a controlled-current source. Outputs from, say, six trapezoidal shapes are summed by using a common collector load for the differential pairs. Another new technique is for 'exponentially' sweeping the output frequency without using an 'exponential' element. The sweep arrangement exploits



Producing an approximation of a sinewave by adding a triangular wave to clipped versions of the triangular wave.

the improved stability of the integrated monostable multivibrator which is used to derive a pulse train of defined pulse width with a repetition frequency proportional to the output frequency of a square/triangle generator (see diagram). This pulse train is used to operate a current-steering gate which diverts a voltage-controlled current into a storage capacitor. The capacitor is buffered by a high input impedance unity-gain amplifier, and the voltage developed at the output of the amplifier becomes the input to the voltage controlled square/triangle generator, completing the loop and leading to the exponential sweep.

The sweep waveform at the input of the triangle/square wave generator is limited in amplitude by the operation of the hysteresis switch which discharges the capacitor when the upper trip level is reached, so that the cycle repeats itself. Sweep control is embodied in the voltage-controlled current source feeding the current switch.

Piezoelectric actuator

A piezoelectric ceramic has been employed by the Royal Radar Establishment as an actuator for producing small movements. The problem with using the material in this fashion is its non-linear large-signal response, creep and hysteresis effects which can lead to errors as large as ±20% in position. R.R.E. have used a simple control loop to eliminate positional errors in a laser beam deflector.

After suitably mounting the piezoelectric strip, two strain gauges were coupled to it – one in front and one behind. Movement of the strip stretches one strain gauge and relieves pressure on the other. The two strain gauges are connected in a bridge, used to counteract non-linearities in the strip. The output of the strain gauge bridge which is a measure of the position on the strip) is compared with a signal proportional to demanded position and an error signal is derived. After integration the error signal is applied to an amplifier which drives the strip. The whole forms a closed-loop servo system with the strain gauge bridge providing the position feedback signal which often is supplied by a potentiometer in a conventional motorized actuator servo.

The ceramic strip has little inherent damping so a system such as described would tend to oscillate. To overcome this, R.R.E. fitted a small dish to the strip containing silicone oil and a steel ball. The mass of the steel ball prevents it from moving much when the strip is moving rapidly and friction between the ball and the dish effectively damps out the fundamental resonance of the actuator.

Measuring gear wear 'on-line'

A method of measuring gear wear without need to remove the gear wheels has been developed at R.A.E., Farnborough. The initial test set up consisted of a motor driven shaft which drove another 'idler shaft' using a pair of the gear wheels on test. The idler shaft drove an output shaft through another pair of gears. The test method consisted of dismantling the set-up after a period of running and measuring any gear wear that had taken place. This method was unsatisfactory for a variety of reasons so an electronic device was developed which continuously measured gear wear, while the gears were rotating.

The same test jig is employed and gear wear is measured in terms of the phase difference between the output of two magnetic transducers mounted in close proximity to precision toothed wheels fitted to the drive and output shafts. The output of each transducer is fed to a μA709 amplifier operating at high gain in an open-loop configuration. The amplifiers square the signal and feed two single transistor buffer stages which speed up the rise time (100ns) and feed the input of an exclusive-OR gate. This gate provides an output when there is a signal at only one of its inputs but not when there is a signal at both inputs at the same time or when there is no input.

It is arranged that the magnetic pick-offs have outputs with 90° phase difference by correctly positioning the toothed wheels. The output of the exclusive-OR gate is fed to a low-pass filter which drives a mean-level meter amplifier. Any gear wear results in a change in phase difference between the outputs of the transducers which causes the width of the pulse from the exclusive-OR gate to alter. This change is registered on the meter which is calibrated in terms of wear. The equipment has an f.s.d. of 0.5mm with a resolution of 2.5μm.

Trace digitizing equipment

An equipment shown by the Atomic Energy Authority (AWRE, Aldermaston) was designed for the rapid and accurate conversion to digital form of any visible trace or pattern on chart or film. The method of reading was chosen to simulate driving a car along the centre of a magnified image of the trace. The record is supported on a carriage which provides motion in two mutually perpendicular directions which constitute the x and y axes of cartesian coordinates: alternatively one of these motions may be replaced by rotation to provide polar coordinates. A magnified portion of the trace is projected onto a screen. The screen itself carries a central datum point with an indicator in the form of an arrow through that point, and may be rotated about its centre so that the arrow points in any desired direction. The screen is coupled to control circuitry which arranges that the carriage moves at exactly the angle selected by the arrow. A pair of potentiometers are driven by the rotatable screen, one giving an output proportional to the sine of the angle between the arrow and the forward direction, the other an output proportional to the cosine of that angle. These signals are processed to drive stepping motors which move the carriage. Speed of travel is controlled by a foot control.

Carriage position is continuously monitored by two digitizers the outputs of which are displayed on a series of lamps and can be punched on to paper tape. The usual method of operation is to set the punch rate at about three coordinate pairs per second so that the route is continuously digitized as reading proceeds. A constant punch rate has been found satisfactory since, over a complex route, the operator tends to slow down the driving speed so that more coordinates are punched per unit length. This, of course, is exactly what is required.

Matchbox data logger

The clinical research centre of the Medical Research Council exhibited a six-channel data logging tape recorder which measured only 11 × 6 × 2cm. The recorder was originally designed to be carried by a person so that measurements could be taken over a long period of time, changing the cassette every day. Six analogue voltages are sampled at minute intervals, giving a running time of 27h per cassette.

The storage medium is one of the tiny tape cassettes used in miniature dictation machines. The tape transport is controlled by a watch mechanism in the machine. Once per minute a pick-off from the second hand sets a bistable circuit which in turn starts the tape transport. The motor speed is electronically governed. When the tape speed has stabilized pulses from a clock generator are fed to the recording amplifier and to a staircase generator. The output of the staircase generator is compared with the analogue input signal in a differential amplifier. When the two are equal the input clock pulses are inhibited, the motor control bistable is reset stopping the tape motion and the staircase generator is reset so that the circuit is ready to accept the next sample in a minute's time. The number of pulses recorded on the tape is proportional to the value of the analogue input voltage.

On replay both analogue and digital outputs are available. A digital-to-analogue converter provides the analogue output while the digital output is obtained by feeding the recorded pulses into b.c.d. counter chains. The digital output can be fed straight to a paper tape punch or printer for subsequent computer input and analysis. During replay the tape motion can be controlled by the output registering device. With the machine The Clinical Research Centre has come up with a novel, inexpensive and effective solution to their own, and probably many other peoples data collection problems.

Feedforward Error Control

All electronics engineers are familiar with feedback, as a means of reducing amplifier distortion and control system errors, but probably only those who have worked in process control will have heard of *feedforward*. To many engineers in the communications and domestic equipment fields it may be an unknown technique. Nevertheless, feedforward, as a means of error correction, was originated by H. S. Black[1] in 1924, several years before his more famous invention of feedback. (That is, if we allow that Black really did invent feedback: some engineers might hold that it first appeared in the form of James Watt's steam engine governor, or perhaps earlier as a speed controller in windmills.)

The essential principle of feedforward is that a delayed version of the input signal is formed and compared with the output signal of the amplifier (or corresponding control system device), the amount of delay being equal to that of the transit time of the amplifier. From this comparison an error signal is derived, and is injected into the main information stream in such a way as to cancel out the distortion and noise introduced by the amplifier. What this principle does that is not done by the feedback principle is to recognize that amplifiers take a finite time to transfer a signal from the input to the output. The inherent limitation of feedback is that, in its ideal representation, it assumes simultaneity between output and input so that simultaneous comparison can be made between the two and a true error signal be formed. In practice, of course, the output is delayed, and in order to achieve effective simultaneity the feedback system must respond much more quickly than any rate of change existing in the input signal to the amplifier. In other words the bandwidth of the complete feedback loop, including the amplifier, must be several times greater — perhaps an order of magnitude — than the frequency band occupied by the components of the signal. As is well known, this often leads to stability problems.

In the majority of feedback amplifiers, design techniques have found ways round this inherent limitation of the feedback principle, but in some amplifiers, notably r.f. types in communications systems, the requirement for system bandwidths very much greater than the r.f. bands being amplified can be a serious problem. At Bell Labs in the U.S.A., H. Seidel has demonstrated over a period of several years the advantages of using feedforward for r.f. amplifiers in communications channels of about 20MHz bandwidth. Three experiments have been performed. In the first, with a v.h.f. amplifier in 1968, a 108dB dynamic range was obtained[2]. More recently feedforward control was applied to a high performance, baseband, amplifier used in carrier systems[3]. With the addition of feedforward control, all intermodulation distortion products in this amplifier were said to be reduced by about 40dB over its entire frequency range of 0.5 to 20MHz.

In the third experiment, feedforward was applied to a travelling-wave tube, of a type used in an established microwave radio relay system[4]. The laboratory set-up produced more than 40dB reduction of intermodulation over a 20MHz channel at 4GHz, using a single correction stage, and more than 50dB using a second stage.

Circuit complexity in the last-mentioned two demonstrations was roughly double that of the conventional amplifier, but Bell say that integration methods available at both v.h.f. and microwave frequencies suggest that there will be no major difficulty in accommodating much of this increased hardware.

In sum, the feedforward technique, according to Bell Labs, has three main advantages: (1) it does not substantially reduce amplifier gain; (2) gain-bandwidth is 'consumed' entirely within the band of interest; and (3) it is independent of the magnitude or shape of the amplifier delay.

References

1. Black, H.S. U.S. Patent 1,686, 792, issued 9th October 1929.
2. Seidel, H., Beurrier, H.R., and Friedman, A.N. "Error-Controlled High Power Linear Amplifiers at VHF", *Bell Syst. Tech. J.* vol.47, No.5, May-June 1968, pp.651–722.
3. Seidel, H. "A Feedforward Experiment Applied to an L-4 Carrier System Amplifier", *IEEE Trans. on Communication Tech.* vol.Com-19, No.3, June 1971, pp.320–325.
4. Seidel, H. "A Microwave Feed-Forward Experiment", *Bell Syst. Tech. J.* vol.50, No.9, Nov. 1971, pp.2879–2916.

Low-noise Audio Amplifiers

by H. P. Walker, M.Sc

Minimizing transistor noise-figure does not necessarily lead to an optimum low-noise design. In this article the effect of circuit configuration is included with a discussion of optimization with a complex source impedance, and a design procedure is given.

Much has been written on the optimization of transistor noise figure and while with very low and very high source impedances the transistor may be a severe limitation, in many audio amplifiers using modern low-noise transistors the circuit configuration and associated resistors can be the main factor determining signal-to-noise ratio. In this article both the effect of configuration and the limitations imposed by transistor noise are considered in achieving an optimum design.

Optimum configuration

High-quality audio amplifiers usually incorporate overall negative feedback to obtain gain stability and reduce distortion. Assuming that a low output impedance is required, there are two input configurations: series feedback, Fig. 1(a), and shunt feedback (virtual earth), Fig. 1(b), the former being voltage sensing and the latter current sensing.

The majority of audio input transducers are designed as voltage sources feeding an input resistance greater than the source impedance: for example, a magnetic pickup or tape head presents an inductive impedance which is less than the recommended 50-kΩ load up to frequencies of 10 to 15kHz. An exception would be a ceramic pickup driving a load resistance of say 200kΩ in a transistor pre-amplifier[1,2] where conditions approximate to current drive at low frequencies, the turnover occurring at around 1 or 2kHz.

When the source impedance is purely resistive, maximum signal-to-noise ratio is obtained when the input resistance is greater than the source resistance. This is a simple application of the well-known rule[3] 'no resistive attenuation before amplification' and it is clear that if the input resistance equals the source resistance, then Johnson noise* is reduced by 3dB and the signal attenuated by 6dB resulting in a 3-dB loss of signal-to-noise ratio.

Consider now the signal-to-noise ratio obtained for the two cases of Fig. 1. Identical frequency-dependent sources are loaded with an input resistance of R_{in} in both configurations. To examine the effect of configuration alone, assume that the amplifier A is noiseless, has infinite gain and high input impedance, thus creating a virtual earth at node 'E' in Fig. 1(b), and for Fig. 1(a), causing negligible loading at the input and therefore sensing the noise and signal voltage across R_{in}.

The signal-to-noise ratio for the series circuit, derived in Appendix 1, at a frequency f for a bandwidth δf, is

$$\left|\frac{V_o}{V_n}\right| = \frac{V(f)}{\sqrt{4kT\delta f.\left(\dfrac{|Z(jf)|^2}{R_{in}}+R(f)\right)}} \quad (1)$$

assuming R_e is sufficiently small to contribute a negligible thermal noise voltage. Impedance $Z(jf)$ is the complex source impedance $R(f)+jX(f)$, where the resistive part $R(f)$ alone generates Johnson noise $\sqrt{4kT.R(f).\delta f}$.

The signal-to-noise ratio for the shunt circuit, derived in Appendix 1, is obtained by summing signal and noise currents at node 'E'

$$\left|\frac{V_o}{V_n}\right| = \frac{V(f)}{\sqrt{4kT\delta f[R_{in}+R(f)]}} \quad (2)$$

assuming that the feedback resistor R_f is sufficiently high to contribute a negligible noise current $\sqrt{4kT.\delta f/R_f}$. For practical purposes R_f should be made at least three times the impedance of the input arm.*

Comparing these two equations the only difference is in the bracketed terms in the denominator of each expression. The signal-to-noise ratio of the series feedback input will be greater than that of the shunt feedback input when

$$\frac{|Z(jf)|^2}{R_{in}} < R_{in} \quad \text{or when} \quad Z(jf) < R_{in}.$$

That is, at a spot frequency f, the signal-to-noise ratio of the series circuit will be superior to the shunt feedback circuit if the source impedance at that frequency is less than the required input resistance. With a magnetic pickup or tape head, the source impedance is predominantly inductive and less than the input resistance at frequencies

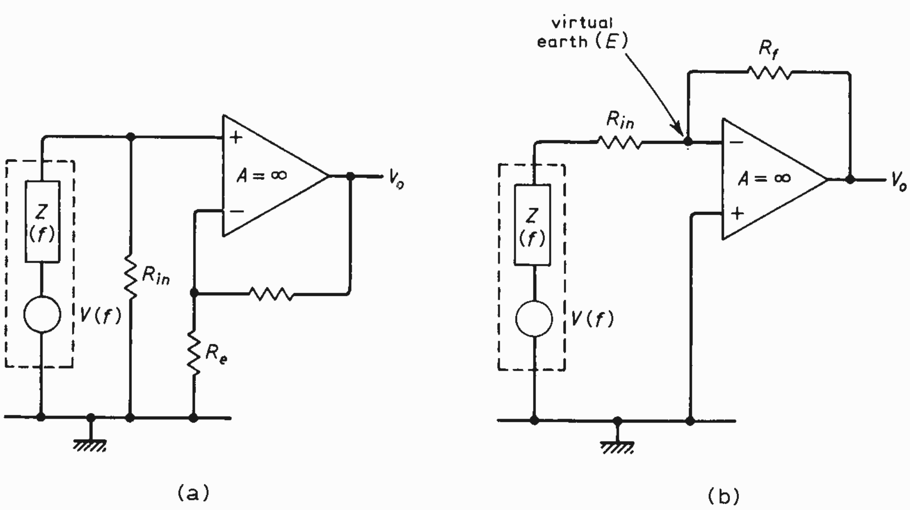


Fig. 1. Signal-to-noise ratio for series (a) and shunt (b) feedback arrangements, as derived in Appendix 1, is shown in graphical form in Fig. 2.

*The thermal noise generated by a resistive impedance in thermal equilibrium with the surroundings, and is $V_n = \sqrt{4kTR\Delta f}$ where $k = 1.38\times10^{-23}$ joules/deg C, T = absolute temperature in K, R = resistance in ohms, Δf is the noise bandwidth and V_n is the r.m.s. noise voltage.

*This same condition applies to biasing resistors connected across the input of amplifier A. Although the connection of a low resistance from the summing junction 'E' to ground will have a negligible effect on the closed-loop gain, particularly with a high-gain amplifier, the signal-to-noise ratio will be seriously reduced because of the high noise current being injected into the virtual earth.

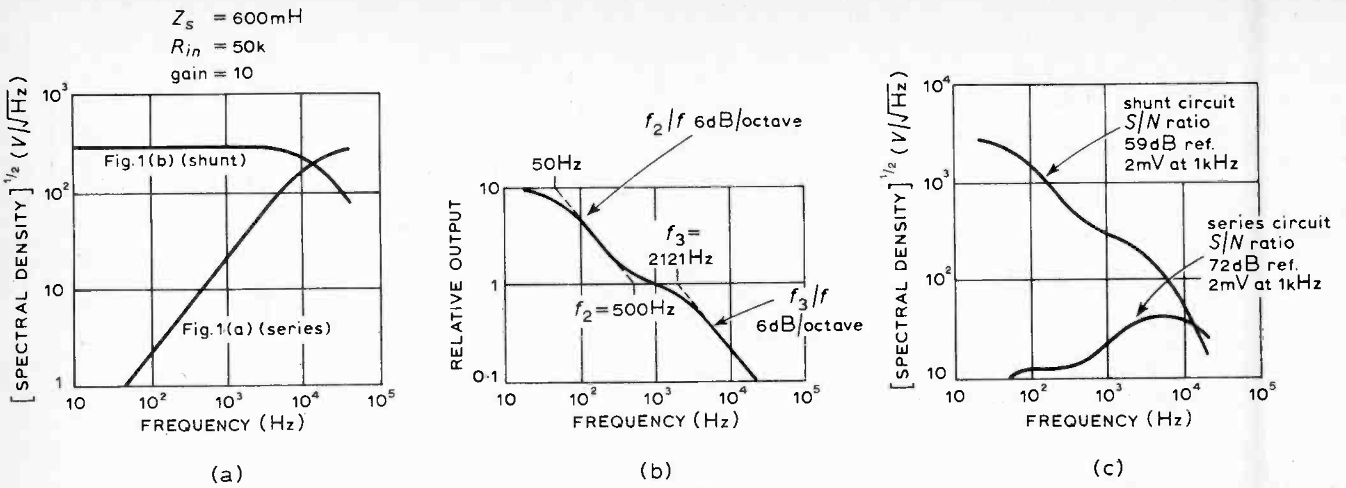


Fig. 2. Signal-to-noise ratio of series circuit is better over most of the band–illustrated in (a) where noise voltage per unit bandwidth falls at 6dB/octave for the series circuit. Curves shown at (c) apply after equalization curve at (b) has been applied. (Signal-to-noise ratio figures are derived in Appendix 2.)

below the turnover in the region 10 to 15kHz. This implies that the series circuit will provide a better signal-to-noise ratio over most of the audio band.

To illustrate this, the r.m.s. noise voltage per unit bandwidth at the outputs of the two circuits is plotted against frequency to obtain the spectral density* functions shown in Fig. 2(a) using typical values of R_{in} as 50kΩ and transducer inductance of 600mH. The graphs indicate that for the series circuit the spectral density falls at 6dB/octave below the turnover frequency of 13kHz due to the inductive source impedance, whereas for the shunt feedback circuit the spectral density is nearly constant up to 13kHz and falls at 6dB/octave beyond this frequency. This latter result is as expected for at low frequencies when $|Z(jf)| < R_{in}$, the noise current in R_{in} entering the virtual earth is nearly constant with frequency; in other words signal-to-noise ratio is independent of source impedance when $|Z(jf)| \ll R_{in}$.

Now pass the outputs of these two amplifiers through an R.I.A.A. equalization network with unity gain at 1kHz and with the frequency response shown in Fig. 2(b). The resulting spectral density functions are shown in Fig. 2(c). Notice that apart from a higher average noise level than that of the series circuit, the shunt feedback configuration generates a large portion of the total noise in the band below 1kHz where the subjective effect is the more disturbing. To calculate the overall signal-to-noise ratio we must find the total noise power by integrating the square of the functions plotted in Fig. 2(c), see Appendix 2. The signal-to-noise ratio, with full R.I.A.A. equalization in the band 100Hz to 20kHz, obtained for the series and shunt feedback circuits respectively are 72dB and 58.5dB referred to 2mV at 1kHz, a typical signal from a tape head or low-output magnetic pickup cartridge. Practical measurements on these two circuits give results in good agreement with the theoretical (Appendix 2).

With a low-impedance resistive source such as the output of a negative feedback amplifier, it is usual to specify a high input resistance so that moderately-sized coupling capacitors can be used and so that several inputs can be fed simultaneously. Equations 1 and 2 indicate that in these circumstances the series input will provide the better signal-to-noise ratio, though usually signal levels are high and noise is not a problem.

As an example the reader may like to compare the performance of the two line amplifiers of my stereo mixer[4] design; the series input of Fig. 16 gives a residual noise level (i.e. with $R_s = 0$) nearly 15dB lower than that of Fig. 15 which is a shunt feedback arrangement. When the source resistance equals the input resistance (e.g. 600-Ω line matching), it follows from the equations that both configurations will give the same signal-to-noise ratio.*

Finally an interesting property of the shunt feedback circuit is that, contrary to common experience, it will generate minimum noise when the input is open circuit as no noise current flows in the input arm and only the feedback and biasing resistors remain. In practice this effect may be masked or even reversed by deterioration of transistor noise figure when operating with such a high source impedance. This is discussed in the next section.

To summarize: the series feedback configuration gives the better signal-to-noise ratio when the source approximates to voltage drive, while the shunt feedback circuit is superior for current drive conditions ($Z_s > R_{in}$). The designer must also ensure that the feedback resistors, R_e for the series circuit and R_f for the shunt circuit, do not introduce an unnecessary source of noise as implied in the derivation of the equations.

Noise in transistors

The equivalent noise generators of Fig. 3 are a universal representation of any noisy amplifier[3,5]. These generators may be thought of as equivalent noise resistances[3,6] R_{nv} and R_{ni}, which in the case of bipolar transistors are a function of h_{FE} and the collector current. There exists, for any value of these parameters, an optimum source resistance[5], $R_{s\,opt} = \sqrt{R_{nv} \cdot R_{ni}}$, which minimizes the noise contributed by the amplifying device.

For silicon bipolar transistors*[3,7,8]

$$V_n = \sqrt{4kT(r_b + 1/2g_m).\delta f}$$

hence $$R_{nv} = (r_b + 1/2g_m) \quad (3)$$

$$I_n = \sqrt{\frac{4kT\delta f}{2h_{FE}/g_m}}$$

hence $$R_{ni} = 2h_{FE}/g_m \quad (4)$$

assuming no correlation between I_n and V_n, and $g_m = qI_e/kT$. In the case of field effect transistors, as with valves, the value of R_{ni} is extremely high and the circuit can often be reduced to an equivalent noise resistance (typically several kilohms at low audio frequencies) in series with the input[6].

It can be shown[6] that noise figure

$$= 10\log_{10}\left(1 + \frac{R_{nv}}{R_s} + \frac{R_s}{R_{ni}}\right)\text{dB} \quad (5)$$

and it follows that the larger the ratio R_{ni}/R_{nv}, the lower the optimum noise figure becomes and the greater the range of R_s over which a good noise figure (<2dB) can be obtained.

However, the expression for R_{nv}, given in equation 3, includes one term which is inversely proportional to collector current

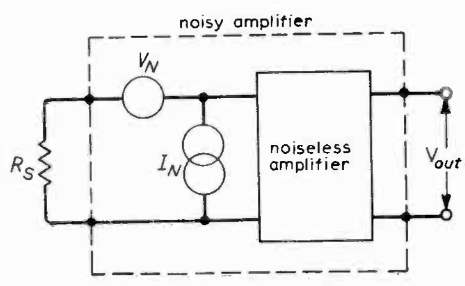


Fig. 3. Equivalent noise generators shown are usually thought of as equivalent noise resistances.

*If the matching input resistance is established by a combination of shunt and series feedback (see footnote on page 236), no loss of signal-to-noise ratio occurs, except that due to the feedback resistors, as the amplifier is both voltage and current sensing. See for example the microphone amplifier described in B.B.C. Monograph No. 46, Feb. 1963, 'Application of Transistors to Sound Broadcasting'.

*Spectral density is usually defined as mean-square noise voltage (i.e. power) per unit bandwidth, but as the plot is of r.m.s. noise voltage per unit bandwidth, the vertical scales in Fig. 2 are in $V_{r.m.s.}/\sqrt{f(\text{Hz})}$.

*We assume that $h_{FE} \gg 1$; also neglect leakage current effects or for practical purposes assume a collector-emitter voltage of less than 5V. The 'effective' base resistance, r_b, is obtained by noise measurement, and is not the h.f. value quoted by manufacturers.

(and therefore under the control of the designer) and a fixed term, r_b, the 'effective' base resistance[9], which is a characteristic of a particular transistor and relatively independent of collector current.

So if we let $R_s = r_b \approx 300$ and make $I_C \geqslant 0.5$mA, the absolute minimum noise figure would be 3dB. Obviously there will be a very severe limitation on achievable noise figure when $R_s < r_b$ and so a matching transformer is normally used with a low-impedance microphone[2] (Z_s = 30 to 50Ω or 200 to 600Ω) to step-up the impedance to the 10 to 30kΩ region where transistor noise figure has an optimum value of less than 1dB. The alternative is to use the technique of parallelling n transistors[6,10] and so reducing the series noise resistance by a factor n; a practical example is the medium-impedance microphone amplifier (Fig. 7) of the stereo mixer[2]. It has also been found[8] that p-n-p transistors have lower effective base resistances than n-p-n types. Figs 4(a) and (b) show the noise performance of a suitable p-n-p transistor, Motorola 2N4126*, which for I_C = 0.5mA† gives a 3-dB noise figure at 1kHz when R_s = 100Ω, (hence R_{nv} < 100Ω), and with R_s = 200Ω a noise figure of 2dB at 1kHz, rising to 3dB at 100Hz.

Both bipolar and field-effect transistors suffer from flicker noise‡, which varies from one device to the next and is very hard to predict. With bipolar transistors it may be represented by an increase in I_n (or a decrease in R_{ni}), Fig. 3, below a certain frequency, and may be characterized by the extra noise generator[8] $|I_f|^2 = K.I_B{}^{\gamma}.f^{-\alpha}.\delta f$, where γ and α are approximately unity and K varies widely with different transistors. Since this generator is proportional to base current, its effect will be reduced by using a low collector current and a device with a high h_{FE} (ref. 12). Fig. 5 demonstrates the relation between collector current and low-frequency noise performance of the Texas TIS97. Flicker noise in j.f.e.ts appears in the voltage generator and can be represented by an increasing value of R_{nv} at low frequencies. Thus for a good noise figure (<4dB) it is preferable to operate bipolar transistors with source resistances less than say 200kΩ, unless selected devices are used, and with low collector currents of say 10μA or less. On the other hand, as the series noise resistance dominates at low frequencies in j.f.e.ts, best noise performance will be achieved with high source resistances of greater than 50kΩ; again this lower limit can be reduced with selected devices.

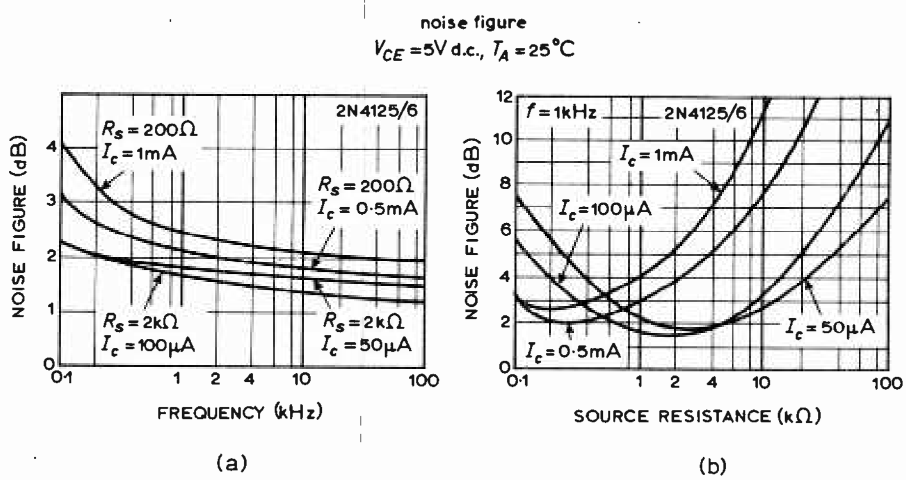


Fig. 4. Noise performance of a p–n–p transistor giving 2-dB noise figure at 1kHz with R_S = 200Ω.

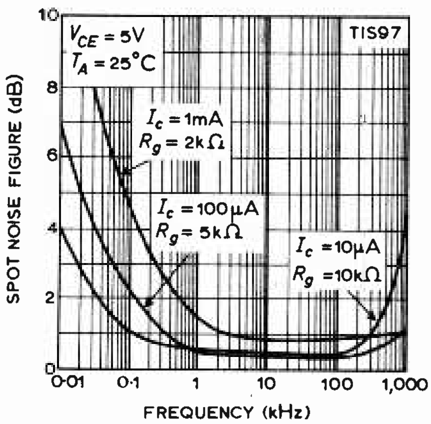


Fig. 5. Effect of flicker noise at low frequencies is reduced by using a low collector current.

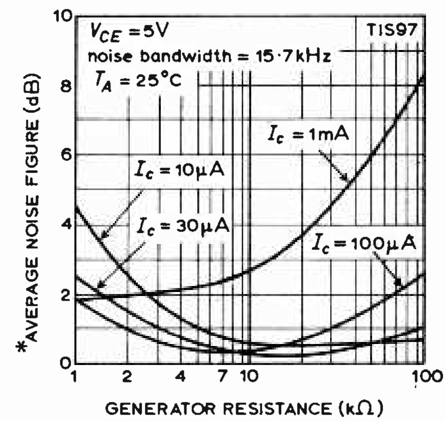


Fig. 6. Broadband noise figure is less than 2dB at I_C of 100μA for a mean source resistance of 5kΩ.

Now two points to consider when optimizing noise figure in a practical circuit. Firstly, equation 5 is true only for a fixed source resistance; practical sources loaded by an input resistance present a complex source impedance, $Z_s(jf) = R_s(f) + jX_s(f)$, of which only the real part $R_s(f)$ is responsible for thermal noise from the source. Analysis of Fig. 3 under these conditions gives the mean-square noise output voltage as

$$\bar{V}_{no}{}^2 = 4kT\delta f\left(R_s(f) + R_{nv} + \frac{|Z_s(jf)|^2}{R_{ni}}\right) \quad (6)$$

at a frequency f for a bandwidth δf, and the spot noise figure is

$$F(f) = 10\log_{10}\left(1 + \frac{R_{nv}}{R_s(f)} + \frac{|Z_s(jf)|^2}{R_s(f).R_{ni}}\right) \quad (7)$$

The broadband noise output voltage can be found by integration of equation 6. Optimization is now much more tedious particularly when there is excess noise in the noise current generator at low frequencies and when there is an equalization network as is usual in tape and disc pre-amplifiers. A brief discussion of this optimization technique is available from the editorial office.

However, for practical purposes a good approximation can be made by finding the range over which the source impedance varies and then choosing a collector current which gives a good noise figure for the same range of source resistances, either from manufacturers' data sheets or by calculation using equation 5. For our example of the magnetic pickup or tape head loaded by a 50kΩ resistor, the source impedance varies from less than 1kΩ at 100Hz to nearly 50kΩ at 20kHz, having a geometric mean of about 5kΩ.

The noise figure is optimized for this source resistance when the collector current is in the region 70 to 100μA. Fig. 6 shows that the Texas TIS97 gives a broadband noise figure of less than 2dB over the required range of source resistances when the collector current is 100μA. The computed signal-to-noise ratio referred to 2mV at 1kHz of an R.I.A.A.-equalized amplifier (similar to Fig. 3 in the stereo mixer[2]) is shown in Fig. 7, plotted against input stage collector current; the maximum is achieved for a current of 35 to 40μA.

The second point concerns the effect of feedback and also transistor configuration. The fact that applying negative feedback affects input impedance may lead to the erroneous assumption that it will also alter the noise figure. Since signal and noise are both reduced by the same factor[6], the open and closed-loop noise figures are the same provided the noise bandwidth remains constant. Thus noise figure must be optimized under open-loop conditions and negative

*A reader, Mr Curl, of San Francisco, has kindly pointed out the excellent noise performance of the Motorola 2N4401 and 2N4403.

†Operating the transistor at a current of several mA causes R_{nv} to increase as the noise current generator, I_n, begins to develop a significant noise voltage across r_b; the generators will then be slightly correlated. It is probably for this reason, particularly when excess noise is present, that higher current transistors do not give an improved equivalent series noise resistance although the ohmic base resistance is less.

‡Flicker noise (1/f noise) is known to arise in bipolar transistors from generation and recombination at defects (i.e. dislocations and impurities) in the emitter-base junction where there is a high concentration of minority carriers[11,12], while burst ('popcorn') noise results from surface states existing in the passivating silicon dioxide layer on the base region. As such, both effects are a function of the processing used in manufacture, and the improved noise performance of modern transistors results from fewer impurities and from low-defect processing. Excess noise in f.e.ts[7] is due to bulk generation/recombination in the vicinity of the channel and is relatively independent of operating conditions.

feedback used to increase or decrease input impedance*. For example, consider the virtual earth configuration of Fig. 1(b). In a low-noise amplifier, the input arm should have a lower impedance than either the feedback arm or biasing resistors, so the source impedance (open loop) seen by the first transistor is very nearly that of the input arm. As a magnetic-pickup amplifier, this impedance will tend towards R_{in} (50kΩ) at low frequencies, so the first transistor should be run at a low collector current of 10μA or less for a satisfactory flicker noise performance, though perhaps a j.f.e.t. would be a more suitable input device in this configuration.

The fact that the input becomes a virtual earth when the loop is closed does not alter the noise figure, as once the source resistance is defined the transistor noise can be represented by a single current generator across the input which injects the same current regardless of whether the loop is open or closed.

Likewise, despite the differing input impedance of the three transistor configurations, all give approximately the same noise figure with a given source resistance and collector current. For example a low source impedance (<200Ω say) may intuitively be thought to give best noise figure with the very low input impedance of a common-base amplifier. This is not so as the base resistance, which limits the noise figure, is still in series with the signal source and the base-emitter junction.

Design procedure

The important points in low-noise audio design can be summarized under the following headings:

Step 1: configuration. By using equations 1 and 2, determine which feedback arrangement will provide the better signal-to-noise ratio with the particular source impedance and input resistance.

Step 2: feedback components. Check that the feedback loop does not cause a deterioration in the signal-to-noise ratio. Keep series resistors in the input circuit low in value and shunt resistors high.

Step 3: first stage noise figure. The source impedance comprises the total impedance (including that of the source itself) seen by the input transistor between its base and emitter[6]. When this is purely resistive, the optimum first stage collector current is found directly from equations 3, 4 and 5. With a complex source impedance, the approximate method described in the previous section may be used, or a more accurate result obtained by using equation 6 as discussed in Appendix 3. The input transistor should be run with a V_{CE} of less than 5V to prevent excess noise due to leakage currents.

Step 4: minimize noise contribution of later stages. The first-stage gain should be high to minimize the noise contribution of the second stage. If the second stage is run at h_{FE} times the current in the first stage (as with a Darlington connection), the shot noise in the base current of the second transistor will equal the shot noise in the collector current of the first. Also the second stage may generate flicker noise, so the ratio of the collector currents should be much less than h_{FE}. E. A. Faulkner[6] recommends that the second stage be run at the same current as the first, though usually other requirements such as gain and harmonic distortion must be considered.

Step 5: biasing. In the shunt circuit, biasing resistors connected to the input should be at least three times the input resistance to prevent avoidable noise current flowing into the virtual earth. In the series circuit the biasing resistors should be greater than or equal to the specified input resistance; never pad out the input impedance with series resistance. In addition to Johnson noise, resistors carrying a direct current give rise to excess noise (in proportion to their voltage drop) so avoid large voltage drops across biasing resistors, although low-noise resistors and the low voltages in transistor circuits make this a secondary consideration.

*In fact this is a useful advantage of feedback, as it allows impedance levels to be modified without altering noise performance; in an open-loop configuration, shunt or series buffering resistors would be required with a consequent deterioration in signal-to-noise ratio.

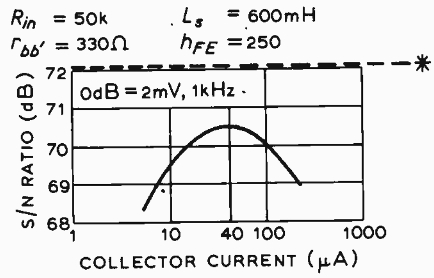


*Fig. 7. Calculated signal-to-noise ratio of an equalized amplifier as plotted against I_C of first stage. *Broken line represents theoretical maximum with noiseless transistor. (Graph neglects excess noise.)*

Example

The amplifier shown in Fig. 8 was primarily intended for a magnetic pickup[2], and as the source impedance is less than the required input resistance over most of the audio band, a series feedback arrangement is chosen (Step 1). The graph of Fig. 7 shows the optimum first-stage collector current to be 40μA (Step 3), whereas the approximate analysis gives 80μA. Note, however, that the difference in signal-to-noise is only 0.2dB. Resistors R_1 in parallel with R_4, the biasing resistor which carries only the very small base current to Tr_1, provides the required input resistance of 50kΩ. Resistor R_7 is kept as low as possible (Step 2) bearing in mind the loading effect of the feedback network on the emitter follower Tr_3. The voltage across Tr_1 is about 2.5V (Step 3), and the high collector load, R_6, allows Tr_2 to be current driven for low distortion. The optimum second-stage collector current for minimum distortion with the particular value of R_6 is about 0.5mA and this gives a satisfactory collector current ratio (Step 4) of 12. The open-loop gain is in the range 5000 to 7000 and harmonic distortion about 1% for 5V$_{r.m.s}$ output.

Appendix 1

Signal-to-noise ratio of series-feedback circuit

The signal voltage appearing across R_{in}, Fig. 1(a), at a frequency f is

$$|V_{in}| = \frac{V(f).R_{in}}{|R_{in}+Z(jf)|}$$

where $Z(jf) = R(f)+jX(f)$.

Short out $V(f)$; then noise voltage across R_{in} at a frequency f for a bandwidth δf is

$$\overline{V_n^2} = 4kT.R_{in}.\delta f.\left|\frac{Z(jf)}{R_{in}+Z(jf)}\right|^2$$

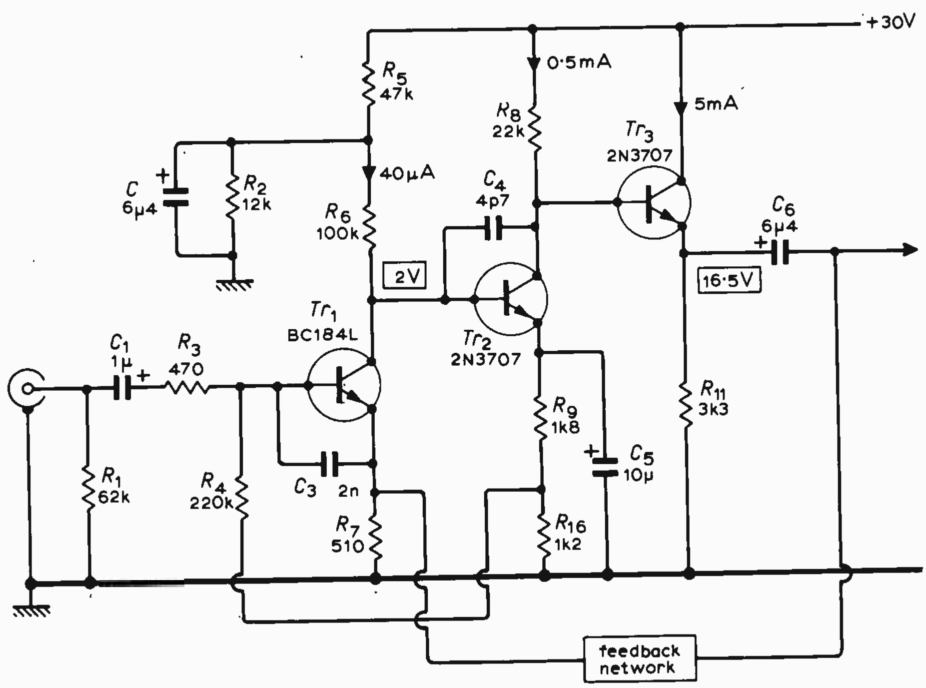


Fig. 8. Circuit of pre-amplifier for a magnetic pickup used to illustrate low-noise design procedure.

due to R_{in}

$$+4kT.R(f).\delta f.\left|\frac{R_{in}}{R_{in}+Z(jf)}\right|^2$$

due to real part of $Z(jf)$

Therefore r.m.s. noise voltage

$$\overline{V_n} = \frac{R_{in}}{|R_{in}+Z(jf)|}.$$

$$\sqrt{4kT.\left(\frac{|Z(jf)|^2}{R_{in}}+R(f)\right).\delta f} \quad \text{(A1)}$$

Hence signal-to-noise ratio is

$$\left|\frac{V_{in}}{\overline{V_n}}\right| = \frac{V(f)}{\sqrt{4kT.\left(\frac{|Z(jf)|^2}{R_{in}}+R(f)\right).\delta f}}$$

Shunt-feedback circuit

The signal voltage at a frequency f appearing at the output, Fig. 1(b), is obtained by summing the currents at node 'E'.

$$\frac{V(f)}{|Z(jf)+R_{in}|}+\frac{V_o}{R_f} = 0$$

$$\text{Therefore } V_o = -\frac{R_f.V(f)}{|R_{in}+Z(jf)|}$$

Short out $V(f)$ and equate the mean square noise currents flowing into node 'E'. Assume R_f is sufficiently large that its noise current can be neglected.

$$\frac{4kT(R_{in}+R(f)).\delta f}{|R_{in}+Z(jf)|^2} = \frac{\overline{V_n}^2}{R_f^2}$$

at a frequency f over a bandwidth δf.

$$\text{Therefore } \overline{V_n} = \frac{R_f}{|R_{in}+Z(jf)|}$$

$$\times\sqrt{4kT(R_{in}+R(f)).\delta f} \quad \text{(A2)}$$

$$\text{Thus s/n ratio is } \frac{V(f)}{\sqrt{4kT(R_{in}+R(f)).\delta f}}$$

Appendix 2

Signal-to-noise ratio of magnetic pickup amplifiers with R.I.A.A. equalization

Assumptions: Pickup cartridge purely inductive; piecewise linear approximation to R.I.A.A. curve, as in Fig. 2(b).

The subject of noise in frequency-dependent networks is treated in several texts[12]. When a noise voltage or current with a certain spectral density is passed through a frequency-dependent network the resulting output has a spectral density equal to the product of the input spectral density function and the square of the magnitude of the transfer function (equation A3).

$$S_o(f) = S_i(f).|H(jf)|^2 \quad \text{(A3)}$$

Series circuit. From equation A1, the noise voltage at a frequency f for a bandwidth f is

$$\overline{V_n}(f)\Big|_{B=\delta f} = \left|\frac{R_{in}}{R_{in}+j2\pi fL}\right|.$$

$$\times\sqrt{4kT.\left(\frac{|j2\pi fL|^2}{R_{in}}+0\right).\delta f}$$

Let $L/R_{in} = 1/2\pi f_t$ where L is the inductance of the pickup.

$$\overline{V_n^2}(f)\Big|_{B=\delta f} = \left|\frac{1}{1+jf/f_t}\right|^2.4kTR_{in}.(f/f_t)^2.\delta f$$

$$= S_i(f).\delta f$$

The R.I.A.A. network can be characterized by three regions

$|A(jf)|^2 = (f_2/f)^2$ for $f_1<f<f_2$,
,, $= 1$ for $f_2<f<f_3$,
,, $= (f_3/f)^2$ for $f_3<f<f_4$.

Output spectral density is

$$S_o(f) = S_i(f).|A(jf)|^2$$

Total mean square voltage over the band f_1 to f_4 is $\int_{f_1}^{f_4} S_o(f).df$. Thus

$$\overline{V_{no}}^2 = 4kTR_{in}.\int_{f_1}^{f_4}\frac{(f/f_t)^2}{1+(f/f_t)^2}.|A(jf)|^2.df$$

This integral is evaluated in three parts corresponding to the three regions of the R.I.A.A. characteristic. With L = 600mH. R_{in} = 50kΩ, f_1 = 50Hz, f_2 = 500Hz, f_3 = 2120Hz and f_4 = 20kHz,

$$\overline{V_{no}}^2 = 4kTR_{in}.(0.65+18.8+290)$$

Thus the r.m.s. noise voltage $\overline{V_n} = 0.5\mu V$, which referred to 2mV at 1kHz gives a signal-to-noise ratio of 72dB.

Shunt circuit. Starting from equation A2 and assuming $R_f = R_{in}$, but neglecting the noise current due to the feedback resistor, the mean square noise voltage is

$$\overline{V_{no}}^2 = 4kTR_{in}.(4000+1600+1600)$$

Hence the r.m.s. noise voltage is 2.5μV, which gives a signal-to-noise ratio of 58.5dB referred to 2mV at 1kHz.

To check these results, noise measurements were made on the series feedback circuit shown in Fig. 8 and on the shunt feedback circuit described by Linsley-Hood (July 1969 issue). The inputs of the amplifiers were loaded with a 600-mH inductor to simulate the cartridge. The signal-to-noise ratios for the series and shunt circuits referred to 2mV input at 1kHz were 70dB and 58dB respectively. These results show good agreement with the theory. Measurement bandwidth was less than that used in the calculation and transistor noise is also making a contribution.

References

1. Burrows, B. J. C., Ceramic Pickup Equalization, *Wireless World*, July and August 1971, pp.321–4 & 379–82.
2. Walker, H. P., Stereo Mixer—1, *Wireless World*, May 1971, pp.221–5.
3. Baxandall, P. J., Noise in Transistor Circuits, *Wireless World*, vol. 74 1968, pp.388–92 & 454–9.
4. Walker, H. P. Stereo Mixer—2, *Wireless World*, June 1971, pp.295–300.
5. Haus, H. A. *et al*, Representation of noise in Linear Twoports, *Proc.I.R.E.*, vol. 48 1960, pp.69–74.
6. Faulkner, E. A., Design of Low-noise Audio-frequency Amplifiers, *Radio and Electronic Engineer*, vol. 36, 1968, pp.17–30.
7. Mantena, N. R., Sources of Noise in Transistors, *Hewlett-Packard Journal*, vol. 21, 1969, No. 2. Swiontek, M. C. and Hassun, R., Influence of Transistor Parameters on Transistor Noise Performance—a simplified Presentation, *Hewlett-Packard Journal*, vol. 16, 1965. No. 7.
8. Chenette, E. R., Low-noise Transistor Amplifiers, *Solid-State Design*, February 1964, pp.27–30.
9. Faulkner, E. A. and Harding, D. W., Some Measurements on Low-noise Transistors for Audio-frequency Applications, *Radio and Electronic Engineer*, vol. 36, 1968, pp.31–3.
10. Faulkner, E. A., Optimum Design of Low-noise Amplifiers, *Electronics Letters* vol. 2, 1966, pp.426–7.
11. Jaeger, R. C. and Brodersen, A. J., Low-frequency Noise Sources in Bipolar Junction Transistors, *I.E.E.E.Trans.* vol. ED-17, 1970. pp.128–36.
Plumb, J. L. and Chenette, E. R., Flicker Noise in Transistors, *I.E.E.E.Trans.* vol. ED-10, 1963, pp. 304–8.
Conti, M., Surface and Bulk Effects in Low-Frequency Noise in N-P-N Planar Transistors, *Solid-State Electronics* vol. 13, 1970, pp. 1461–9.
12. Hsu, S. T., Noise in High-gain Transistors and its Application to the Measurement of Certain Transistor Parameters, *I.E.E.E. Trans.* vol. ED-18, 1971, no. 7.

Meetings

LONDON

1st. IEE — Colloquium on "Digital frequency synthesis in communication systems" at 10.00 at Savoy Pl., W.C.2.

2nd. IERE/IEE — Colloquium on "Hybrid computing systems" at 14.30 at 9 Bedford Sq., W.C.1.

3rd. SERT — "The registration of technician engineers and technicians" by A. J. Kenward at 18.30 at Mullard House, Torrington Pl., W.C.1.

5th. IEE/I. Meas. Control — Colloquium on "Industrial applications of queueing theory" at 10.00 at Savoy Pl., W.C.2.

8th. IEE — Discussion on "Experiences with amorphous semiconductor devices" at 17.30 at Savoy Pl., W.C.2.

9th. AES — "Transformers and the audio engineer" by P. J. Baxandall at 19.15 at the Mechanical Engineering Dept., Imperial College, Exhibition Rd., S.W.7.

10th. I.Phys./IEE — Colloquium on "Semiconductor memories" at 10.00 at Savoy Pl., W.C.2.

10th. IEE — "The applications of electroluminescence and acoustic surface waves" by Dr. E. V. D. Glazier at 17.30 at Savoy Pl., W.C.2.

10th. SERT — "Digital processing of radio and television signals" by Dr. B. Moffatt at 19.00 at the I.T.A., 70 Brompton Rd, S.W.3.

11th. IERE/IEE — Colloquium on "Implantable cardiac pacemakers" at 14.30 at 9 Bedford Sq., W.C.1.

11th. IEE — "The future role of the technical journal" by M. G. Lowe at 17.30 at Savoy Pl., W.C.2.

12th. IEE — "Integrated navigation systems for aircraft and hovercraft" by D. C. Price at 17.30 at Savoy Pl., W.C.2.

18th. RTS — "Graphics in B.B.C. television" by Colin Cheesman at 19.00 at the ITA Conference Suite, 70 Brompton Rd., S.W.3.

GLASGOW

31st. SERT — "The registration of technician engineers and technicians" by A. J. Kenward at 19.30 at McClellan Galleries, Sauchiehall St.

IPSWICH

10th. SERT — "Video tape recorders" by C. H. W. Allvey at 19.30 at Ipswich Civic College, Rope Walk.

Electronic Building Bricks

23. The gate

by James Franklin

The action of an electronic gate is analogous to that of a farm gate or level-crossing barrier — it allows things through during any period it is opened under control. In the electronic case the 'things' allowed through are successive values of an electrical variable representing information (Fig. 1). In practice the information might be samples of a signal (see Part 16, Fig. 2), a sequence of pulses representing numbers, or a sequence of cycles of an oscillation.

Looking at the electrical action of the gate more closely, we can consider it basically as an on/off switch, connected into the information path, which is usually a straightforward conductor in a circuit. The switch, however, is not operated by hand or other mechanical means because it is not actually mechanical, but by electrical control signals.

The basic idea of a gate as a switch is valid but is verbally confusing because, as can be visualized, the gate is opened to allow information to pass through when the switch is closed (circuit completed), and closed to information when the switch is opened (circuit broken). It is better to think directly in terms of the actual electronic components that are used to construct gates in practice. These are

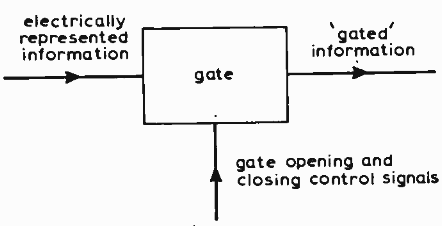


Fig. 1. Basic function of the gate.

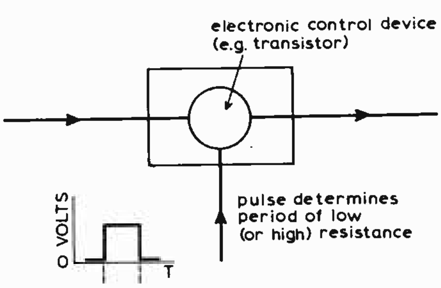


Fig. 2. Practical gate is an electronic device of which the resistance can be electrically controlled.

mainly the transistor, the thermionic valve and the rectifier. As we have seen from Parts 9 and 19 these devices provide a conduction path for electrons, the resistance of which can be varied by means of an electrical control signal (e.g. a varying voltage). In fact the resistance can be varied from a very low value, almost equivalent to the on/off switch being closed, to a very high value, almost equivalent to the switch being open. Moreover this transition can be achieved very rapidly by the control signal, in a time usually much shorter than that taken by a mechanical switch. Without specifying any particular component we can therefore think of the gate as an electronic control device acting as an electronic switch.

The electronic switch thus visualized is turned 'on' (low resistance) or 'off' (high resistance) by applying one or the other of two extreme values of control signal. Typically these would be two voltages, one being zero voltage and the other a constant positive or negative voltage. In practice this would mean the switch would be turned 'on', if normally 'off' (or turned 'off', if normally 'on') by a voltage pulse, as shown in Fig. 2, and would remain in that state for the duration of the pulse. The two voltages forming the pulse would constitute the gate control signals indicated in Fig. 1.

Another example of the process of 'gating', besides that in Fig. 2 of Part 16, is shown in Fig. 3. Here the signal applied to the input of the gate is a sequence of short pulses (a), the gate control signals are the two voltages of the square pulse (b) and the signal emerging from the gate is the short sequence of pulses (c), the length of which is determined by the duration of the pulse (b). One application of this principle is in measuring the frequency of events or oscillations. If, as shown, the gate control pulse has a duration of 1 millisecond, and it allows four pulses (or cycles of any oscillation) through in the 'open' time, the frequency of the input signal pulses (or cycles of oscillation) is 4 per millisecond, or 4000 per second (4000Hz). To make this measurement, of course, the pulses emerging from the gate while it is opened must be counted by an electronic counter. Yet another application is in colour television receivers — to select a 'burst' of ten cycles of an oscillation, used for colour synchronizing, from the rest of the television picture waveform.

So far we have mentioned only applications in which the duration of the gate control pulse is much shorter (Part 16) or much longer than that of the significant variations in the input. If the 'open' period of the gate is comparable in duration with a particular part of the input information (say a single pulse), and occurs at the same time, then we can consider the gate in a different light — as a coincidence detector. This is illustrated in Fig. 4. At (a) the gate 'open' period occurs at a different time from the information pulse, so the gate is closed to the information and there is no output

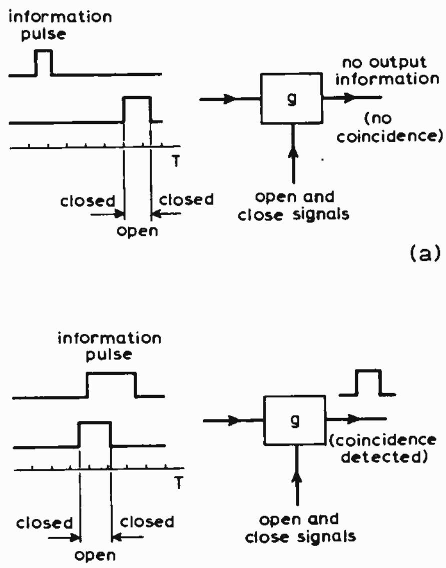


Fig. 3. Example of 'gating' a number of pulses (c) from a train of short pulses (a).

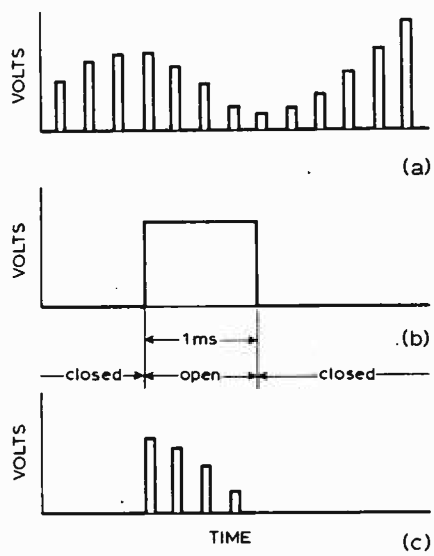


Fig. 4. Gate acting as a coincidence detector: (a) non-coincidence of pulses; (b) coincidence of pulses.

from the gate. This absence of output information indicates lack of coincidence. At (b) however, the gate is open for at least a part of the period of the information pulse, and the presence of information at the output of the gate indicates that coincidence has occurred.

Circuit Ideas

Voltage-controlled triangle/square generator

The triangle/square generator described by G. B. Clayton (*W.W.* December 1969) consisting of an integrator and comparator may be modified to give voltage control of frequency by replacing the diode bridge with f.e.t. switches.

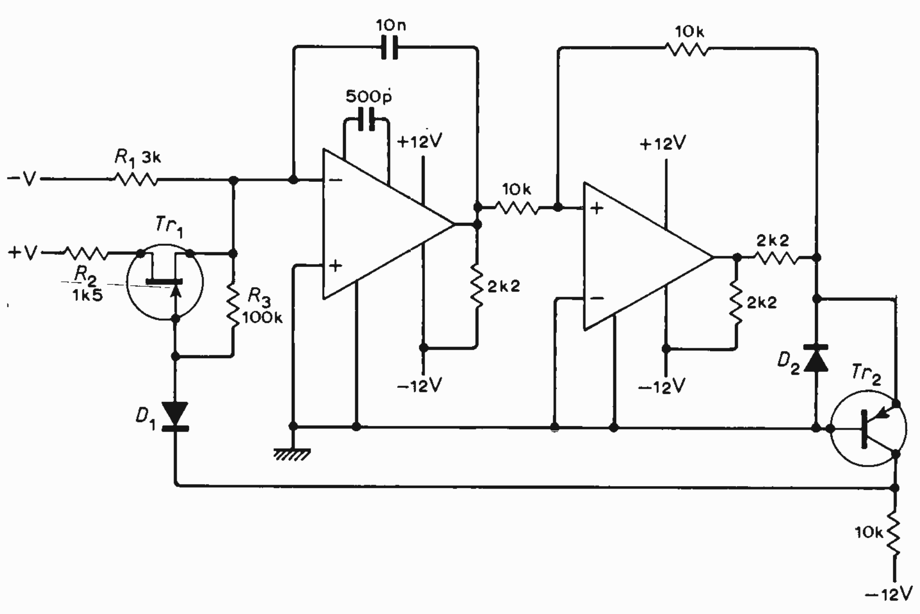
A practical circuit for producing symmetrical waveforms only, may be made using one f.e.t. switch. The mark-space ratio is given by:

$$\frac{1}{(R_1/R_2)-1}$$

so for $R_1 = 2R_2$ the ratio is unity, and R_2 may include a preset to fix the ratio exactly. SL701C amplifiers are used to maintain good waveform at the higher frequencies (100kHz) and a diode and transistor used to clip the square wave and produce the f.e.t. control voltage. Compensation is not required for the comparator but a small capacitor (10–50pF) across D_1 or R_3 may be needed to balance-out any stray capacitance.

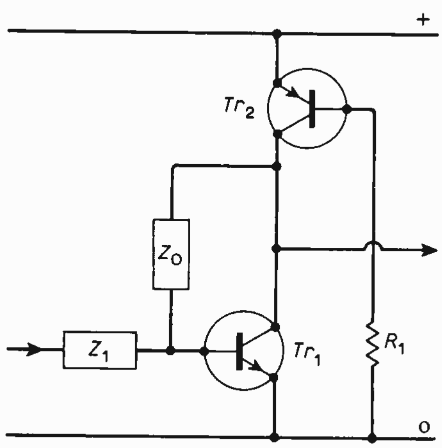
Single-voltage frequency control may easily be provided by the use of a unity gain inverting amplifier, and the frequency may then be varied using one potentiometer or a sweep unit. Using one of the standard techniques for triangle-sine-wave conversion, the unit forms the basis of an audio sweep generator.

The circuit as it stands suffers from poor frequency stability against temperature due to the temperature coefficients of D_2 and Tr_2, but this may be improved by increasing the amplitude of the square-wave using 3V zener diodes.
R. J. Tidey,
Oxford.



Simple d.c. amplifier

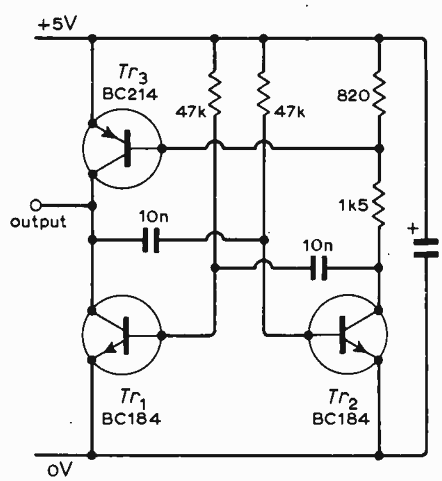
In non-reversing control systems an amplifier giving only one polarity of output is acceptable, and in many cases the amplifier is not required to have its input and output at the same d.c. level. For such circumstances the circuit shown gives a high voltage gain and low drift. Current gain can be added by means of emitter followers.



The input and feedback are mixed at the base of Tr_1. The collector load for Tr_1 is the collector circuit of Tr_2 which provides high voltage gain. The common collector current is controlled by R_1 which gives an almost constant base current to Tr_2, so that the collector current increases with temperature and compensates for changes in the characteristics of Tr_1. Control accuracies of 0.1% can be achieved with this type of amplifier.
R. M. Carter,
Lincoln.

Fast rise-time multivibrator

One of the disadvantages of the ordinary astable multivibrator is that the output waveform has a rise-time determined by the time constant formed by the collector load resistor and the coupling capacitor to the opposite base. The method of improving rise-time by using a diode to isolate the output from the coupling capacitor has the disadvantage that the rise-time is degraded by capacitive loads.



In the circuit shown, a fast rise-time is obtained by replacing the collector load resistor of Tr_1 by a p-n-p transistor Tr_3 which is switched on and off by the collector current of Tr_2. The danger of excessive dissipation in Tr_1 or Tr_3 if oscillation ceases can be reduced by inserting a resistor in series with Tr_1. With the component values shown in the circuit diagram, rise and fall times of 0.5μs were achieved and even with a load of 0.05μF the rise and fall times were less than 2.5μs.
C. R. Masson,
Edinburgh.

An invitation

If you have developed or happened upon an original circuit configuration to perform a simple or a complex operation, or have used standard components in an unconventional manner, send a concise description, in the form of a circuit diagram and notes, and we will consider its publication as a circuit idea. £5 is paid for each contribution published.

Doppler Effect

Fundamentals of the phenomenon as it occurs in sound waves

by 'Cathode Ray'

Just as the foreigner who wants to know the English words for a container, for the activity of a pugilist with his fists or of a mariner with his compass, for the most expensive seats at the opera or the free seats for the jury in the courtroom, for a shrub used for edging, for a Christmas gift for the postman, for a yellow check pattern at busy urban crossroads, and for a cottage used by huntsmen, is astonished to learn that they are all 'box', so some of the readers of *Wireless World* and similar (though of course inferior) literature may be somewhat bewildered by the appearance of the Doppler effect in a variety of quite different contexts. It crops up in radar, in loudspeaker distortion, in radio-wave propagation studies and in astronomy — both radio and optical varieties. Almost invariably, if any explanation at all for the term is vouchsafed, one is told that it is what causes the drop in pitch of the sound emitted by a rapidly moving vehicle as it goes past. Even if the writer goes so far as to offer some explanation of why this happens it is usually so brief (in order not to divert too much attention from his main theme) as to leave plenty of room for misunderstanding.

For instance, one can easily get the idea that the pitch-drop is due to the change in the speed with which the sound waves reach the listener; faster when the source of sound is approaching, and slower when it is going away. After all, that is what happens with bullets fired from a moving vehicle. And it is usually taken for granted that the Doppler effect in radio and light waves is the same in principle as with sound, and that exactly the same formulae apply, when allowance is made for the difference in wave speeds.

Even Doppler himself seems to have lacked a completely correct grasp of the effect named after him. He, by the way, was an Austrian physicist who lived from 1805 to 1853 and first pronounced upon his effect in 1842. Basically he had the right idea, but he went much too far in crediting it with the differing colours of the stars. (Temperature is mainly responsible for that, though of course Doppler effect is of very great importance to astronomers, but it appears as a shift in the positions of the spectral lines on the frequency scale.)

Confirmation of the Doppler theory by experiment was carried out (not by Doppler) in 1845. We can pleasurably visualize the modus operandi as described by Alexander Wood in his book *'Acoustics'*: 'The first experiments were carried out by Buijs Ballot on a single-track railway between Utrecht and Maarsen. A trumpet was carried on the locomotive and three others were used by groups posted at the side of the track. The trumpets were sounded alternately on the locomotive and at the side of the track, and the apparent change of pitch was observed both for a moving source and for a moving observer by musicians whose estimate of small intervals of pitch was considered to be reliable.'

To keep things simple — or at least not so complicated as they could be — let us assume two things. One is that all the speeds to be considered are uniform, or constant. The other is that all movements are directly towards or away from the observer — or listener, if you prefer. Short of actual suicide, anyway. That is to ensure that the observed rise or fall in pitch will itself be constant. If the track of the source of sound passes wide of us the change in pitch is gradual. During 1944-5, when we were cast unwillingly in the role of observer in experiments carried out by the Nazis, we became quite clever at judging whether there was any need to dive for shelter when a V1 (flying bomb or doodlebug) was heard approaching. An early and gradual drop in pitch of the guttural note, and one could carry on unconcernedly — 'I'm all right, Jack'. A constant and rapidly swelling note, and one got underground at the double, or even treble.

First, let us suppose that we, the observers, are stationary, and so is the air — there is no wind. A source of sound of constant frequency f hertz (Hz) is also stationary, some distance away. It might be easiest to suppose it is a motor bike with the engine running in neutral, so that the sound is in the form of pulses and we can draw the peaks of the air-wave pulses as circles surrounding the source (Fig. 1). (This is a variant of the stone dropped in the pond, beloved by writers of elementary books on the principles of radio.) The sound pulses are numbered in the order in which they are emitted; No. 1 is just assaulting our ears at the instant depicted, when No. 9 is on the point of emerging from that part of the machine they call the silencer. The radial distance between successive peaks is the wavelength of the sound, denoted by λ as in radio. If the speed or velocity of sound in metres per second (m/s) is called V, then in one second f waves or pulses have spread out over V metres. So the length of each of them (λ) is V divided by f:

$$\lambda = \frac{V}{f}$$

But you can forget about that for a little while, because the pitch of a sound depends on its frequency, not its wavelength. If you have any doubt about that, don your skin-diving kit and listen to a constant-frequency sound under water,

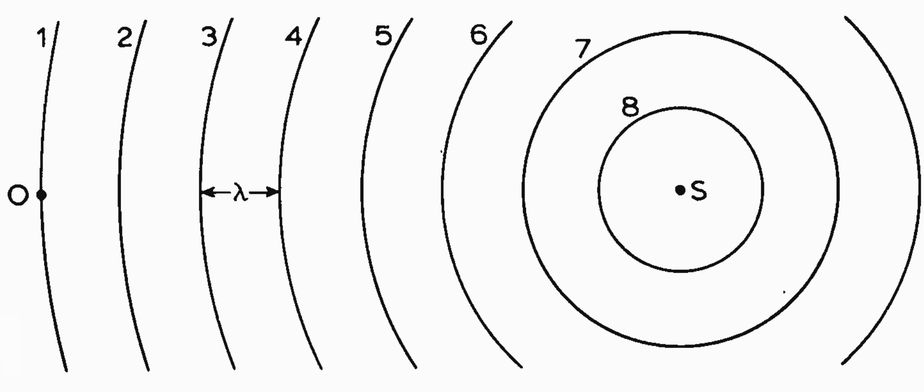


Fig. 1 The first few wave peaks emitted from a stationary source of sound, S. The first of them has just reached a stationary observer, O.

in which sound travels more than four times as fast as in air so the waves are more than four times longer. Yet the pitch is the same as in air.

In the conditions assumed, the sound waves reach our ears with the same frequency as they were emitted from the source. Obviously. Mere distance has no effect on the pitch of the sound we hear, though it mercifully reduces its intensity. The statement just made, though plausible, is however not quite accurate. And I don't mean the 'mercifully' bit, which betrays that my motor-cycling days are over, and might be contested by those for whom they are not. The fact is that pitch is a subjective thing (that is to say, is produced by our own senses, nerves and brain) and depends slightly on loudness as well as on frequency. But, like the trumpeters of Utrecht, let us ignore that complication and assume that pitch is a measure of the frequency of the sound reaching the ear.

Next, keeping our fixed distance from the revving engine, let us suppose that a steady wind springs up. If it blows direct from the source to us, then its speed (call it v_m, for velocity of the medium, the air) is added to V, so that the sound waves come to us at a speed of $V + v_m$ m/s. The time they take, though shorter than with no wind, is the same for all the waves, so the frequency with which they reach us is the same as that with which they were generated, whatever the direction and speed of the wind, so long as it is steady.

Now let us suppose that the motor bike is coming straight for us at a speed of v_s m/s, and that its rider is cleverly managing to keep its engine revs the same as when it was standing still. The important thing to keep in mind is that once a sound has been committed to the air its progress is in no way affected by what its source then does, whether it stays still or moves. So nothing about wave No. 1, emitted when the machine was instantaneously in the same position as in Fig. 1, has changed. But by the time No. 2 starts, $1/f$ seconds later, the source has moved nearer. The distance nearer is equal to speed × time, v_s/f metres, represented in Fig. 2 by the distance between S positions 1 and 2. So wave-front No. 2 is drawn from that new centre, 2. Similarly for the other six shown. Because the waves are closer together along the line of motion towards us at O, they reach us at shorter intervals of time. In other words, to us their frequency is higher than f.

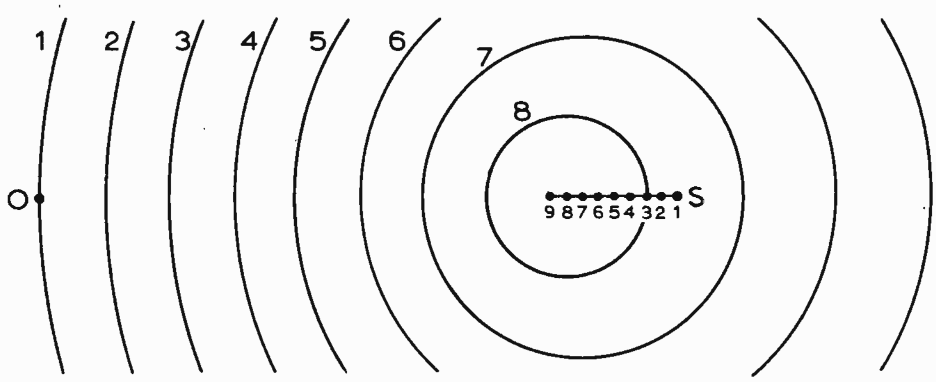


Fig. 2. In this situation the observer is still stationary, but the source is moving towards O at constant speed v_s.

With the stationary source, the wavelength, as we have already noted, was V/f metres. Now we have found that in our direction it is shorter by v_s/f metres. So as far as we are concerned the wavelength is not V/f but $\frac{V}{f} - \frac{v_s}{f}$

The corresponding frequency, which we will call f', is V divided by the wavelength:

$$f' = V \div \frac{V - v_s}{f} = \frac{fV}{V - v_s}$$

If the motor bike happens to be rather unusually fast, so that $v_s = V$ (a condition known as Mach 1 among aviators, who do this sort of thing and think little of it) the wave starting-point scale stretches all the way from S to O, and what we observe is nothing at all until the arrival of wave No. 1. Its arrival is likely to be somewhat blurred by the arrival of all the other waves simultaneously *and* the source itself. That could well bring our observations to an abrupt end; but if the rider manages, in spite of his hurry, to achieve a near miss, we get nothing worse than a sonic boom. Looking back at the formula we see that making $v_s = V$ drives f' to infinity, which is another way of putting it.

If a source of sound *exceeds* Mach 1 ($v_s > V$), then it arrives before any of its sound, which then follows in reversed order in time, corresponding to negative values of f'. After a few months to enable us to study the better-known aspects of the Doppler effect by means of the V1, the Nazis extended our course of study into its more bizarre manifestations by organizing the supersonic V2. The fact that it arrived before any of its associated sound precluded one from taking evasive action, but enabled those not on target, if they were interested, to note the unusual acoustical accompaniments.

When the sound source has gone past, the sign of v_s is of course reversed, so the apparent frequency (if we are still alive to hear it) is

$$f' = \frac{fV}{V + v_s}$$

and this is clearly lower than f.

To take some reasonable figures, suppose f is 50 Hz, corresponding to 3000 r.p.m. with a single-cylinder two-stroke. At 15°C and normal atmospheric pressure, V is 342 m/s. If the speed is 45 m.p.h., v_s is 20.1 m/s. So when v_s is towards us, $f' = (50 \times 342)/(342 - 20.1) = 53.15$ Hz. This drops to 47.25 Hz when v_s is away from us. The sound we hear is thus about 6% sharp or flat, amounting to about a semitone, according to whether its source is coming or going. So the total drop in pitch is about a whole tone.

If now we reverse the procedure, moving ourselves at 45 m.p.h. past a standing noise maker, we might expect exactly the same thing to happen. The relative speed between the source of sound and ourselves is the same. In the diagram, Fig. 3, the wave fronts are of course the same as in Fig. 1. The corresponding positions of the observer — us — are shown on the left. After a period of one wave, during which time wave No. 2 moves to the shown position of No. 1, we are no longer there but have already been passed by it on our way to observer position 2. So the *observed* time interval between waves 1 and 2 — and between every succeeding pair of waves — is less than the interval between them at the source. So the observed frequency is higher, as expected. And if we tried it we would again notice a drop of about a whole tone as we passed the source of sound. But to make sure let us again go through the calculation in detail.

Because we are moving towards the source of the sound, relative to us the speed of the sound is $V + v_o$, where v_o means the speed of the observer. Frequency being velocity divided by wavelength, the frequency to us is not just V/λ but $(V + v_o)/\lambda$. And λ, as we have noted, is V/f.

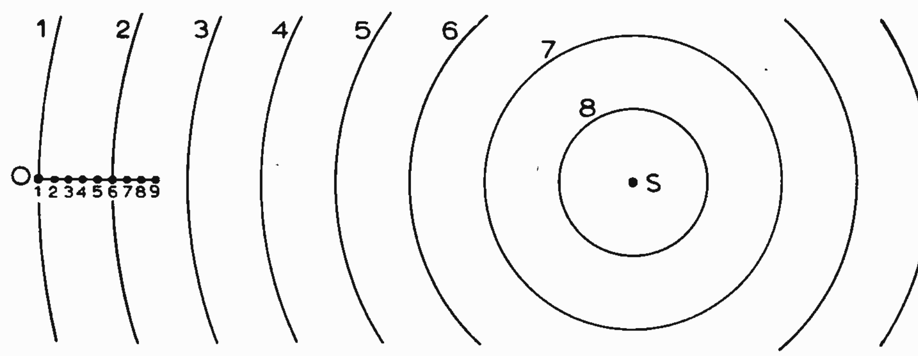


Fig. 3. Here it is the source that is stationary and the observer moving towards it at constant speed v_o.

So the observed frequency, f', is now

$$\frac{f(V+v_o)}{V} \text{ or } f(1+\frac{v_o}{V})$$

If we compare this with the result for the moving source we can see it is not the same. Using the same values as in our example, with v_o also 45 m.p.h. we find the observed frequencies are 52.95 and 47.05. These are equally above and below 50, whereas with the moving source the high frequency was 3.15 Hz above 50 and the low one 2.75 below. The total drop in Hz seems to be 5.90 in both cases however, and so they are to sliderule accuracy. But if we took the trouble to be more precise we would find a difference of about 1 in 3000 — hardly enough to be noticeable even to the reliable musicians of Utrecht (or anywhere else). And if we reckon the total drop in frequency as the ratio of the low to the high, it is *exactly* the same, being $(V - v_s)/(V + v_s)$ in one case and $(V - v_o)/(V + v_o)$ in the other. Musical intervals (of pitch) are such ratios, so even an ideal musician would always hear exactly the same interval for a given relative speed between him and the source of sound. But he would have to know which was moving in order to determine the absolute pitch of frequency of the sound at the source.

The inequality in the sharpening and flattening of the pitch of sound when it is the source that moves, quite small in our example, is much greater if the ratio of v_s or v_o to V is larger. When equal (Mach 1) f' goes up to infinity and down to zero.

It may still be puzzling some (especially those who are thinking of the magic word 'relativity' and are remembering that not long ago in connection with Fig. 1 we satisfied ourselves — I hope — that the speed, if any, of the air made no difference) why the Doppler effect for a given relative speed between source and observer should depend in the slightest degree on which was said to be stationary and which moving. After all, that is only a convention, depending on the 'frame of reference' (as Einstein called it) one chooses. In such matters as these one usually chooses the surface of the earth. But someone else would be perfectly entitled to choose the motor bike as the origin in his frame of reference, and say it was always stationary and the observer always moving. Who has never been confused when in a train in a station, looking at another train alongside?

The explanation is that the rule about movement of the air making no difference is true only when there is no relative movement between source and observer. Although we worked out the formula for Fig. 2 on the basis of a moving source, we could equally well have treated it as a variety of the Fig. 3 situation, the only difference being the existence of a 45 m.p.h. wind blowing from O to S so that we, the observers, would feel no breeze in our faces as we were swept along, and might almost imagine we were standing still.

If it hadn't been that I find it clearer, and believed you might too, to deal with one thing at a time, I could have saved a lot of space by following the usual textbook line and deriving one formula to cover all cases — moving source, observer and medium. It is

$$f' = f\left\{\frac{V \pm v_m \pm v_o}{V \pm v_m \mp v_s}\right\}$$

You use the upper of the + and − signs for $\left\{\begin{matrix} v_m \\ v_o \\ v_s \end{matrix}\right\}$ when the $\left\{\begin{matrix} \text{medium} \\ \text{observer} \\ \text{source} \end{matrix}\right\}$ is moving towards the $\left\{\begin{matrix} \text{observer} \\ \text{source} \\ \text{observer} \end{matrix}\right\}$ For Fig. 2, v_m and v_o are zero, and when we omit them from the above formula we get the one we found for that case. Similarly for Fig. 3, where v_m and v_s are zero. But if in Fig. 3 we choose to regard the observer as stationary, then v_o is zero and $v_m = v_s$, giving the same result.

The textbooks also go into the modifications that have to be made when motion is not directly towards and away from the observer, and into what happens when observation is via a moving reflector. This last condition has practical importance with radio waves, as you may have noticed if you failed to spot the police radar in time to slow down. But I think we'd better leave radar along with radio waves for another article. Meanwhile, lest it be thought that sound Doppler is completely out of place in these pages, I will briefly describe what it is liable to do to the reproduction of sound by moving diaphragms, as in almost all loudspeakers.

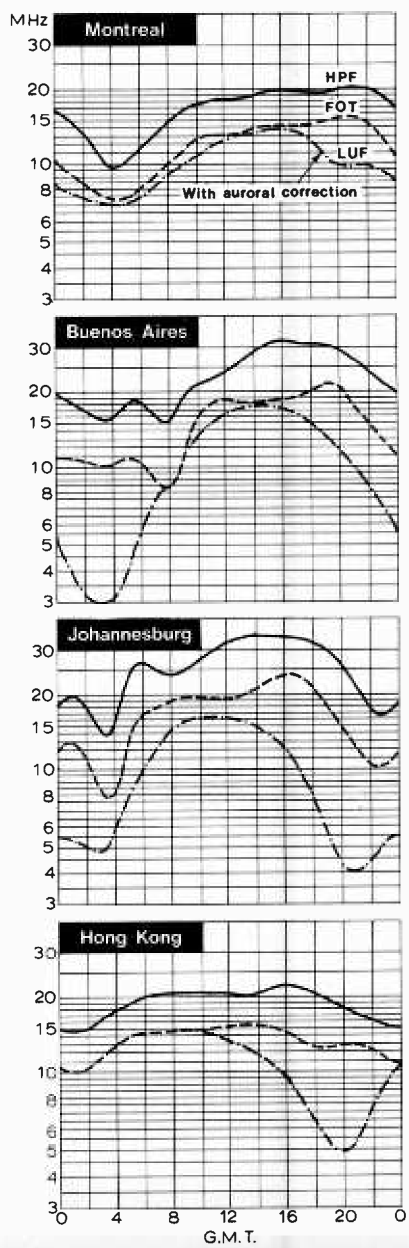
Suppose that in order to punish, in a manner befitting the offence, a motorist who was addicted to excessive use of his horn, he was condemned to sit in the driving seat, with his horn permanently on, while his car was mounted on a mechanism that kept on making it alternately surge forward with highly promising acceleration and then frustratingly retreat to the starting point. If we, in order to study his reactions, took up our position along the line of his oscillation, but well clear of its peak displacement, enough has already been said to enable us to forecast that reception of the horn's steady note would be marred by what in the context of gramophones and tape recorders would be called wow. It will perhaps help us to imagine what we would hear if, instead of his horn, the motorist was playing a flute solo or some other musical composition in the middle and upper ranges of frequency. It would be unpleasantly distorted.

This is roughly what takes place when programmes are reproduced by a single moving-coil loudspeaker. Although the amplitudes through which the diaphragm moves at the very lowest frequencies hardly compare with those in the unconventional punishment just described, they are vastly greater than at high musical frequencies — which nevertheless are much more audible. The *velocity* of the low-frequency movement could well be comparable. Frequency modulation of, say, a 1000 Hz note at 50 Hz is unpleasant.

That is one reason why in any fi system professing to be hi the reproduction of the full frequency range is usually divided up among two or more diaphragms, or the amplitude of movement is reduced by an impedance-matching enclosure, or both.

H.F. Predictions — May

The probability of a skywave path existing can be judged from the value of working frequency relative to HPF (10%) and FOT (90%). These curves are constructed by applying an ionospheric index to standard charts of ionospheric characteristics and are for practical purposes independent of equipment parameters. Actual communication however is only possible if the resulting signal-to-noise ratio exceeds a required threshold value. The number and variations of factors determining s/n ratio are such that it is not possible to draw generalized curves for the 10% and 90% probability of achieving all service thresholds. Individual cases would have the same form as the chart LUFs (which are the 90% curves for commercial telegraphy using several kilowatts, directional aerials and good sites assuming 100% skywave probability at all frequencies) but would be displaced vertically up or down. Circuit reliability is the product of skywave and threshold probabilities, for example at the intersection of FOT and LUF on the charts reliability of the stated service is predicted to be 81%.



I.E.E.E. Show New York

In America, writes a U.K. correspondent, I have always associated the wide open spaces with the prairies and the west. This year, however, the wide open spaces were very much in evidence at the I.E.E.E. show at the New York Coliseum. The number of exhibitors was down drastically; only about two thirds as many as last year. The exhibition organizers tried valiantly to boost the number of stands by including open areas, each furnished with tables and chairs, in amongst the stands and labelled 'Technical Discussion Areas'. The large vacant areas of the exhibition halls were curtained off but this did not fool anybody! Attendance, too, was very much down on previous years. This was accentuated by the earlier opening at nine o'clock. Usually it was ten o'clock before any visitors arrived. However, this did have two advantages: those who came early were able to see the exhibition in relative comfort and the majority of the visitors had definite objects in view.

Despite the smaller number of stands there were plenty of interesting exhibits. I was most surprised, however, to find, sandwiched between instrument companies, a surplus equipment dealer.

Mini-calculators were very much in evidence. The cheapest, by North American Rockwell, was intended for the consumer market. With it one could add, subtract, multiply or divide for less than a hundred dollars. Most surprisingly it used a liquid crystal display. Unlike the majority of digital displays, the liquid crystal display uses reflected light and requires very little power. For the home-constructor there was a do-it-yourself kit including all the parts necessary for a simple calculator at a cost of $98.50.

Video tape recorders were shown by Sony and Panasonic (Matsushita). While they are not quite advanced enough to make the home cine-camera obsolete, they are pointing that way. The performance of a Sony colour camera, in particular, was most impressive with very near natural colour reproduction. This camera uses a single pick-up tube, called the Trinicon, which operates on a new principle. It is basically a photoconductive, single-beam tube, but uses a striped RGB optical filter and a corresponding set of indexing electrodes. The output of the tube is a composite of the luminance signal, a coded colour signal (three-phase p.a.m.) and a coding reference signal. In the camera the luminance signal is extracted by a low-pass filter and the colour and reference signals are separated by correlation techniques; colour difference signals $R-Y$, $G-Y$ and $B-Y$ are obtained by synchronous detection and finally the luminance and colour difference signals are combined in an encoder to give a standard N.T.S.C. colour signal output.

Sony also demonstrated a 17-in version of their Trinitron colour c.r. tube in a projection television equipment. With a final anode voltage of only 25kV, it gave an adequately bright picture on a 38-in diagonal back projection screen. In addition they showed a 50-in front projection (reflecting screen) system, but this was demonstrated only in subdued light so that it was difficult to assess the brightness of the picture. Hitachi also have a new colour tube system, which they claim simplifies production and reduces the effects of beam landing error in the corners of the tube face. This is achieved by a new optical lens used in the colour phosphor deposition process in tube manufacture. The surface of the lens is divided into several hundred areas, each of which has a surface inclination that will give the exact refraction angle required for each corresponding portion of the phosphor screen. Panasonic showed equipment for 'printing ½-in video tapes from a master. The printing works on the contact principle when the master and duplicate tape are wound together onto a spool. A 'transfer' field is applied to the bifilar tape and then the tapes are rewound onto their respective reels. This process takes only 6 minutes for a 2400-ft reel and it is claimed that the master is good for several thousand prints.

While not strictly colour television, the electronic painting system shown by G.T.E. Laboratories was most interesting even if the results were not very inspiring. In essence electronic painting is achieved by 'sketching' on the screen of a television camera using a penlight. The resultant 'picture' is stored in a memory, the colours being commanded by the 'artist' and displayed on a television monitor.

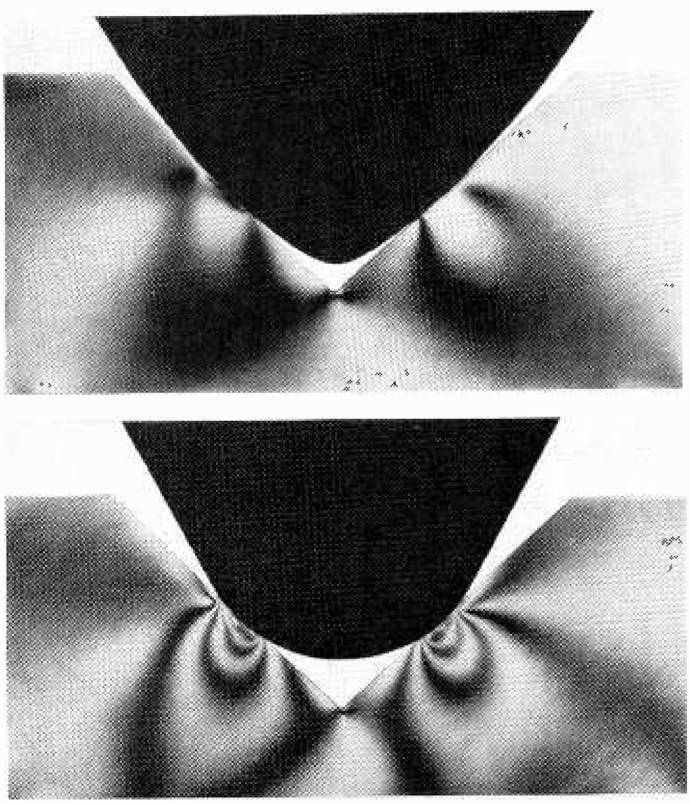
An ignition interlock system designed for General Motors was displayed by North American Rockwell. It is called the 'Phystester' and offers an interesting method of preventing a motorist from driving while drunk and yet allowing him to drive if he is sober enough to pass a test. The device operates as follows. When the driver turns his ignition key on, the word 'code' is illuminated in a display portion of the unit. The driver now has to insert, on a keyboard, a sequence of five digits which are the car's ignition code. Until he has done this the unit remains inert, so that this feature is also a theft prevention measure. If he has inserted the combination correctly, a 'set' button is illuminated. If the combination was wrong the keyboard turns off and the driver must turn the key off and on again to restart the process. The driver must then press the 'set' button, when a five-digit random number will flash on the display. The driver must now insert this number. During this the word 'brake' will be illuminated and he must put his foot on the brake until the 'brake' command is extinguished. He must then finish inserting the number. If he has done this correctly, a 'start' button is illuminated and he can start the car. If not the 'set' button lights again and he has to repeat the test. If he fails three times the word 'code' appears again and the car will not start for half an hour. The only thing wrong with this unit, remarks our correspondent, is that it includes an override switch so that the driver need only do the test if he wants to!

Miniature tubeless TV camera *Constructed by RCA Laboratories, U.S.A., this research model, measuring 2 × 2¼ × 3¼ in, uses a solid-state image sensing panel consisting of a 0.2in square m.o.s. integrated circuit with 1,408 photo-sensitive elements. The method of scanning, based on 'bucket brigade' charge transfer, was described in the March issue, p. 138. RCA's research into solid-state image sensors follows two main courses: standard silicon i.c. technology with charge-transfer scanning; and thin-film deposition of elements with scanning by x − y addressing. The thin-film method at present allows much larger numbers of photo-sensitive elements to be fabricated on a panel. So far a 512 × 512 element panel has been made, and Paul K. Weimer, leader of the research team, has stated that 'present wire-grill masking facilities would allow fabrication of integrated sensors having up to 1000 × 1000 elements on one-mil centers'.*

Letter from America

Some time ago, Electro-Voice modified their quadraphonic decoder to deal with CBS SQ records and then triumphantly advertised 'The 4-channel war is over'. The advocates of discrete systems like JVC-RCA could paraphrase John Paul Jones and reply — we have only just begun to fight! RCA have just announced that the first of their discrete discs will be released in May; the price will be the same as ordinary stereo records. Station KIOI in San Francisco has been broadcasting experimental transmissions using the Dorren 'Quadracast' system which is ideal for the JVC-RCA discs. JVC report that much progress has been made in overcoming some of the problems which many critics thought were inherent in the system. As *W.W.* readers may be aware*, the JVC discrete system uses a multiplex arrangement which requires a response up to 45 kHz. As might be expected, this not only posed recording problems (particularly at the inner grooves) but also meant that a relatively expensive pickup had to be used. Earlier JVC discs got over some of the difficulties by not using more than two-thirds of the disc — but the life of the record was still restricted due to high-frequency erasure. However, JVC developed a new stylus called the Shibata (see illustration) which gives a better contact with the record grooves. According to JVC, this stylus enables the groove wall pressure to be reduced to one-fourth and so the signal/noise ratio and amplitude response are much improved. Now, RCA have come up with a new record compound which uses multiple resins plus lubricating stabilizers which will further reduce record wear and erasure of the carrier frequency. According to W. H. Dearborn, RCA's director of operations, the new disc is fully compatible and can be played as many as 100 times with an ordinary pickup at 5 grams without significant deterioration! But of course any four-channel record system not only has to be compatible, it must be capable of being used for broadcasts and the big problem remaining is getting FCC approval. However, tests by station KIOI indicate that the Quadracast system radiates less sideband energy than ordinary two-channel stereo. Meanwhile, tests are going on in New York with two f.m. stations broadcasting discrete records and tapes. One station radiates front left and rear right, and the other front right and rear left. Not particularly compatible *and* listeners need two receivers — but the response has been remarkably good.

* See pp 486-8 October 1970 issue.



New Shibata stylus (top) developed by JVC claims to reduce record wear by reducing pressure to one-quarter of that with an elliptical stylus (bottom) and to improve phase and amplitude response. Stress patterns shown are for equal downward force.

Coming back to matrix systems, it seems that more and more manufacturers are making equipment for the CBS SQ system and the latest list includes Kenwood, Lafayette, Harman-Kardon, Sherwood and Toshiba — as well as Sony who came in first. So far about 30 SQ records are available. Records made with the Sansui and Electro-Voice matrix systems are also appearing in the shops. How do the various matrix arrangements compare in practice? Recently I spent some time listening to the available systems using their respective decoders and appropriate records. The four were Sansui, UMX-Denon, Electro-Voice and SQ. All gave good results with little to choose between them. Tests were also made for mutual compatibility — i.e. records made by system A played through system B, C and D's decoders. Conclusions were that all systems were compatible enough to give a good surround sound but in some cases location suffered. For instance, a UMX record played via an SQ decoder had the effect of moving an instrument placed at the rear extremes to a position nearer the front. Such deviations were easy enough to detect in a very heavily damped room with little reflected sound but they were heavily masked under normal domestic listening conditions.

Advocates of a discrete system speak very smugly of a 45-dB separation between channels, saying you really cannot hear a peep from channels A, B, or C if you listen closely to channel D's speaker. And it is perfectly true — you can't: but the question is, how relevent is this kind of separation? An instrument playing in one corner of your living room is not heard at a 45 dB lower level in any of the other corners — unless you have an exceptionally large house. It seems to me that the only valid case for a discrete system is the undeniable fact that it *does* give the composer a better tool, or medium — especially for electronic music. Even here, it is surprising what effects are achieved by such composers as Walter Carlos in his 'Switched-on Bach' or Mort Subotnick with 'Touch'. (Both these recordings are CBS SQ, the first using a Moog and the second a Buchla synthesizer.)

It might be asked — what are the reactions of American hi-fi enthusiasts and music-lovers to all this quadraphonic confusion. Well, no doubt about it — many prospective buyers are adopting a wait-and-see policy.

G. W. Tillett

World of Amateur Radio

More f.m. on v.h.f.?

A new group has been set up in the U.K. to promote greater use of amateur narrow-band f.m., particularly in the 2-m and 70-cm bands. Among the proposals is the adoption of 144.48 and 433.20MHz as national calling frequencies, and the allocation of zonal f.m. frequencies: 144.4 (west); 144.8 (south); 145.2 (midlands); and 145.6MHz (north). The group is recommending that both deviation and audio bandwidth should be limited to 3kHz, preferably with audio compression.

Another boost for f.m. is likely to follow the establishment of an amateur f.m. repeater station in the Cambridge area; it is understood that the M.P.T. is prepared to license its use on an experimental basis, provided that satisfactory standards can be agreed with the R.S.G.B. The use of f.m. repeaters has become a prominent feature of v.h.f. operation in a number of countries, used by both fixed and mobile stations.

The recent noticeable trend towards n.b.f.m. has been encouraged by the fact that there is less likelihood of f.m. transmissions causing interference to TV, radio or audio equipment. An interesting form of s.s.b., which takes advantage of this property of constant amplitude transmissions, is reported in use in Holland. Several amateurs, including PA0EPS, are now using infinitely clipped, phase-locked s.s.b. on 144MHz. In this sytem, a 9-MHz s.s.b. signal is infinitely clipped to provide a constant-amplitude (but spreading) signal; then another 9-MHz oscillator is phase-locked to this signal, and carefully adjusted to provide a compromise between bandwidth and intelligibility; this signal is then heterodyned to 144MHz. The resulting transmission has constant amplitude and thus shares with f.m. advantages from an interference viewpoint and yet provides the long-haul DX effectiveness of s.s.b.

QSL cards — a rethink?

From time to time amateurs question the long-established habit of sending QSL cards to 'confirm' routine contacts: the recent rises in postal charges is causing some rethinking on this topic. Yet old habits die hard, and QSL cards are still very much an integral part of the hobby. Certainly they go back (in Britain) to at least January 1921 when W. F. Corsham, G2UV, sent out his first card. The establishment in Britain in 1926 of the world's first bulk-handling QSL Bureau helped to reduce costs for those amateurs in no great hurry to obtain cards.

How many cards are sent today? Arthur Milne, G2MI, the R.S.G.B. QSL manager throughout the post-war period, estimates that about 35,000 cards per week, or some 1.75 million a year pass through this bureau alone. We suspect that the greatest volume of cards in a single bureau are those passing through Box 88, Moscow; the A.R.R.L. does not operate an integrated outgoing system.

Checking my own log (as an h.f. operator who endeavours to reply to incoming cards but normally initiates cards only for 'wanted' DX contacts) I find that roughly a third of all contacts made outside the U.K. results in the receipt of a card. The figure is about about 40% for stations in Eastern Europe. Relatively few inter-U.K. contacts result in cards — and indeed a significant proportion of British amateurs seem already largely to have opted out of routine QSL exchanges. But, by and large, the proportion of one-third has remained remarkably consistent for many years.

On the bands

Spring is traditionally the time for good long-distance conditions on the h.f. bands, and the usual flurry of activity centred around the A.R.R.L. and other DX contests. This year, the shortened BERU contest, however, resulted in no great burst of Commonwealth activity — although Caribbean activity, with VP2AAA, VP2MU, VP2LY, 8P6DR, etc, seemed on the increase; and the final hours were submerged by a Russian contest. But generally 14 and 21MHz have been providing many good openings with extremely strong North and South American signals in the late evenings on 14MHz. Even 28MHz sprang into life occasionally, and there were signs that this band is open for long-distance working more often than current activity suggests.

The U.S.S.R. '50th anniversary' group of stations have maintained an extremely high level of activity, using the various Russian prefixes. JT0AE, in Ulan Bator, in the once extremely rare Zone 23, has been active on both 14 and 21MHz c.w. with a Czech operator. Station KC4DX, on an expedition to Navassa, is expected to be active May 12th-15th.

Stew Perry, W1BB reports that the German amateur DL9KR using a 30-ft vertical aerial with umbrella top loading and 2000 ft of radials (maximum length 70 ft) on a flat roof worked all continents on 1.8 MHz within three days of erecting this aerial — his very first contact was with K2ANR, the second with ZD9BM! The Swiss station HB9CM is now also working all continents on 1.8 MHz.

The call GB2IW has been issued to the R.S.G.B. for the Intruder Watch and will be used for collating information on the activities of non-amateur stations using exclusive amateur frequencies. During the second half of 1971, more than 90 intruders were heard, a number due to spurious emissions from stations working on allocated frequencies. Regular intruders include a number of broadcasting stations between 7000 and 7100kHz, several U.S.S.R. teleprinter stations around 14,190kHz, a number of diplomatic stations around 21,100kHz, a station in Saudi Arabia on 21,105kHz, and an Italian diplomatic station on about 21,194kHz. As a result of Intruder Watch activities, assurances have been received from the Ceylon Broadcasting Corporation and the Iran National Airlines that in future their stations will avoid using 21,445kHz and 14,005kHz respectively.

In brief

The Canadian Department of Communications has agreed to drop age restrictions on amateurs; previously applicants have had to be at least 15 years old. . . . Dr J. A. Saxton, president of R.S.G.B. in 1970, has accepted an invitation to serve again as president during the Society's Diamond Jubilee Year in 1973. . . . The Chiltern mobile rally is being held on May 28 in the grounds of the estate of Sir Francis Dashwood in the village of West Wycombe, near High Wycombe, Bucks — one attraction is the annual Steam Rally in the vicinity (details from Chiltern Amateur Radio Club, P. J. Perkins, Loakes House, Loakes Park, High Wycombe; telephone High Wycombe 21612). . . . The Cheshire Homes Amateur Radio Network Fund has been closed after raising £840 to equip Cheshire Homes with communications receivers. . . . The Nottingham club will use the call GB3LEC during the Long Eaton Carnival (May 20-21). . . . The Radio Amateur Old-timers' Association has its reunion on May 5 and the B.A.R.T.G. teleprinter convention is on May 20. . . . The 10GHz activities of Des Clift, VK5CU (former G3BAK), and Barry Wallis, VK5ZMW, has resulted in a new Ausralian record of 61 miles.

Pat Hawker, G3VA

I.E.A. Exhibition

Exhibitors at the 9th international show

The international exhibition of Instruments, Electronics and Automation is to be held at Olympia, London, from 8th to 12th May and this year has nearly 600 exhibitors. Opening time is from 10 a.m. to 6 p.m. with a late night opening until 8 p.m. on the last two days. Admission is 30p but is free to overseas visitors. A list of manufacturers and distributors at the exhibition is shown below and in addition to these there will be several publishers, banks and ancillary services to the electronics industry present. Judging by the amount and variety of pre-show publicity literature we have received from exhibitors, this show should provide many new products and techniques.

The show is organized by Industrial Exhibitions Limited.

Exhibitors
A.C.B. Division
Acbars
Adams & Westlake
Adar Associates
Addo
Advance Electronics
Advance Filmcap
Ad-Yu Electronics
AEG
AIL
Air Control Installations (Chard)
Akers Electronics
Aladdin
Albert Measurements
Allard et Compagnie
Allied International
Alma Components
Alston Capacitors
American Technical Ceramics
Americon Microwave
Ampex (Great Britain)
Analog Devices
Analogic
Analytical Measurements
Andermann
Antiference
APEM
APR
APT Electronic Industries
Arcolectric Switches
Ardente Industrial Services
Arkon Instruments
Arrow-Hart
Aston Electronics
Ates Electronics
Aumann
Auriema
Austen, Charles Pumps
Automation
Auxitrol
Avco Electronics
Avel-Lindberg
Avery
Aviquipo
Avo
Aztec Instruments

Badger Meter
Bafco
Bailey Meters & Controls
Bailey Stamp
Bakelite Xylonite
Baldwin Lima Hamilton Electronics
Barden Corporation
Batley Valve
Bauch, F.W.O.
Becuwe G. & Fils
Belden Corp.
Belix Company
Bell & Howell
Belling & Lee
Bertain Associates
BICC-Burndy
Biccotest
Bifurcated & Tubular Rivet
Biomation
Birchall, D. J.
Birch-Stolec
Bird Electronic Corp.
B. & J.
B & K Laboratories
Blakeborough, J.
Bofors
Bogen, Wolfgang
Borens Fabriks
Bosch, Robert
Bourns
B. & R.
Bradley, G & E
Brady, W. H.
Brion Leroux
Bristol Automation
Britec
Britimpex
British Calibration Service
British Insulated Callenders Cables
British Manufactured Bearings
BSI
Brookdeal Electronics
Brooks Instrument
Bryand Southern Instruments
Budenberg Gauge
Burgess Micro Switch
Burnt Hill Electronics
Burr-Brown Research
Bush Beach Engineering

Callins International
Cambion Electronic Products
Cannon Electric
Carton Industries
Cedenco
Celdis
C.G.S. Resistance
Channel Electric Equipment
Chardwick Helmuth
Chemi-Circuits
Chequers Engraving
Chessell Data Recorder
Chimex Engineering
C.P. Clare Electronics
Clarke-Hess Communication Research
Claude Lyons
Clemac
Cliff Products
Coil Winding Equipment
Cole Electronics
Colvern
Comet Tool
Compteurs Schlumberger
Compulite
Computer Automation
Computing Techniques
Contraves Industrial Products
Control Data Corporation
Control & Instrumentation
Controls & Automation
Core Memories
Cornish Sign Manufacturing
Cossor Electronics
Coutant Electronics
Counting Instruments
Crater Controls
Critchley Bros.
Crogate Transformers
Crompton Parkinson
Crouzet
Croydon Precision Instrument Co.
Culton Instruments
Cwmbran Development Co.
C.Z. Scientific Instruments

Dana Electronics
Daniels Manufacturing Corp.
Data Dynamic
Data General
Data Laboratories
Datasonde
Datel Systems
Datwyler
Davall, S. & Sons
Davian (Instruments)
Davu Wire & Cables
Davy & United Instruments
Daystrom Schlumberger
Deac
Delta Metal Electronics
Denis C. & Co.
Dept. of Trade and Industry
Diamond H. Controls
Digilin
Digital Equipment
Digital Systems
Diodes Inc.
DISA
DIT-MCO International
D-Mac
Dodwell Process Controls
Doric Scientific
Double C Precision Engraving
Dymar Electronics
Dynamco
Dynacast
Dynamic Transmissions
Dyna-Quip
Dynatel

East Grinstead Electronic Components
Edgcumbe Peebles
Educational Measurements
Efco
Egen Electric
Electrical Remote Control Co.
Electricole
Electrolube
Electro Mechanisms
Electronic Associates
Electronic Flo-meters
Electronic Instruments
Electronic Machine Control
Electronic Memories
Electronic Visuals
Electrons Inc U.S.A.
Electronum
Electrovert Manufacturing Co.
Elektrobau Mulfingen
Elektronska Industrija
Elesta AG Electronik
Elga
Elgenco
Ellison Instrument
Eltromet
EMI
E & M Labs
Emtel
EMT
Endress & Hauser
Energy Conversion
Engel & Gibbs
Engineering Enterprises
English Numbering Machines
Environmental Equipment
Equipment & Services
Erg Industrial Corp.
Erie Electronics
Ernest Turner Electrical Instruments
E.S.L. Engineers
Esterline Angus
Ether
Euchner
Euro Electronic Instruments
Euro Electronic Rent
Eurogauge Co.
Eurotherm
Ever Ready
Evershed & Vignoles
Exact Electronics

Facit Office Equipment
Farnell Instruments
Feedback Instruments
Fenlow Electronics
Fieldon Electronics
Fieldtech
Filhol, J. P.
Fine Tubes
Fine Wires
Fischer & Porter
Flow Metering Instruments
Fluidyne Instrumentation
Fluke International
Foster Cambridge
Foster Transformers
Foxall Instrument Housings
Foxborough-Yoxall
F.R. Electronics

Gamma Scientific
G.D.S. Sales
G.E.C.-Elliott
G.E. Electronics
General Radio Company
Gentran
George Kent
Gerber Scientific Europe
GHZ Devices
Girling
GMT
Gordos
Gossen
Gothic Electronic Components
Gould Ionics
G.P.S.
Grace, W. G.
Greenpar Engineering
Gresham
Grubb, Parson & Co.
Grundig
Gubelin Fabrikation
Guest International
Guideline Instruments
Gulton Europe

Hans Grieshaber
Hartmann & Braunn
Harvard Apparatus
Hayden Laboratories
Heath (Gloucester)
Heimann
Hellerman
Hengstler
Hewlett-Packard
Highland Electronics
Hird-Brown
Hoffmann Charts
Holsworthy Electronics
Honeywell
Hottinger Baldwin Messtechnik
Howaldtswerke-Deutsche Werft
Howe Richardson Scale
H.S.B. Controls
Huber Continental
Hugh James Group
Hutson Industries
Hybrid Systems
Hymatic Engineering Co.

Ide, T & W
I.L.C. Data Device
Imhof-Bedco
Impectron
Industrial Instruments
Infotronics
Inland Motor Division
Instrument & Ancillaries Co.
Insuloid
Intercole Systems
International Light
Intersil Memory
Intertechnique
Introl
Iskra

Jaco International
Jermyn Industries
J.H. Associates
Jones, Walter & Co. (Engineers)
Julie Research Labs.

Kaymar Trolleys
Kay Metzeler
Kay Ray
K.D.G. Instruments
Keithley Instruments

Kempston Electrical
Kent Automation Systems
Kent Instruments
Kepco
Kerry Ultrasonics
Keyswitch Relays
Kinetrol
Kistler Instruments
Klippon Electricals
Kodak
Kongsberg Vapenfabrikk
Kovo Foreign Trade Corp.
Krohnhite
K.S.M. Electronics
Kulite Semiconductor Products
Kumag
Kywowa Electronic Instrument Co.
J. B. Kitts

Landis & Gyr
Lan Electronics
Leeds & Northrup
Leland Leroux
Lemco AG Burgdorf
Lemosa
Leonard Wadsworth
Levell Electronics
Lewis, H. K.
Licon Electronics
Lindsey, C. S.
Lion Mechanical Products
Lire
Litronix
Littelfuse
Livingston Hire
Lloyd, J. J. Instruments
Lock, A. M. & Co.
Logimetrics
London Chemical Co.
Lucas Aerospace
Lyons Instruments

Magnetic Devices
Magnus
Markem
Markovits, I.
Martron Associates
Masoneilan
Mast Development
Matrix Corp.
May Precision Components
Mayes, W. H. & Son (Windsor)
M.B. Electronics
M.B. Metals
Mechanism
Metar
Meter-Flow
Metronex
Meyer, William A.
Micro Computer System
Micro Consultants
Micro Devices
Micro Metalsmiths
Micro Movements
Miles Roystone
Millivac Instruments
Mills & Rockleys
Milton Ross Co.
Mimic Electronics
Mine Safety Appliances Co.
Mitsubishi Corporation
M KK L
M.L. Engineering
Monolithic Memories
Monsanto Chemicals
Montford Instruments
Moore Reed & Co
Motorola Semiconductors
Mullard
Multitone Electric Co.
Murex
Mycalex Instruments

National Engineering Laboratory
National Physical Laboratory
Nelson-Ross Electronics
Neoflex
Newmarket Transistors
Newport Instruments
NH Research
Nightingale Chemical
Nig Mason
Nombrex
Norgren, C. A.
Norma
Novar-Nixon
Nuclear Diodes
Nu Devices Inc.
Nuclectrohms

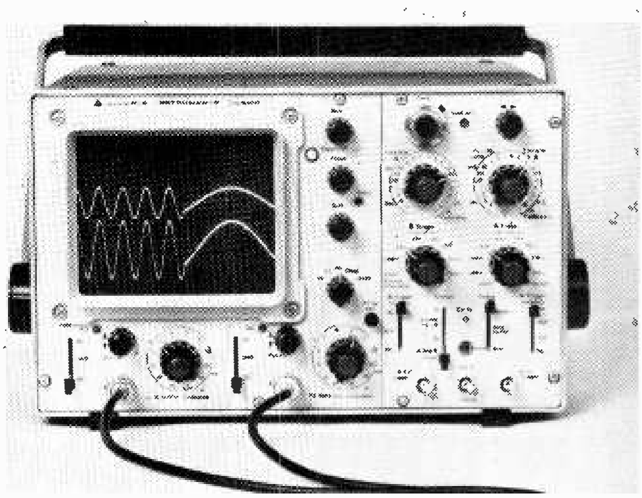
Oliver Pell Control
Olson-Horiba
Omni Spectre
Optical Works
Orbit Controls
Ortec
Oxy Fluid Controls

Panax
Pedoka
Penny & Giles
Perena
Perivale Controls
Permanoid
Philips
Plasmoulds
Platon, G. A.
Pluritec Italia
Portescap
Pozzi
Praxis-Oxford
Precision Electronic Termination
Precision Tool & Instrument Co.
Prewitt
Princeton Applied Research
Printed Motors
Proper Equipment
Prosser Scientific Instruments
P.S.B. Instruments
P.S.I.
Pye
Pyrotenax

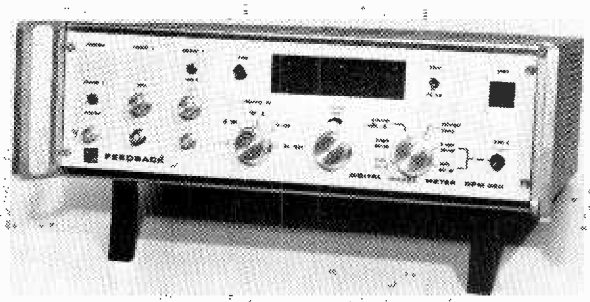
Quality Monitoring Instruments
Quickdraw

Racal
Radford Laboratory Instruments
Radiatron
Radio Aid (Engineering)
Radiometer
Rafi Raimund Finsterhoelzl
Redac
Redlake Labs
Reliance Gear
R.E.S.
Research Inc.
Research Instruments
Resistances
Reutlinger, Dr. & Sons
Reyrolle, A. & Co.
Rhodes, B. & Son
Rikadenki Kogyo
Rilton Electronics
R.K.B. Precision Products
Rockland Systems
Ron Channel Engineering Co.
Rosemount Engineering Co.
Ross Courtney
Rotronic
Roxburgh Electronics
Royal Worcester Industrial Ceramics
Rueger
Russenberger

Safety in Mines Research
Saft (U.K.)
Sage Electronics
Sangamo Weston
Saunders Electronics
Scama
Shaevitz Engineering
Schurter, H.
Schwitter Electronic
S.E. Computer Peripherals
S.E. Labs
Semicomps
Semiconductors Specialists
Sercel
Serck Audco
Service Instrument
Servo Consultants
Setaram
SFIM
SGS
Shackman Instruments
Shandon Southern Instruments
Showa Measuring Instruments
Showa Sokki
Siegert Widerstandsbau
Siemens
Sifam Electrical Instrument
Sigma
Simtec Industries
Sintrom Electronics
Sinus Electronics
S.I.R.A.
Sirco Controls
Sivers Lab
Smiths Industries
Solartron/Schlumberger
Solderstat
Sonicstore
Sonnenschein, Accumulatorenfabrik
Sontranic
Souriau et Cie
Southern Watch & Clock
South London Electrical Equipment



A dual-trace oscilloscope type OS3000 made by Advance Instruments is a 40MHz unit having a 5mV/cm sensitivity at full bandwidth.



The new Feedback Instruments digital phasemeter DPM380 operates in the range 1Hz to 100kHz and has a binary coded decimal output.

Sovirel
Spear Engineering
Special Products Distributors
Spectra-Tek
Spectronics
Sprague Electric
Stackpole
Standard Graph Sales
Institute Dr Ing Reinhardstraumann
Superior Electric Nederland
Surrey Steel Components
SWISSAP
Switchcraft
Symonds Engineering
Symonds, R. H.
Symot
Systron Donner

Tally
Tape Recorder Spares
Tau-Tron
Taylor
Techmation
Techni Drive
Tectonic
Tekdata
Tekman Electronics
Telectron
Teledyne
Telemechanics
Telford
Telsec Instruments
Tempatron
Tempil
T.E.M. Sales
Tensolite
Texcel Electronics
Texscan
Thermo Electric
Thermo Systems Inc.
Thorn Bendix
Thousand & One Lamps
T.I.L.
Tinsley, H. & Co.
T.O.A. Electronics
Tobias Venson
Topper Cases
Tormo
Torsion Balance Co. (Ireland)
Toyota
Tranchant Electronics
Transducers (CEL)
Transaco Machine
Trio
Trumeter Co.
20th Century Electronics
Tyco

Unimatic Engineers
United Electric Controls
Univel

Vactric Control Equipment
Varta Batteries
Veeder-Root
Venner
Vero Electronics
V-F Instruments
Vibration Instruments
Vidar Corporation
V.T.M. (UK)

Wade
Wandel & Goltermann (UK)
Wang Laboratories
Warren Spring Laboratory
Watanabe Instruments Corporation
Watsons Anodising
Waycom
Wayne Kerr
Weir Electronics
Weller Electric
Wendell Fabrics
Wentworth Instruments
West Hyde Developments
Westinghouse Electric
Weyfringe
Whiteley Electrical
Willsher & Quick
W.T.W.

Yokogawa Electric Works

Zellweger
Zettler
Z. & I. Aero Services

About People

The Minister of Posts and Telecommunications has appointed a committee to advise him on new regulations to protect radio and television services from interference caused by the ignition systems of petrol engines. The present regulations have been in operation since 1953. The Advisory Committee has been asked to take account of regulations, based on recommendations of the International Special Committee on Radio Interference (C.I.S.P.R.), adopted by the United Nations Economic Commission for Europe for standardizing the suppression requirements of motor vehicles. The members of the Committee are: **Sir Olliver Humphreys** (formerly director of research GEC Ltd); **A. H. Ball** (Society of Motor Manufacturers and Traders); **P. A. T. Bevan** (formerly chief engineer I.T.A.); **Mrs. P. F. Finnis** (National Federation of Women's Institutes); **G. A. Graham** (B.B.C.); **Major A. J. Jackson** (R.T.R.A.); **E. M. Lee** (Radio Industry Council); **R. A. Lovell** (Standing Joint Committee of the R.A.C., A.A. and R.S.A.C.); **W. Nethercot** (formerly assistant director Electrical Research Association); and **A. Plackett** (National Chamber of Trade).

Neil Borley, who is 36 and has been with Adcola Products Ltd for the past six years as design manager, has become design and development director. Prior to joining Adcola Mr Borley was a design engineer with Creeds. Adcola have also announced the appointment of **'Wally' Birbeck,** aged 49, as works director. He joined the company 14 years ago as works manager.

Malcolm R. King, B.Sc (Eng), M.Sc., joined the research and development staff of B & W Electronics, of Worthing, several months ago and has been mainly concerned with the design of the acoustic line employed in the DM2 loudspeaker. Mr King graduated from Southampton University with a first class honours degree in mechanical engineering and he followed graduation with an M.Sc. in advanced acoustics at the Institute of Sound and Vibration Research, Southampton.

E. S. Williams

E. S. Williams is appointed general manager of Technitron Inc., of Camberley, Surrey. During his service as a regular officer in the Royal Air Force until 1963 he spent several years at the Aeroplane and Armament Experimental Establishment, Boscombe Down. Prior to joining Technitron Mr Williams was with Kollsman Instruments Ltd initially as the technical manager and since 1967 as marketing manager.

Independent Telecommunications Consultants Ltd, of Bolton, Lancs, have appointed **L. T. Arman** to the board. Mr Arman was at one time in that part of the General Post Office which is now the Radio Regulatory Department of the Ministry of Posts and Telecommunications. He subsequently went to Pye Telecommunications Ltd and latterly was with ITT Mobile Communications Ltd. He is a member of the Minister of Posts and Telecommunications Advisory Committee on mobile radio communications.

Frank Stone, export director of A. F. Bulgin & Co. has retired after 48 years with the company. He had been a director since 1958. Mr Stone was a member of the Export Committee of the Radio & Electronic Component Manufacturers' Federation from 1947-1964 (the last year as chairman).

Maurice Ridgewell, who has been with Auriema Ltd for a little more than two years, was recently appointed sales director. Mr Ridgewell was previously sales manager at Decca Radar Instruments Division and before that was with the Elliott Group. Since last June he has been general sales manager of Auriema while still retaining special responsibility for microwave products.

R. F. Hall has been appointed marketing director of Plessey Interconnect. Mr Hall joined Plessey at Swindon in November 1970, as export executive for the former Components Group. He came from Honeywell International, Brussels, where he was director, Microswitch and Precision Components Group, Continental Europe. He was with Honeywell for more than 15 years, ten of which were spent based in Frankfurt/Main and Brussels.

Robert D. Phillips has joined Racal Instruments Ltd., of Windsor as export sales manager with special responsibilities for E.F.T.A. and East European countries. Mr Phillips was manager of the Instruments Division of B & K Instruments Ltd and has also been with both Lyons Instruments and Marconi Instruments.

Kenneth Otway has joined Ampex South Africa (Pty) Ltd, of Johannesburg, as a director. Before his appointment with Ampex, Mr Otway, who is 41, was the assistant chief engineer with Harlech Television Ltd, Bristol, where he had been employed for thirteen years. Previously he was with the B.B.C. Mr Otway has been a radio amateur for some years. His call sign is G8AFT.

The appointments of **P. D. Cowell** and **P. N. Raison** as joint managing directors and chief executives has been announced by Dubilier Ltd. They succeed **Eric Marland** who has resigned as managing director and chief executive. Mr Cowell and Mr Raison previously held senior management positions in the Electrical Products Group of AMF Inc.

Richard J. Constantine recently joined Farnell Instruments Ltd as internal sales engineer at the Wetherby Office of the company. He was formerly a radio and electronics officer with I.T.T. Marine Radio Co. Ltd. Mr Constantine was radio officer on the last commercial trip of the *Queen Elizabeth* and spent some time on the Royal Research Ship *Discovery* where he started transmitting as one of the few maritime mobile amateurs (call sign G3UGF/MM). Incidentally, Farnell Instruments have their own Amateur Radio Society (G4ADQ).

R. J. Constantine

Gerald H. David, for the past nine years with Airtech Ltd, of Haddenham, Bucks, has become managing director of the recently formed company Aerial Facilities Ltd, of Boston Place, London, N.W.1, which is primarily concerned with the siting of aerials for mobile communications. Mr David, who is 38, started his career in the Transmission Laboratory, Standard Telephones & Cables, Newport, Mon. He later moved to Cunard Steamship Co., and then to Airmec, High Wycombe. In 1963 he went to Airtech, as a project engineer, where he was latterly responsible for Electronic Division Sales.

H. A. R. (Tony) Wiggins has been appointed a director of Television Systems and Research Ltd, of Brentford, Middx, a subsidiary of CIG International Capital Corporation. His appointment follows the acquisition by T.S.R. of the Top Rank Television Division of Rank Audio Visual Ltd. Mr Wiggins was the general manager of the division and retains the same post with T.S.R. He was with Rank Audio Visual for nine years mainly in the television relay and closed-circuit television sectors and prior to that spent some years with Rediffusion Ltd.

R. N. Dawson, 36, has been appointed marketing executive of Plessey Windings at Titchfield. Mr Dawson, who joined Plessey in 1957, has been manager of the company's independent Product Assessment Laboratories for the past four years. Plessey Windings specializes in the production of coil-components.

For the second year prizes, in the form of textbooks, have been awarded by Racal to technicians trained at No. 1 Radio School, R.A.F. Locking. The two prizes, for trainees returning to the school for advanced telecommunications training, are given to students 'who have displayed outstanding effort and/or achievement during their technical training'. The prizes were presented by Air Chief Marshal Sir Francis Fogarty on March 1st to junior technician **M. H. C. Burgess** (R.A.F. St. Athan) and junior technician **A. M. Robertson** (R.A.F. Oakington).

E. M. Wareham (Measuring Systems) Ltd, of Letchworth, Hertfordshire, makers of electronic instruments, have appointed **John Pritchard,** M.I.E.E., as commercial manager. Mr Pritchard, aged 42, was previously sales manager with Texas Instruments. Before that he spent 12 years with Hawker-Siddeley, where he was sales manager.

New Products

Autoranging digital multimeter

Model 167, battery-powered digital multimeter from Keithley Instruments incorporates a $3\frac{1}{2}$ digit l.e.d. display in a hand-held probe. The probe can be slotted into the front panel for bench use. Ranging is automatic with overload protection up to 1200V on any voltage range. The meter will measure voltage from 1mV to 1kV d.c. and from 1mV to 500V a.c. and resistance from 1Ω to 2MΩ. The case and probe body are moulded in ABS plastic. Keithley Instruments Ltd, 1 Boulton Road, Reading, RG2 ONL.

WW332 for further details

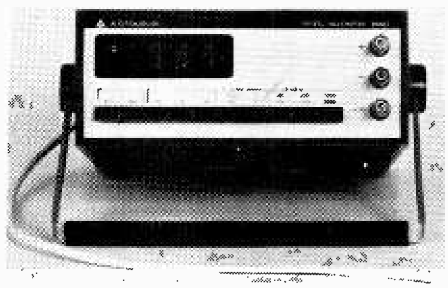
I.C. audio power amplifiers

Impectron have available a 20W audio power amplifier, the EHD-AP4211, made by Matsushita Electric. It is a hybrid thick-film i.c. requiring no external adjustments for temperature compensation, power source fluctuations or feedback. The output can be accidentally short circuited for up to two minutes without any adverse effects. The ambient operating temperature range is -10 to $+60$°C. At 25°C, 1kHz and load resistance 8Ω the voltage gain is 30 $\pm$2dB. Quiescent current is typically 30mA. Frequency range is given as 20Hz to 100kHz, and signal-to-noise ratio 100dB. Input impedance is 50kΩ. Price £3.97 (1 — 25). A 30W version, the EHD-AP 4311, is also available. Impectron Ltd, Impectron House, 23-31 King Street, London W.3.

WW326 for further details

Digital multimeter

A number of improvements to the digital multimeter type DMM2 have been incorporated into the new Advance DMM3. Current ranges for a.c. and d.c. are now in-built — 200nA to 2A d.c. and 200μA to 2A a.c. Voltage and resistance ranges are 200mV to 1kV, and 200Ω to 2MΩ with a resolution of 0.1Ω at 200Ω. Accuracy has been increased by using an improved



attenuator; on d.c. measurements for example accuracy is $\pm$0.1% of reading $\pm$0.1% of full scale. Common-mode rejection is 90dB of the 50-Hz common-mode voltage (with 1kΩ unbalance on the d.c. ranges). Automatic polarity indication is included. Cost has been kept to £130 by using large-scale circuit integration. Advance Electronics Ltd, Raynham Road, Bishop's Stortford, Herts.

WW330 for further details

High-current bridge rectifier

The SDA 129 bridge rectifier made by Solid State Devices Inc, and available from Concorde Instrument Company is capable of handling 25A at 420V (55°C). The p.i.v. is 600V. Peak forward surge (1 cycle) is 150A. Maximum thermal impedance to mounting surface is 1.5°C/W. Concorde Instrument Company, 28 Cricklewood Broadway, London N.W.2.

WW305 for further details

Solid-state relays

A range of static relays — the Solac series from Darpan Controls — has four-terminals. They are solid-state d.c. input/a.c. output switches of medium sensitivity; typically 5mW at 5V. They are logic compatible and 5A types cost £3.25 in quantity. Printed-circuit board-mounting, and base-mounting versions are available. All units operate over the range 6V to 250V r.m.s. 50/60Hz and differ only in the input voltage and output current ratings. Versions are available for nominal input supplies from 5V to 24V and output currents from 2A to 15A. Feedback into logic supplies is low, typically less than 5mV, and all devices have very high insulation resistance between input and output circuits. Darpan Controls Ltd, Bridge Mills, Derby Road, Long Eaton, Nottingham.

WW323 for further details

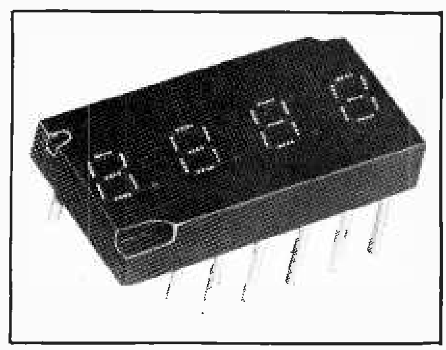
Image intensifier

The F4747 tube from ITT is an image intensifier using a microchannel plate amplifier and proximity focusing. The tube has an 18mm cathode. Input voltage is only 2.7V to give a luminous gain of between 5000 and 35000 and a resolution of 25 to 28 line pairs. Recovery after exposure to a bright-light-source is virtually instantaneous. A fibre-optic image inverter can be specified to bring the output of this single-stage device into correct aspect. Other tubes in the series will be available shortly with 23, 25, 46 and 75mm diameter cathodes, but without any increase in length. ITT Components Group Europe, Valve Product Division, Brixham Road, Paignton, Devon.

WW325 for further details

L.E.D. numeric displays

Guest International have available a series of numeric displays using light emitting gallium arsenide phosphide diodes. Designated the Data Lite 30 Series, these displays are available either with single numerals or encapsulated in groups of three or four. The numerals are 0.125in high and have a high luminance of typically 680cd/m^2 with

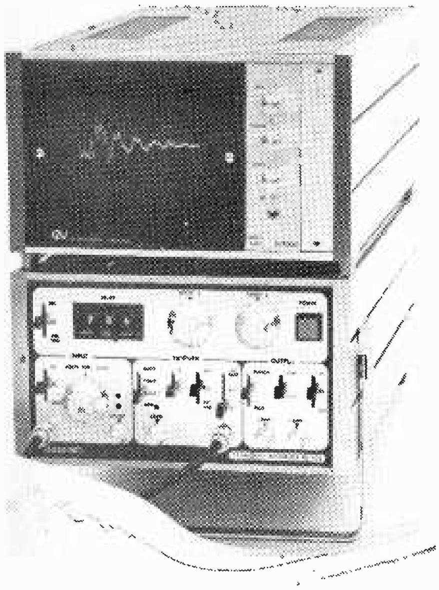


a forward current of 5mA. Displays can be mounted to give ten digits per square inch of panel space. The units are i.c. compatible and the leads fit a 0.10 inch grid. Guest International Ltd, Nicholas House, Brigstock Road, Thornton Heath, Surrey CR4 7JA.

WW316 for further details

High-speed transient recorder

Data Laboratories model DL905 transient recorder allows single voltage transients, repetitive waveforms, or very slowly changing signals, to be recorded, and held indefinitely in its digital memory. The captured signal may be viewed on an

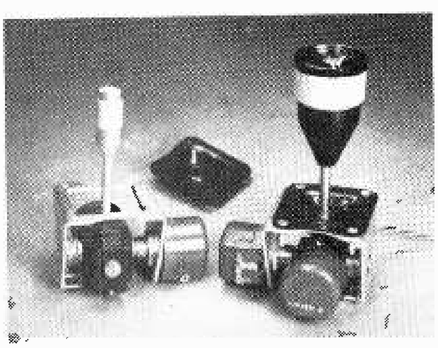


oscilloscope, a permanent record produced on a chart recorder, and a digital readout obtained. Signals preceding a trigger, or occurring spontaneously, can be recorded using a special recording mode, and control of two independent timebases in the switched timebase mode permits the sweep rate to be changed during recording. Connected to a computer the transient recorder becomes a high-speed data acquisition peripheral. It may also be used to improve the frequency response of slower recorders or signal processors, such as signal averagers and spectrum analysers. The instrument consists of a 5MHz 8-bit analogue-to-digital converter, coupled to a 1024 word m.o.s. shift-register memory. Front panel controls for the selection of input gain, sweep time and delay, and trigger conditions, define the period over which the input signal is digitized and stored. A digital-to-analogue converter reconstructs the stored information as an analogue signal. Data Laboratories Ltd, 28 Wates Way, Mitcham, Surrey CR4 4HR.

WW322 for further details

Joysticks and servos

Single, dual and triple axis joystick controls, and a range of miniature d.c. servo mechanisms are available from Flight Link Control. Overall resolution from joystick to servo output can be better than $\pm 1°$, and output torque ranges from 24oz in



upwards. Supply voltages are in the range 6 to 12V. Joysticks are priced from £4 to £10 each, and servos from £10 upwards. In addition to the standard range, non-standard units can be supplied even in very small quantities. Flight Link Control Ltd, Bristow Works, Bristow Road, Hounslow, Middx.

WW327 for further details

Microwave transistor

The V-578 n-p-n silicon transistor from Nippon Electric Company, and available from Impectron, is designed for low-noise broadband amplifier applications to above 5GHz. The device has an f_T of 6.5GHz with a n.f. of 2.7dB at 2GHz and 6dB at 4GHz, and a gain of 13dB at 2GHz and 7.5dB at 4GHz. Microwave Division, Impectron Ltd, Impectron House, 23-31 King Street, London W.3.

WW329 for further details

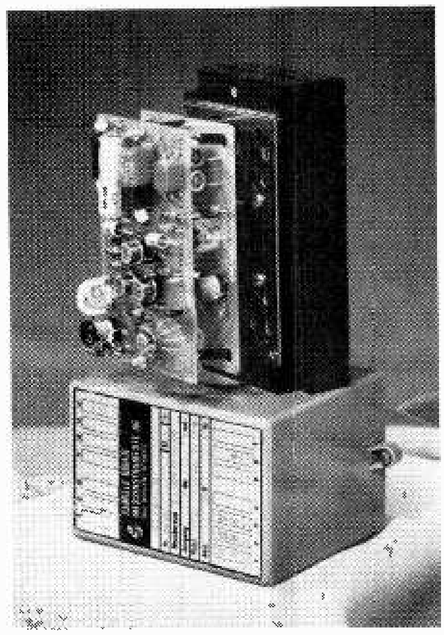
Attache case v.h.f./u.h.f. receiving system

Astro Communication Laboratory's electronically or manually tuned v.h.f./u.h.f. receiving system, type SR-217, has been designed for fixed station and portable applications where small size and light weight are desirable. The complete system fits inside a standard size attache case for ease of handling. The system is mains-powered (20W). A range of 30 to 1000MHz can be scanned in seven base bands, and displayed on a cathode-ray tube. Also, a single base band can be manually tuned to display the characteristics of any single frequency therein. Instantaneous switching between bands and search/manual tuning modes shows any a.m., f.m., c.w., and p.a.m. signals present across the entire frequency range. Seven swept-frequency tuners operate with the signal display unit. The tuners are set by voltage-controlled capacity diodes for electronic scanning, and by potentiometer regulation for manual tuning. R.F. circuits are swept synchronously with the horizontal control voltage for the signal display unit to display linearly frequency from left (low) to right (high) on the c.r.t. screen. An intensity-modulated marker calibrates the c.r.t. display to aid signal location by manual tuning. The three selectable i.f. filter bandwidths are 75kHz, 300kHz and 3MHz. A b.f.o. is built into the 75kHz bandwidth and other i.f. filters available below bandwidths of 150kHz. Astro Communication Laboratory (U.K.), Tower Street, Coventry, CV1 1JP.

WW328 for further details

Solid state converters

Rayleigh Instruments have introduced a range of Camille Bauer solid-state transducers, type Sineax, for conversion of electrical variables into proportional

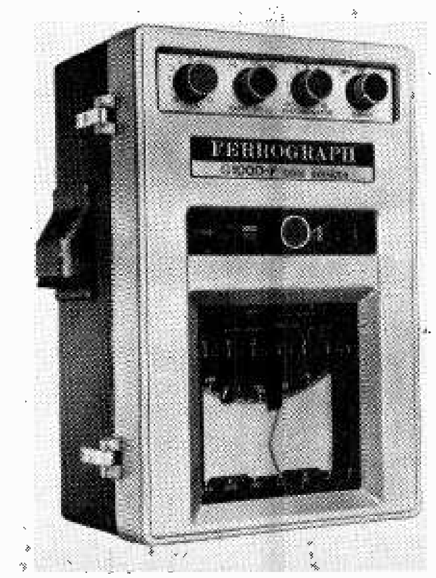


analogue direct current. Transducers conform to DIN specifications and use time division multiplication (pulse width modulation). Active and reactive power, phase angle, frequency, voltage, current, mV signals and resistance variables are converted into a direct current of 0-5, 0-10, 0-20 or 4-20mA impressed current and therefore load independent, to a maximum stipulated limit. Power converters can have unipolar, bipolar symmetrical and bipolar asymmetrical output currents. Rayleigh Instruments Ltd, 271 Kiln Road, Benfleet, Essex, SS7 1RX.

WW330 for further details

173-fathom echo sounder

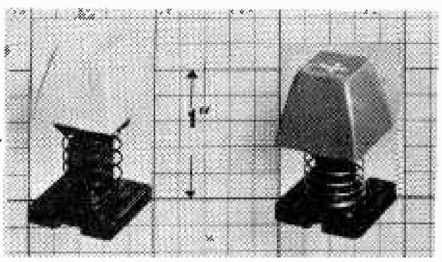
The G1000 inkless echo sounder from Ferrograph has a high operating frequency combined with short pulse length to give high resolving power, supplemented by the inclusion of white-line bottom discrimination. Five scales in all are used to cover the 1,040 ft range (173 fathoms), all overlapping by 40ft. A 'search' position of the range switch permits all scales to be superimposed on a single trace, to avoid



constant switching between scales when searching. Dry recording paper (in 15 metre rolls, 115mm wide) is used and electrically marked. Recording speeds of 0.25, 0.50, 1.00 and 1.50m/h are available: for most economical use the slowest speed is used until interesting soundings are seen. At this point the speed can be changed, without stopping the sounder, in order to improve the lateral resolution. The standard 12V model has a current consumption of 2A, which is lower for the two higher working voltages possible. External control boxes must be ordered as extras for 24 and 32V operation. A variety of dual transducer units is available, each with separate transmit and receive transducers, contained in either bronze or steel casings. Dimensions of the instrument are 340 × 216 × 203mm and the weight of the sounder is 8kg. Price £175. The Ferrograph Co. Ltd, The Hyde, Edgware Road, Colindale, London N.W.9.
WW331 for further details

Mercury keyswitches

Mercutronic keyboard-array keyswitches from Tekdata employ a mercury switching element. The life of each switch is a guaranteed 25 million operations and the modules are impervious to dust, noise and



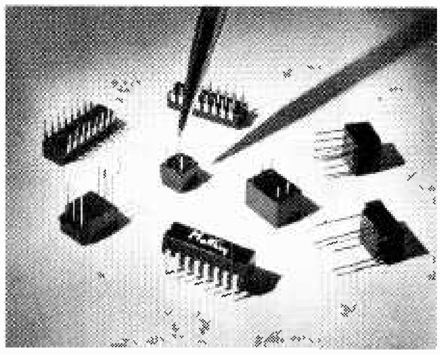
electrostatic charges. Encoding is performed by using m.o.s. chips or diode arrays. A range of standard keyboard layouts is available. Tekdata (Trading) Ltd, Pentagon House, Bucknall New Road, Hanley, Stoke-on-Trent, Staffs. ST1 2BA.
WW321 for further details

Epicyclic reduction drive

Jackson Brothers (London) have introduced a miniature epicyclic 10:1 reduction drive. With an overall length of 23mm and a high torque (576gm/cm), the new 10:1 drive is suitable for use when space behind the front control panel is restricted. Jackson Brothers (London) Ltd, Kingsway, Waddon, Croydon, CR9 4DG.
WW320 for further details

Miniature pulse transformers

Matthey Printed Products have introduced a range of low-power pulse-transformers. Several standard package styles have been selected to accommodate a wide range of characteristics. A design service is available. The pulse transformers can deliver up to 5W at the output terminals with primary inductances from 10μH



to 10mH. Matthey Printed Products Ltd, William Clowes Street, Burslem, Stoke-on-Trent ST6 3AT.
WW319 for further details

Encapsulated reed relays

A range of miniature encapsulated reed relays available from Ralcom Electromechanical Components will switch at frequencies up to 2000Hz. They have a 12VA handling capacity with current up to 600mA and voltage up to 220V. Life is claimed to be 10^7-10^9 operations. Coils for 2-48V operation are available. Standard types are also available with axial connections, and suppressor diodes may be fitted with no increase in size. If no suppressor is fitted, the coil terminations are brought out both ends of the package. Prices range from 96p for normally open single-contact devices in quantity. Ralcom Electromechanical Components, 15 Milldown Road, Goring, Reading, Berks.
WW326 for further details

'Flat' rotary switch

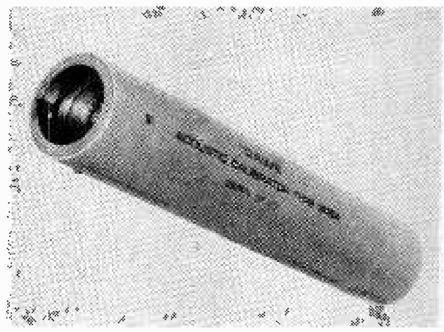
The SB11 flat switch from ITT can have up to five sections with one, two, three, four or six circuits per section. Detent angle is 30° and up to 12 locating stops can be accommodated. The switch is designed specifically for printed-circuit mounting. Mounting dimensions are 29 × 11mm and the switching mode is non-shorting or shorting. The switch is rated at 3VA for operation at 60V. ITT Components Group Europe, Electromechanical Product Division, West Road, Harlow, Essex.
WW324 for further details

Low-power l.e.d. display

The Hewlett-Packard 5082-7405 light-emitting-diode display requires 7mW per digit, and efficiency is increased through the use of an integral moulded lens which enlarges the five digit cluster to a height of 0.112in — it is 0.75in wide. The cluster is in a standard 14-pin d.i.p. Each digit includes a decimal point. A red dye, incorporated in the plastic, filters out all visible light except the 655 nanometer wavelength emitted by the diode. Good contrast is thus assured. In addition, portions of the lead frame are darkened to reduce reflections. The cluster uses 13 pins for the five digits, and the display is i.c. compatible. Price £15.80 (1-15). Hewlett-Packard Ltd, 224 Bath Road, Slough, Bucks. SL1 4DS.
WW316 for further details

Sound pressure calibrator

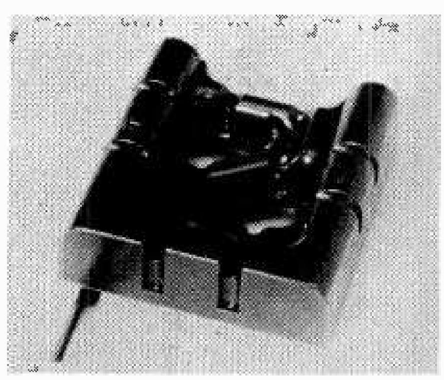
Acoustic calibrator type 1418A from Dawe Instruments is suited to the field calibration of sound level meters and microphone systems. The microphone is inserted directly into the calibrator which automatically produces a 1kHz signal at a sound pressure level of 94dB. (At this frequency the A, B, C and D weighted responses of a sound level meter coincide so that all functions may be directly



calibrated.) The new design is claimed to be more convenient to use, particularly in areas of high noise level, and has an accuracy of ±0.4dB. A built-in battery provides the power. There are no manual controls. Dawe Instruments Ltd, Concord Road, Western Avenue, London W3 OSD.
WW318 for further details

Voltage doubler for line output stages

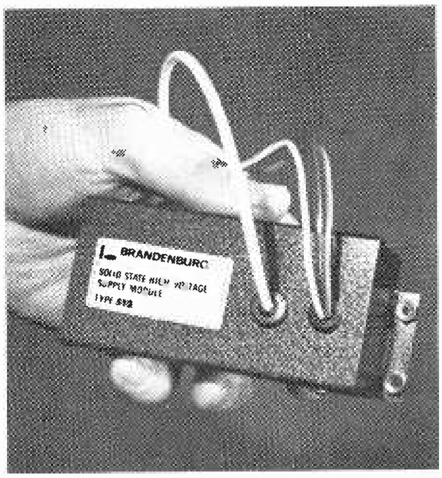
Selenium rectifier voltage doubler type TVK 22, from Siemens, has been developed to reduce failures in black and white television line-output stages due to the high voltage on the coil of the flyback transformer — typically 18 to 29kV at the live terminal. The voltage doubler reduces the size of the winding and the voltage stress; voltage on the live terminal of the coil will now be only 9 to 10kV.



The new unit consists of an encapsulated assembly of the three selenium rectifier sticks and two capacitors in a voltage-doubler circuit. Siemens (United Kingdom) Ltd, Great West House, Great West Road, Brentford, Middx.
WW315 for further details

Oscilloscope e.h.t. module

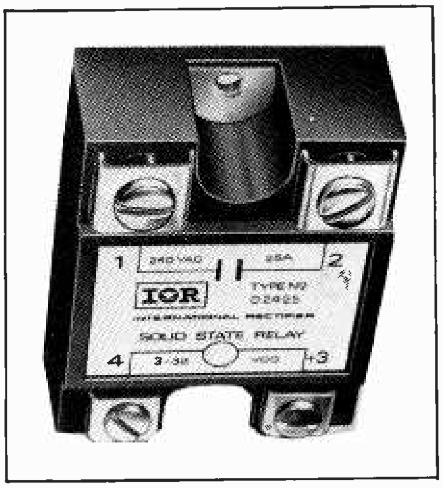
A modular high-voltage supply for oscilloscope tubes has been announced by Brandenburg. The unit provides between 4 and 7kV at a maximum of 120μA. It is capable of withstanding an output short-circuit for up to 10 seconds without



damage. The impregnated e.h.t. components are screened in a metal can with overall dimensions of 50 × 38 × 150mm. The price is £15 singly, with reductions for quantity orders. Brandenburg Ltd, 939 London Road, Thornton Heath, Surrey CR4 6JE.
WW313 for further details

Solid-state relay for a.c. power

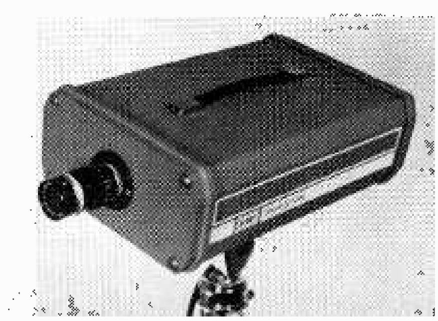
International Rectifier have produced an all-semiconductor, four-terminal, fully isolated ($10^{10}\Omega$, 8pF), a.c. power relay capable of switching up to 25A a.c. The input required is 1 to 3V d.c. into 1.5kΩ with one model or 90 to 240V a.c. or 45 to 200V d.c. into 30kΩ with a second model. Power is switched by a pair of s.c.rs connected in parallel; the s.c.r. junctions being mounted on an electrically insulating, but thermally conducting, substrate. The metalization pattern for interconnection is extended to the edges of the substrate to improve heat transfer. The substrate itself is mounted on an aluminium sheet which transfers heat to a heat sink. Because this sheet is electrically insulated more than one relay can share the same heat sink without having to bother about insulation problems. The s.c.rs are controlled by a modified zero crossing switch which eliminates r.f.i. The switch is arranged to fire the s.c.r., if there is an ON control signal, when the voltage across the two output terminals exceeds about ± 10V. It is this voltage that is used to power the internal circuitry of the relay. This means that a separate



power supply connection is not required but there is about a 4mA drain through the load when the relay is in the off condition. The s.c.r. control circuit derives its input signal from a photo-diode. The input signal is connected via an internal current-limiting resistor (d.c. input), or a resistor and a diode (a.c./d.c. input) to a light-emitting diode. Hence coupling between input and output is by a light beam. The relay can be mounted straight on to a heat sink with the four connections made by lugs, 'quick-connect' push-on-terminals or simply by wrapping wires round the terminal screws. Alternatively the relay can be screwed directly on to a printed interconnection pattern. Three basic models are available, 2, 10 and 15A, but there are variations, a.c./d.c. input etc. International Rectifier, Hurst Green, Oxted, Surrey.
WW312 for further details

Rugged camera

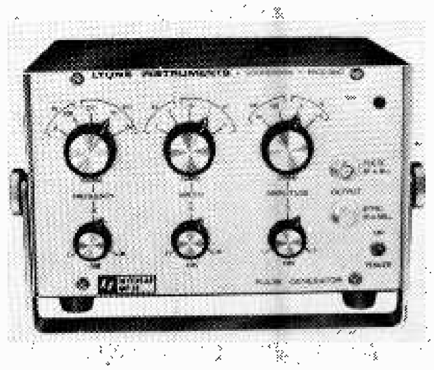
E.M.I. have introduced a new television camera, called Surveyor, which while not being in the low-cost bracket does not achieve the performance of broadcast quality cameras nor their price. It has been designed for general purpose closed-circuit TV and surveillance work and to this end is sealed against the elements, can be serviced very quickly and has no user controls. There are two external connections; one is the mains lead and the other carries the output signal (1V composite video, white positive, for 75Ω termination, to E.B.U. or E.I.A. recommendations, 625/50 or 525/60 standards). The camera has automatic light level compensation and the output will not vary more than 3dB over



the range 18 to 90,000 lux (60% reflectance) for a lens aperture of f1.9. The signal-to-noise ratio is 40dB and geometric distortion (due to the raster) will not exceed 2% of picture height or width. The pick-up tube is an E.M.I. 9745, 9747 or equivalent electrostatic vidicon. A full range of accessories is available including pan and tilt heads, control units etc. E.M.I. Electronics Ltd, Television Equipment Division, Hayes, Middlesex.
WW311 for further details

Low-cost pulse generator

The 'Interlab' type MP25 pulse generator from Lyons Instruments has a p.r.f. of 25Hz to 2.5MHz, pulse width of 100ns to 10ms and output of 50mV to 5V into 50Ω (10V max e.m.f.) with better than 10ns



rise and fall times. Width and p.r.f. jitter are both below 0.1%. There is a +2V sync. output, and the main output stage is capable of 100% duty cycle. Lyons Instruments Ltd, Hoddesdon, Herts.
WW310 for further details

Dual-in-line reed relay

A dual-in-line reed relay in a standard i.c. package is available from Plessey. Single or twin make contacts in a 14-pin package can be supplied. The products are designated type CRK for one-make operation, and CRL for two-make, and are available in the full range of popular coil voltages from 4.8 to 24V. Minimum operating powers are approximately 110mW for the CRK and 290mW for the CRL. Any standard coil and switch terminations can be provided across the fourteen pins, and the rhodium contacts are rated at 3W switching with maximum ratings of 250mA or 28V d.c. Temperature range is −50° to 125°C. Plessey Switching Controls, Abbey Works, Titchfield, Fareham, Hants.
WW308 for further details

Twin-channel audio filter

The Kemo VBF/3 is a two-channel filter variable over a range of any five decades between 0.01Hz and 10kHz. Each channel can be used as a high- or low-pass fourth order Butterworth filter giving a skirt

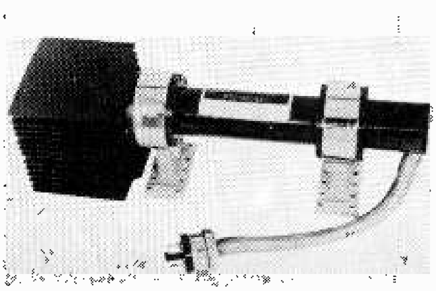
attenuation of 24dB/octave. The channels can be arranged in series or parallel by front-panel function-switch so forming a single band-pass or band-reject filter or low-pass or high-pass unit having 48dB/octave slope. The VBF/3 is one of a wide range of variable filters ranging in price from £130 to £1,500. Kemo (Consultants) Ltd, 42 Chancery Lane, Beckenham, Kent.
WW305 for further details

Solid-state marine radars

A range of solid-state marine radars has been announced by Decca. Superseding the Decca Transar range, there are three groups: 'group 16' has 16in displays with a choice of true or relative motion, 3cm or 10cm solid-state transceivers, and 6, 9 or 12ft scanners; 'group 12' has 12in displays with a choice of true or relative motion and the Decca anti-collision display, 3cm or 10cm solid-state transceivers and similar scanners; and 'group 9' has 9in displays, relative motion on 3cm wavelength, 25kW or 3kW solid-state transceivers, and 4, 6 or 9ft scanners. Thus there are three display sizes, three types of presentation and two wavelengths available. The solid-state transceivers (magnetrons are the only thermionic devices remaining) and solid-state displays provide an exceptionally bright, high-quality picture, with better overall performance. Decca Radar Ltd, Decca House, Albert Embankment, London SE1 7SW.
WW314 for further details

Travelling-wave tubes

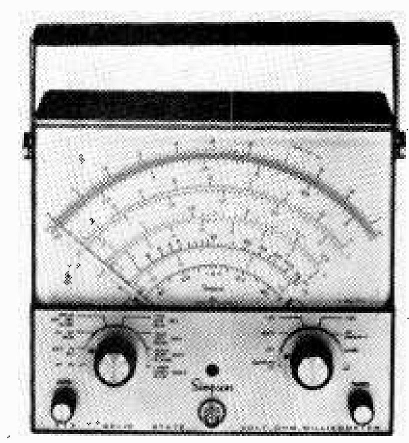
Three travelling-wave tubes from EMI-Varian are for use as output tubes in microwave relay systems. The convection cooled periodic permanent magnet focused tubes have low a.m. to p.m. conversion and are available in the following



configurations: VTX-2612A1 delivering c.w. output of 10W at a nominal gain of 41dB over the frequency range of 10.7 to 11.7GHz; VTM-2613A1 delivering c.w. output of 5W with nominal gain of 42dB over the frequency range of 12.20 to 13.25GHz; and VTU-2614A1 delivering c.w. output of 2.5W at a nominal gain of 37dB over a frequency range of 14.40 to 15.25GHz. Life expectancy for all tubes in the series is better than 20,000 hours. The tubes operate with the collector depressed to 1.9kV (2.0kV for the VTU-2614A1) for greater efficiency. No adjustment is needed for input and output impedance matching. EMI-Varian Ltd, Hayes, Middlesex.
WW301 for further details

Electronic multimeter

A varistor-protected multimeter from Bach-Simpson, model 313, employs a 7-in mirror scale. The instrument (unusually) measures r.m.s. and peak-to-peak values, and has a dB scale.



Specification:

voltage (a.c./d.c.)	10mV—1000V 8 ranges (11MΩ d.c., 10MΩ a.c.)
current (d.c.)	10μA— 1A (5 ranges)
resistance	1—500MΩ (7 ranges)

Battery condition can be checked. Price £59.50. Bach-Simpson Ltd, 331 Uxbridge Road, Rickmansworth, Herts WD3 2DS.
WW309 for further details.

Light-sensitive potentiometer

A potentiometer which is controlled by light, and called the Photentiometer, is available from Photain Controls. It consists of a strip of photoconductive material (either cadmium sulphide or cadmium selenide is available) 21mm long by 0.5mm wide mounted on a ceramic strip 25 × 3.5 × 2.2mm and is complete with lead wires. When connected to a suitable input voltage (up to 25V d.c.) and subjected to a moving strip of light it provides an output voltage which is directly proportional to the position of the light on the strip of photoconductive material. Photain Controls Ltd, Randalls Road, Leatherhead, Surrey.
WW306 for further details

Polystyrene capacitors for vertical mounting

The A615 polystyrene capacitor from Salford Electrical Instruments has a flexible skirt extending below the body and encircling the leads which holds the capacitor away from the board as it is being soldered into position. The range of the capacitors is 1 – 10μF (100V d.c.). Salford Electrical Instruments Ltd, Peel Works, Barton Lane, Eccles, Manchester M30 0HL.
WW308 for further details

Upgraded multimeter

Multimeter type S7A from Guest International has been improved in several ways. The input impedance has been raised from 20kΩ/V to 200kΩ/V up to 30V; to 10MΩ/V from 100V to 1000V, and a.c. frequency response is now specified as less than –1dB at 100kHz and –3dB at 10Hz and 200kHz. New accessories offered are a current shunt, and a voltage probe, which extend the ranges to 10A and 30kV respectively. A carrying case is available. Guest International Ltd, Nicholas House, Brigstock Road, Thornton Heath, Surrey CR4 7JA.
WW314 for further details

Chassis-mounted heat dissipators

Two new-style heat dissipators for use in chassis-mounted TO-3 transistor socket applications are now available from Souriau Lectropon. Incorporating a TO-3 semiconductor hole pattern with clearance holes for the socket hardware, the new dissipators may be removed from the chassis mounting surface or circuit board along with the semiconductor without disturbing the transistor socket or its hardware. Designated HP1-TO3-44CB or HP3-TO3-44CB the dissipators are 2¼in and 3in square respectively. Typical power dissipation for the HP1 type is 20W in natural convection with a 90° case temperature rise. Souriau Lectropon Ltd, Shirley Avenue, Vale Road, Windsor, Berks.
WW307 for further details

Literature Received

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

ACTIVE DEVICES

We have received three publications from Teledyne Philbrick, St. Peter's House, Chichester, Sussex.
Product price list (from Feb. 1972) WW401
Product guide 1972 .... WW402
Wall chart on op-amps and non-linear applications WW403

Photoconductive cells, automatic lamp failure indicator, s.c.rs and triacs made by Quantrol Electronics Inc. are the subject of several data sheets received from Joseph Lucas (Electrical) Ltd, Electronics Product group, Mere Green Road, Sutton Coldfield, Warwickshire WW404

Specifications of a range of precision potentiometers made by Sakae, of Japan, are given in publication 7103. U.K. agents Techni Measure, 3 The Green, Chalfont St. Giles, Bucks. WW405

The series of tri-state (DM series) logic circuits from National Semiconductors (UK) Ltd, Larkfield Industrial Estate, Greenock, Scotland, are described in a set of leaflets which include information on characteristics and applications WW406

Two publications from Ferranti Ltd, Gem Mill, Chadderton, Oldham, Lancs, are:
Ten supplementary sheets to the handbook 'Semiconductor Devices' WW407
Report No. 14: Applications of Ferranti Silicon Opto-electronic Devices WW408

GD127 is a data sheet describing a range of microwave noise generators employing silicon diodes. Nore Electric Company Ltd, 461 Southchurch Road, Southend-on-Sea, Essex SS1 2PL WW409

PASSIVE DEVICES

Speed sensing relays (series SR) are described in a leaflet we have received from MTE Electronics Ltd, Leigh-on-Sea, Essex SS9 5LS WW410

We have two publications from the London Electrical Manufacturing Co. Ltd, Beavor Lane, Hammersmith, London W.6:
A chart of capacitor dielectric characteristics WW411
A capacitor product summary leaflet ... WW412

A leaflet gives full specifications of a range of polycarbonate capacitors (100 to 10,000pF ±10%) from Ashcroft Electronics Ltd, Ashcroft Road, Cirencester GL7 1QY WW413

'Low profile, miniature p.c. socket' is the title of a leaflet describing the development and specifications of the Minisert socket for p.c. boards. Berg Electronics NV, Helftheuvelweg 1, P.O. Box 2060, 's-Hertogenbosch, Holland WW414

Details of 37 different reed switches are given in a catalogue which contains application notes and a fold-out specification sheet. FR Electronics, Switching Components Group, Wimborne, Dorset WW415

Topaz, a range of electro-magnetic two-, three- and four-pole relays, available for chassis or p.c. mounting, are described in a leaflet from Londex Ltd, Electrical and Electronic Controls, P.O. Box 79, 207 Anerley Road, London SE20 8EW WW416

We have received a leaflet describing wire wrapping contacts for all Pye MRAC series connectors. Pye Connectors Ltd, Hitchin Street, Biggleswade, Bedfordshire WW417

Murex Ltd, Rainham, Essex RM13 9DP, have issued two publications:
Magnets for reed switches (Publication M32) WW418
Murex Magnets: performance data (Publication M33) WW419

APPLICATION NOTES

We have two publications from ITT Semiconductors, Footscray, Sidcup, Kent.
Interface circuits necessary between m.o.s. and bipolar digital i.cs due to the incompatibility of logic levels WW420
The use of fast thyristors for 110° colour tube line scanning (reference to type BT119 and BT120 thyristors) WW421

'Numerical Indicator Tubes', is the title of application note No. TP1285, sent to us by Mullard Ltd, Mullard House, Torrington Place, London WC1E 7HD WW422

EQUIPMENT

A leaflet describing the range of Jermyn Industries i.c. sockets, including all mechanical and electrical specifications, is issued by their distributors Intel Connectors Ltd, Henlow Trading Estate, Henlow, Beds. WW423

The Sipel Express Circuit System (equipment for production of single- or double-sided p.c. boards) and the Anglade AF9 Microsoltec System (for mounting flat packs onto p.c. boards with no pre-tinning) are described in a leaflet. Lauriestone Electronics Ltd, 7 Stepfield, Witham, Essex. WW424

A product information sheet describes two Saunders hand-held test probes. One for injection of 100ns t.t.l./d.t.l. compatible pulses and one for visual voltage readout. Both are for use with digital systems. Celdis Ltd, 37/39 Loverock Road, Reading, Berks RG3 1ED WW425

An audio product catalogue from Partridge Electronics, 21-25 Hart Road, Thundersley, Benfleet, Essex, describes a series of panel-mounted modules WW426

Wire stripping and cutting machine accessories made by Eubank Engineering Co. is the subject of a product leaflet from Automation Ltd, Bessemer Road, Basingstoke, Hants WW427

Publication PS1 describes a range of rack-mounting kits made by Farnell Instruments Ltd, Sandbeck Way, Wetherby, Yorkshire, LS22 4DH ... WW428

We have received catalogue No. 27 specifying all the aerials, transmission lines and associated products made by Andrew Antenna Systems, Lochgelly, Fife WW429

Two publications sent to us by Shure Electronics Ltd, 84 Blackfriars Road, London S.E.1, are:
'Studio microphones' (describes 11 microphones for studio use) WW430
'Microphone circuitry' (describes the range of equipment available associated with microphone usage) WW431

The Tau-tron Series 525 programmable data generator is described in a leaflet. The generator can operate from 1 bit per second to over 3Mb/sec. Lyons Instruments Ltd, Hoddesdon, Herts WW432

We have a leaflet on the Agastat electromechanical, electronic and fluid power control instruments. Amerace Esna Ltd, Chantry Road, Kempston, Bedford WW433

The T1 series of micro-miniature incandescent lamps (tubular, approx. 3mm diameter) are specified on a data sheet from Ragan Precision Industries Inc., 9 Porete Avenue, N. Arlington, N.J. 07032, U.S.A. WW434

B & K Laboratories Ltd, Cross Lances Road, Hounslow TW3 2AE, have made several equipment price changes which are in the February 1972 price list WW435

'A guide to your oscilloscope requirements' is a brochure detailing a range of oscilloscopes and accessories with notes on applications and measurement techniques. SE Laboratories (Engineering) Ltd, North Feltham Trading Estate, Feltham, Middlesex WW436

Marconi Instruments Ltd, St. Albans, Herts, have sent us a booklet 'The Quiet Ones' describing the range of signal generators, TF2011, TF2012, TF2013 .... WW437

GENERAL INFORMATION

Details are given in a leaflet of a course 'Science and People' which starts in September 1972 at Edge Hill College of Education, St. Helens Road, Ormskirk, Lancashire L39 4QP. The course is intended as an alternative route into teaching for sixth formers and suitably qualified men and women.

We have received details of three three-day courses at Twickenham College of Technology, Egerton Road, Twickenham, Middlesex. Commencement dates are given.
New Materials (16th May 1972)
Electronic Engineering for Mechanical Engineers, (6th June 1972)
Production Testing (25th April 1972)

We have received a copy of the Public Address Engineers directory, which includes details of product lists, and the Association's testing standards. Association of Public Address Engineers Ltd, 6 Conduit Street, London W1R 9TG ..... Price 50p

Four publications covering I.E.C. recommendations have been sent to us by the British Standards Institution, 2 Park Street, London W1A 2BS.
Publication 244-5A. First supplement to Publication 244-5 (1971). Methods of measurement for radio transmitters Part 5 .... Price £4
Publication 382. Analogue pneumatic signal for process control systems .... Price 65p
Publication 381. Analogue d.c. current signals for process control systems .... Price 65p
Publication 255-3. Electrical relays Part 3. .... Price £7.50

Channel allocations to u.h.f. television transmitting stations in the U.K. are given in leaflet 4003 (14) issued by the Engineering Information Department, B.B.C., Broadcasting House, London W1A 1AA

Hatfield Instruments Ltd, Burrington Way, Plymouth, Devon PL5 3LZ, have sent us the 'Short-form catalogue 1972' containing information on their products WW438

Three leaflets entitled 'What's new from magnetic' give details of a range of push-button switches, keyboards and motors. Magnetic Devices Ltd, Exning Road, Newmarket, Suffolk WW439