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"To promote the advancement
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Electronic Engineering in Ocean Technology

ONE of the most significant things about the Institution's Conference which is to be held at Swansea on 21st–24th September 1970, is its title. The 1966 Conference at Southampton was called 'Electronic Engineering in Oceanography'. The present Organizing Committee thought that it could detect a very important change in thinking after this gap of four years; there is now a much more marked interest in the applied, technological and commercial side of the oceans. Four years ago, ocean technology was only just emerging as a subject in its own right; now it is established. Admittedly it hardly figures yet in university and college courses; one cannot take a degree in it—but even this will soon change. Industry and Government are conscious of the emergence of this field: there is a Government interdepartmental Committee on Marine Technology; firms in numerous countries are active in marine dredging, underwater oil and gas operations, underwater construction, underwater instrumentation including submersible vehicles, and so on. At the exhibition and conference 'Oceanology '69' held at Brighton in February 1969, nearly 2000 people registered. Ocean Technology has arrived.

In this field of Ocean Technology, electronic engineering plays a leading part. All underwater operations (and many surface operations too) require proper instrumentation and most of this is electronic in whole or in part. Although this is also largely true of the pure science field of Oceanography, and the reasons which made the 1966 Conference a success still apply, the Committee believed that the Institution of Electronic and Radio Engineers should take its full and proper part in the new and bigger field of Ocean Technology, and it is expected that the papers and discussion at Swansea will demonstrate the correctness of this judgment. Since the field is now so much enlarged as compared with the 1966 Conference, it was felt necessary to specify special topics in electronic engineering on which the Conference would concentrate. These are the gathering, transmission processing and display of information from the ocean, and the list of sessions and indication of the papers set out in last month's *Journal* show the organization and coverage of the subject matter. The final programme now available from the Institution includes 39 papers in the six main sessions.

Electronic systems for ocean technology have to operate at sea under harsh conditions and care has been taken to include some discussion of these problems in several papers. To improve understanding of some of the practical problems, it has been arranged that a research vessel, the R.R.S. *John Murray*, operated by the Natural Environment Research Council, and a naval survey ship, H.M.S. *Hecate*, will be open to inspection at Swansea during the Conference. The increasing use of computers in ships, aimed at reducing both time delay associated with shore processing of information and the excessive amount of data otherwise required to be taken back to the shore base, will also receive special attention; some of this equipment is in fact in the ships which will be at Swansea.

One of the themes stressed in the 1966 Conference was the problem of food production from the oceans and fishing applications will receive due attention. Paradoxically in these days of unattended instrumentation there is renewed interest in 'underwater man' and diving, submersibles and habitats all bring problems in electronic engineering which are to be discussed at the Conference.

The programme as a whole should, therefore, provide an excellent opportunity for scientists and engineers of many disciplines, whether as users or designers, to gain an up-to-date picture of a very significant area of ocean technology.

D. G. TUCKER

INSTITUTION NOTICES

Birthday Honours

Council's congratulations have been sent to Mr. John O'Connor on his appointment as a Member of the Fourth Class of the Royal Victorian Order (M.V.O.) in Her Majesty's Birthday Honours List published in June. First elected an Associate Member of the Institution in 1953 and recently re-instated as a Member, Mr. O'Connor is Chief Wireless Engineer (Field Services) with the Directorate of Telecommunications of the Home Office.

Revised Editions of I.E.R.E. Conference Proceedings

The final Proceedings of two I.E.R.E. Conferences have been revised and are now available. These are:

'Thick Film Technology'

(Held in London in April 1968)

'Electronic Weighing'

(Held in London in October 1968)

The prices of these Proceedings which comprise 22 and 14 papers respectively and extend to over 200 pages, are £3 10s. 0d. each post free. Orders, preferably accompanied by remittances, should be sent to: Publications Department, I.E.R.E., 9 Bedford Square, London, WC1B 3RG.

Obituary

The Council has learned with regret of the sudden death in June last at Ndola, Zambia, of Mr. George A. Ashley (Graduate); he was 59 years of age. Since 1966 Mr. Ashley had held the appointment of Lecturer in Technical Studies at the Evelyn Hone College of Further Education, Lusaka; last October his Department was transferred to a new college in Ndola. Before going to Zambia Mr. Ashley was an Assistant Lecturer at the City of Bath Technical College. Registered as a Student of the Institution in 1941 while serving in the R.A.F., Mr. Ashley was elected an Associate in 1949 and was transferred to Graduate in 1964.

The British United Provident Association

Owing to increased administrative commitments it has become necessary to change the arrangements for handling the registrations of subscribers to the Institution's B.U.P.A. group. With effect from the next renewal date—1st November 1970—subscriptions will be paid direct to B.U.P.A. and as the Institution is asked to do no more than give its name and goodwill to the Group, all correspondence will be dealt with by the Cardiff Branch of B.U.P.A.

Group subscribers will appreciate that this change-over will entail signing a new Banker's Order for their

subscriptions and cancelling existing Banker's Orders. They are asked to assist by completing and returning to the Cardiff Branch of B.U.P.A. the new Banker's Order form which will be sent to them in September.

Conference on Computers for Analysis and Control in Medical and Biological Research

A Conference on 'Computers for Analysis and Control in Medical and Biological Research' will be held at the University of Sheffield, on 7th–9th September 1971. It is being organized by the Control and Automation Division of the Institution of Electrical Engineers in association with the Institution of Electronic and Radio Engineers, the Institute of Electrical and Electronics Engineers (U.K. and Republic of Ireland Section), the Institute of Mathematics and its Applications, the Biological Engineering Society and the British Computer Society.

The Conference is intended to interest those already working in the medical and biological field, where computers are used for research purposes and to provide discussion amongst those interested in the following topics:

Models of physiological systems and processes, control systems, harmodynamic muscular, respiratory, biochemical systems, ionic equilibrium.

On-line computers for control of experiments, neurophysiology, behavioural sciences etc.

Analysis of data from on-line data acquisition, e.g., e.c.g., unit potentials etc., pattern recognition, cardiac output measurements.

Clinical research in medicine, pathology, on-line control in clinical chemistry, statistical techniques.

Offers of contributions for the Conference programme are invited and intending authors should submit a 250-word synopsis to the I.E.E. Conference Department before 5th November 1970.

Further details and registration forms will be available in due course from the Manager, Conference Department, I.E.E., Savoy Place, London WC2R 0BL.

Reprints of Journal Papers

Reprints are prepared of all papers published in the *Journal* and copies may be obtained from the Institution, price 5s. 0d. each (post free). Requests for reprints may be made using the form which is included in the end pages of most issues of the *Journal*. It is particularly asked that remittances be sent with orders to avoid book-keeping entries and thus reduce handling costs.

volts under the influence of the reference voltage, the output voltage at any time T_2 will be

$$v = v_1 - \frac{V_r T_2}{RC} \quad \dots\dots(2)$$

where

$$T_2 = \text{ramp-down time}$$

$$V_r = \text{reference voltage.}$$

When $v = 0$ and substituting for v_1

$$\frac{T_2}{T_1} = \frac{V_{in}}{V_r} \quad \dots\dots(3)$$

This shows that the ratio of the two integration times is directly proportional only to the ratio of the input voltage and reference voltage, therefore the result has no dependence whatsoever on the RC time-constant of the integrator. Now the number finally stored in the counter will be

$$N_2 = T_2 f \quad \dots\dots(4)$$

where f = clock frequency.

Also during the ramp-up time the count accumulated $N_1 = T_1 f$. Substituting for T_1 and T_2 in equation (3):

$$N_2 = \frac{V_{in} N_1}{V_r} \quad \dots\dots(5)$$

Therefore the final count stored in the counter is directly proportional to the input voltage, as the other two factors in the equation, N_1 and V_r , are constants. Hence the voltage measured is, as stated earlier, not only independent of the time-constant of the integrator, but is also independent of the clock frequency. This basic principle shows that it should be possible to make a very accurate and stable digital voltmeter. However, unless extreme care is taken in the design, arrangement of logic, and the peripheral equipment necessary to organize this technique into a full digital multi-range system, the simple basic advantages can easily be eroded away.

2. Main Sources of Errors in a Dual Ramp Voltmeter and Consideration of their Effects

The main sources of error are the following:

- (1) Variation in resistance of the semiconductor switches used to switch the integrator between the input voltage and the reference voltage.
- (2) Voltage drift of input, integrator, and zero detect amplifiers.
- (3) Transients injected by the integrator input switches.
- (4) Non-synchronization of the clock to the ramp waveforms.

- (5) Instability of the reference voltage.
- (6) Temperature coefficient and instability of the gain determining resistors in the input amplifier and attenuators.
- (7) Short-term stability of the clock frequency.
- (8) Linearity of ramp.
- (9) Series and common mode interference signals.

If care is not taken with all of these items they can add up and spoil the expected performance of the d.v.m. It is possible, however, to design a dual ramp voltmeter, with an overall accuracy of $\pm 0.01\%$, a temperature coefficient of less than 10 parts in $10^6 \text{ deg}^{-1} \text{ C}$ and a zero stability of less than one digit for a change of 20 deg C, plus the zero drift due to the input amplifier.

In order to achieve this it is necessary to hold the temperature coefficient of items that are under the control of the design, to within ± 1 part in $10^6 \text{ deg}^{-1} \text{ C}$.

2.1. Variations in Resistance of the Semiconductor Switches

2.1.1. The effects of variation in 'on' resistance

Junction f.e.t. switches are normally used for this purpose. The variation in the 'on' resistance of these f.e.t. switches must be small compared with the integrator resistor. Figure 1 shows these switches connected into the integrator. The simplest solution to this problem would be to make the value of resistor R very high. However, the integrating current is determined by this resistance and the reference voltage. This current must be large compared to the input current of the integrator amplifier, or else the computed ramp rate of the integrator will be in error. Therefore, there are two conflicting requirements.

- (a) To keep the resistance R high, to minimize the effects of the variation in the 'on' resistance of the f.e.t. switches.
- (b) To keep the resistance R low, to increase the integrating current thereby reducing the errors caused by the input current to the integrator amplifier.

One solution to this problem would be to use a chopper-stabilized amplifier for the integrator. With care, this type of amplifier can be designed to give an input current of less than 10^{-11} amps, allowing the integrator resistor R to be made high enough for the f.e.t. switches to present no problems in their 'on' state. This is a very expensive solution to the problem and if it were used in a digital voltmeter design, would probably result in an unacceptably high price for the instrument.

Another solution to the problem is to use a f.e.t. as the input device for the integrator and types with

a worst case gate current of 10^{-9} A at 50°C are available. Using f.e.t.s it will be necessary for some form of voltage drift correction to be applied, which can be achieved by using a time-sharing technique which is described later.

Now the maximum allowable input current to the integrator for a worst case error will be given by the expression

$$I_c \geq \frac{100I_e}{E} \quad \dots\dots(6)$$

where

E = worst case % error

I_e = offset current of the integrator amplifier

I_c = computing current of integrator flowing in R .

The value of the integrator resistance R will be

$$R = \frac{V_r}{I_c} \quad \dots\dots(7)$$

Therefore

$$R \geq \frac{EV_r}{100I_e} \quad \dots\dots(8)$$

Taking an example let

$$I_e \leq 10^{-9} \text{ A}$$

$$V_r = 10 \text{ V}$$

$$E \leq 0.001\%$$

Then

$$R \leq \frac{0.001 \times 10}{100 \times 10^{-9}} = 100 \text{ k}\Omega$$

In this example it can be seen that a change in resistance of 1 ohm in the f.e.t.s 'on' resistance will cause a change of 0.001% to the integrating current. However, if both S1 and S2 'on' resistance changes by the same amount and in the same direction then the errors will cancel, therefore it is the differential change of resistance between the two f.e.t.s which will cause an error. The main cause of change in the 'on' resistance will be temperature variations and the change is approximately $1\% \text{ deg}^{-1}\text{C}$. Therefore taking the worst case temperature coefficient for this function to be 1 part in $10^6 \text{ deg}^{-1}\text{C}$ then the maximum differential temperature coefficient of the two f.e.t.s will have to be $0.1 \Omega \text{ deg}^{-1}\text{C}$, namely one millionth of 100 k Ω . Also any change in the source resistance feeding them must be kept to within this limit.

Taking the example above again and letting the 'on' resistance of the f.e.t. be 50Ω then the temperature coefficient will be $0.5 \Omega \text{ deg}^{-1}\text{C}$, therefore for the 1 part in $10^6 \text{ deg}^{-1}\text{C}$ specified the matching between the two f.e.t.s will have to be better than 5:1; this is readily achievable. These errors will affect the

calibration accuracy but will have no effect on the zero stability.

2.1.2. The effects of the 'off' resistance of the integrator switching f.e.t.s

During the 'off' state there will be two main sources of error. The drain to source leakage, which will depend on the f.e.t.s 'off' resistance, and the drain-source voltage. In the case of S1 (Fig. 1) this will be the input voltage and in the case of S2 it will be the reference voltage.

The second source of error will be the gate-source leakage. When one of the f.e.t.s is switched to the 'off' state, the other f.e.t. is switched to the 'on' state, therefore the leakage current will flow through the low resistance of the 'on' f.e.t. and the source resistance of the voltage generator feeding it. This, in the case of the reference voltage, will be the output resistance of the Zener diode, and in the case of the signal voltage, the output resistance of the input amplifier.

In practice the voltage developed by the leakage current flowing in this resistance will be very small, compared with either the reference voltage or the output voltage from the input amplifier. For example, for a worst case error of 0.001%, a 10 V reference and a leakage current of 10^{-9} A, the combined output impedance could be

$$\frac{10 \times 0.001}{100 \times 10^{-9}} = 100 \text{ k}\Omega$$

The output resistance of the amplifier and voltage reference must be considerably less than this figure, and to meet the requirements stated in 2.1.1.

2.2. Voltage Drift of Input, Integrator, and Comparator Amplifiers

2.2.1. Effects of voltage drift of the input amplifier

The purpose of the input amplifier is to convert the input voltage to an amplitude which is within the scale of the reference voltage and to convert the input impedance of the integrator, to a high impedance of at least 1000 M Ω for a good quality d.v.m. The performance requirement for this amplifier, regarding voltage drift, will be dependent on the maximum sensitivity required for the instrument. The effect of this drift will be reduced as the amplifier is switched to less sensitive ranges, therefore no one figure can be stated for the overall drift effect of the input amplifier on the instrument. This is why manufacturers often specify a temperature coefficient for the instrument in parts in $10^6 \text{ deg}^{-1}\text{C}$ plus an extra incremental voltage drift for temperature change. This incremental voltage drift, which is usually specified in $\mu\text{V deg}^{-1}\text{C}$, will also include any thermal e.m.f. generated in the input range switching circuitry of the instrument.

2.2.2. The effects of voltage drift of the integrator amplifier

Voltage drift in the integrator amplifier will have the effect of directly altering the ramp rate thus changing the calibration of the voltmeter.

Equation (5) states that $N_2 = V_{in}N_1/V_r$ and this assumes that V_{in} and V_r are the voltages switched across the integrator resistor R1 (Fig. 2). However if there is drift in the integrator amplifier, then the virtual earth will not be at zero volts but will be at the drift voltage. Also as V_r is of the opposite polarity to V_{in} then in one case this drift voltage will add and in the other case it will subtract. Let the voltage drift of the integrator amplifier be V_e .

Then

$$N_2 = \left[\frac{V_{in} + V_e}{V_r - V_e} \right] N_1 \quad \dots\dots(9)$$

The full formula of error caused by this is given by:

$$E = \left[1 - \frac{V_r(V_{in} + V_e)}{V_{in}(V_r - V_e)} \right] \times 100\% \quad \dots\dots(10)$$

But at full scale where $V_{in} = V_r$

$$E = \frac{2V_e}{V_r} \times 100\% \quad \dots\dots(11)$$

Taking the example where $V_r = 10$ volts and the maximum allowable error is to be 1 part in $10^6 \text{ deg}^{-1}\text{C}$, i.e. $0.0001\% \text{ deg}^{-1}\text{C}$.

Then

$$V_e \leq \frac{EV_r}{200} = \frac{0.0001 \times 10}{200}$$

$$V_e \leq 5 \mu\text{V deg}^{-1}\text{C} \quad \dots\dots(12)$$

At zero the error will look like an input voltage to the integrator of V_e . The zero drift will depend on the least significant bit size referred to the input of the integrator. For the example taken, this will be

$$\frac{V_r}{N_1} = \frac{10}{20,000} = 0.5 \text{ mV} \quad \dots\dots(13)$$

To meet the requirement of not more than one bit zero shift for 20 degC temperature change the zero stability of the integrator will need to be

$$\frac{500 \mu\text{V}}{20} \text{ deg}^{-1}\text{C}, = 25 \mu\text{Vdeg}^{-1}\text{C}.$$

2.2.3. Effects of voltage drift in the zero detect amplifier

Voltage drift in the zero detect amplifier will result in zero drift but will not cause calibration errors. The maximum allowable voltage drift of this amplifier will depend on the full scale voltage swing of the integrator amplifier. The higher the voltage swing, the higher will be the allowable voltage drift of the zero detect amplifier.

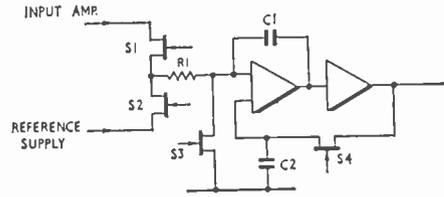


Fig. 2. Circuit for correction of drift in integrator and zero detect amplifiers.

It is not always economically possible to provide a very high voltage output from the integrator amplifier, and if a nominal swing is taken to say 10 V , then the same calculation for voltage drift will apply when calculating the allowable zero drift in the integrator amplifier, namely $25 \mu\text{V deg}^{-1}\text{C}$, but as stated later there are other factors affecting the zero stability, therefore a drift figure of about $10 \mu\text{V deg}^{-1}\text{C}$ will be required.

2.2.4. Time sharing drift correction of the integrator and zero detect amplifier

The drift of both the integrator and zero detect amplifiers can be corrected by the use of time sharing techniques. This is possible because the dual ramp voltmeter is not a continuous process and time can be taken between readings to correct drift.

Figure 2 shows a circuit which drift corrects both integrator and zero detect amplifiers together. After a digitization has taken place S3 and S4 are switched into their 'on' state, therefore the drift of both amplifiers is stored on C2. Provided there is no voltage developed across S3, all of the drift which would cause the ramp rate to alter, and cause calibration errors, will be cancelled. During the drift correct period the output voltage of the zero detect amplifier will be at the voltage developed across C2, which will be the drift voltage referred to the input of the two amplifiers. Therefore, drift of the zero detect amplifier referred to its input will be the voltage across C2 divided by the gain of the zero detect amplifier. If the input to the integrator amplifier uses a pair of f.e.t.s, connected as a long-tailed pair then a drift figure for these of $1 \text{ mV deg}^{-1}\text{C}$ is easily achievable. The gain of the zero detect amplifier need then be only 100 to achieve the required drift figure of $10 \mu\text{V deg}^{-1}\text{C}$. The gain of this amplifier will have to be made higher for other reasons.

The main causes of error with this technique are decay of the voltage stored on C2 during the digitization period of the instrument, transients injected when operating the f.e.t. switches, stray current flowing through S3 during the drift correct period, and thermal e.m.f.s generated across S3.

If there is a drain on the voltage stored on C2 during the digitization period, this will look like a small ramp voltage which will add to or subtract from the ramp function, thereby causing calibration and zero errors. Therefore this drift must be kept to less than one bit, referred to the input of the integrator, over the digitization period. The main source of decay for the capacitor will be the input current to the integrator amplifier and the leakage across the f.e.t. S4.

As stated earlier, when using f.e.t.s this can be kept to within 10^{-9} A over the working temperature of the instrument. It can be easily calculated that a 1 μ F capacitor is more than adequate to keep the decay to less than one bit for a 40 ms sample period.

2.2.5. Switching errors in time shared amplifiers

When switching S3 and S4 from the 'on' to the 'off' state the gate-drain capacitance will cause the switching voltage to be transferred into C1 and C2 respectively.

The amplitude of this error voltage will be approximately

$$V_e \approx \frac{V_s C_{GD}}{C_1} \quad \dots\dots(14)$$

where

V_e = error voltage

V_s = amplitude of the gate switching waveform

C_{GD} = drain-source capacitance of the f.e.t. + strays.

Now taking the example again where one bit is equal to 0.5 mV let

$$C_1 = 0.2 \mu\text{F}$$

$$C_{GD} = 5 \text{ pF}$$

$$V_s = 10 \text{ V}$$

$$V_e = \frac{5 \times 10^{-12} \times 10}{0.2 \times 10^{-6}} = 0.25 \text{ mV}$$

This occurs at the output of the integrator amplifier and with the constants taken represents a half-bit error. However, the change in the error should be small; the main variation would be changes in the drain-source capacitance with temperature, this would have to change by a factor of 2:1 to give a half-bit zero error. The initial half-bit error can of course be balanced out in the initial setting-up of the instrument. This error could be reduced by increasing C1 but the value of this capacitance is determined by the integrating current through R_1 , the output swing of the integrator amplifier, and the computing time. The optimum value for R_1 has already been determined, therefore the value of C1 is fixed. The zero error caused by the f.e.t. transients injection into the drift correcting storage capacitor C2 will be an

order lower because C2 can be an order higher than C1.

2.3. Transients Injected by the Integrator Input Switches and Switching Reference for Polarity Decisions

2.3.1. Error caused by integrator input switches

When the f.e.t. S1 and S2 (Fig. 2) are switched they will inject a transient into the integrator and cause an error voltage to be developed across C1, the magnitude of which will depend on whether the f.e.t. is being switched to the 'on' state or to the 'off' state. If it is being switched 'on' the error is in fact negligible, but for completeness this error will be determined.

The gate switching waveform injects a transient via the drain-gate capacitance. This decays exponentially through the output resistance of the input amplifier, in the case of S1 and the output resistance of the Zener diode in the case of S2.

The total error voltage at the output of the integrator will be:

$$dV = \frac{1}{R_1 C_1} \int_0^{\tau} V_g e^{-t/RC} dt \quad \dots\dots(16)$$

For $\tau = \infty$

$$dV = \frac{RC_{GD}}{R_1 C_1} V_g \quad \dots\dots(17)$$

where

dV = total error voltage at the output of the integrated amplifier

R = output resistance of either the input amplifier or the reference source

C_{GD} = gate drain capacitance

R_1 = integrator integrating resistor

C_1 = integrator integrating capacitor

V_g = f.e.t. gate switching voltage.

Now let

$$\text{f.e.t. gate switching voltage } V = 10 \text{ V}$$

$$\text{f.e.t. drain-gate capacitance } C = 5 \text{ pF}$$

$$\text{source resistance } R = 100 \Omega$$

$$R_1 = 100 \text{ k}\Omega$$

$$C_1 = 0.2 \mu\text{F}$$

Then

$$dV = \frac{5 \times 10^{-12} \times 100 \times 10}{100 \times 10^3 \times 0.2 \times 10^{-6}} = 0.25 \mu\text{V}.$$

As stated earlier this error is negligible provided that the source resistance is kept low.

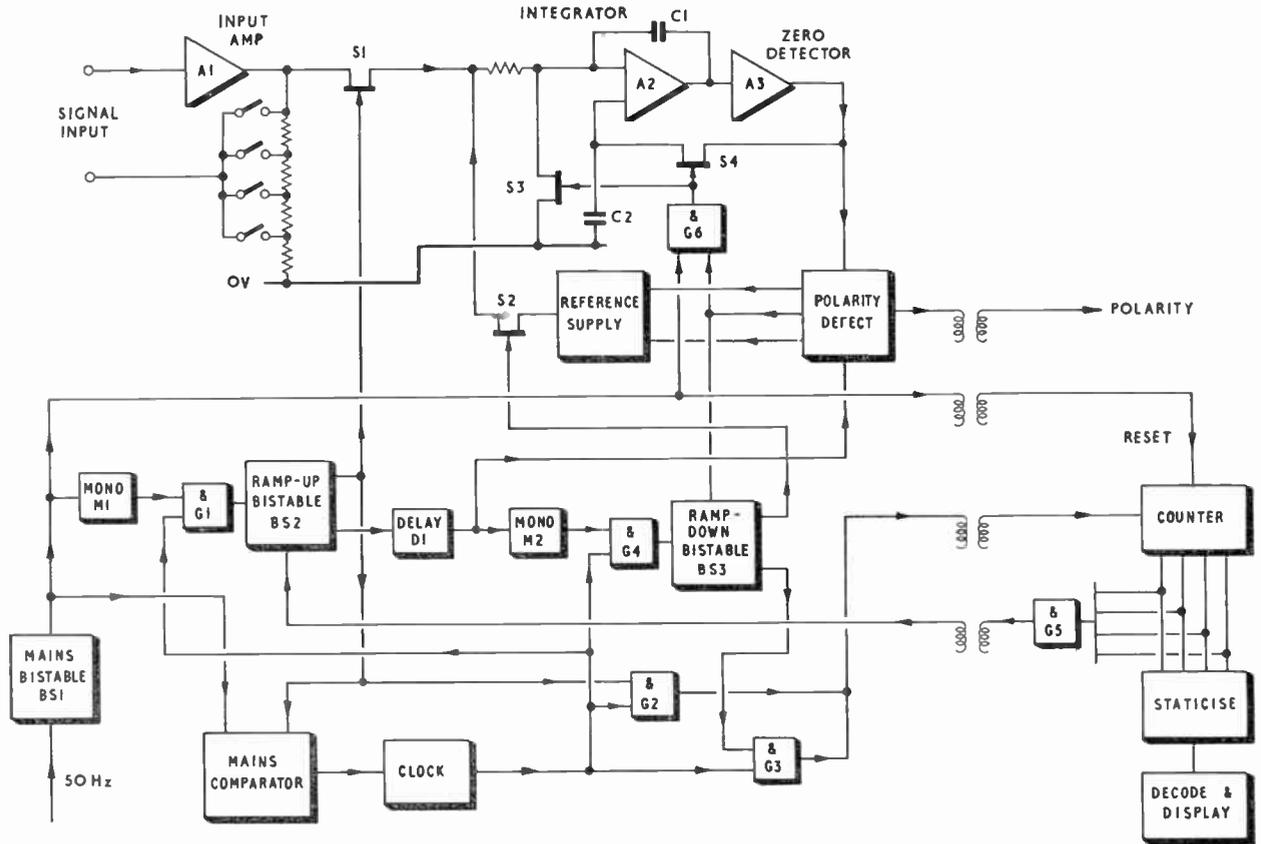


Fig. 3. Dual ramp voltmeter with circuitry to determine correct polarity after switching disturbances.

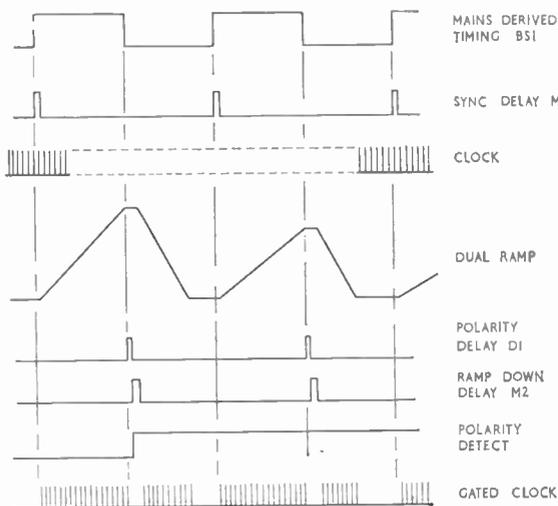


Fig. 4. Timing diagram for dual ramp voltmeter of Fig. 3.

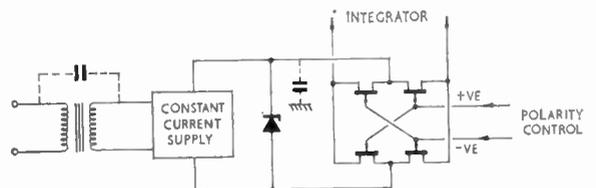


Fig. 5. Supply to Zener diode which determines polarity.

In the condition when the f.e.t. is being switched to the 'off' state the error is much larger. After the f.e.t.s drain source goes open circuit, the remainder of the gate switching voltage is applied to the input of the integrator via the gate-drain capacitance.

The resultant error voltage developed across C_1 (Fig. 2) will be directly proportional to the ratio of C_{GD} to C_1 .

Therefore

$$dV = V_g \frac{C_{GD}}{C_1}$$

The only time that this will in fact cause an error is when the input f.e.t. S1 (Fig. 1) is being switched off at the end of ramp-up.

Now taking an example, let

$$V_g = 20 \text{ V}$$

$$C_{GD} = 5 \text{ pF}$$

$$C_1 = 0.2 \text{ } \mu\text{F}$$

then

$$dV = \frac{20 \times 5 \times 10^{-12}}{0.2 \times 10} = 0.5 \text{ mV.}$$

From equation (13) the least significant bit size is 0.5 mV, therefore the error in example taken is 1 bit.

The actual gate voltage V_g , which is effective in causing this error, will depend upon the amplitude and polarity of the input voltage. Therefore both zero and calibration errors will be caused. As the error is only about one bit in magnitude and will be adjusted out during initial setting-up, the remaining error will depend on the variation of C_{GD} with time and temperature. In practice this has not been found to be a serious problem.

2.3.2. Transient causing wrong logic decisions to be made

These switching transients can cause another type of error, especially when the d.v.m. is measuring voltages near zero. At the end of the ramp-up period a polarity decision has to be taken to decide the polarity of reference voltage required to ramp the integrator back to zero volts. A logic decision is taken depending upon the output polarity of the zero detect amplifier. If this decision is taken at the same time as the input to the integrator is switched, then any disturbance at the input to the zero detect amplifier, especially when the integrator is near zero volts, may cause the zero detect amplifier to change its polarity and cause a wrong polarity decision to be made. If this happens then, depending upon the rest of the organization of the d.v.m., the ramp will continue in the same direction, looking for a zero that does not exist, and a completely erroneous reading will result.

The more usual case is that the zero detect logic for the end of count is switched in at the end of the ramp-up period, and a spurious pulse takes the zero detect amplifier through zero: then a zero count will be recorded, which will be in error. This can cause the instrument to have a zero which is two or three bits wide. These switching spikes can be transmitted to the zero detect amplifier via h.t. rails, common earths or direct radiation. This amplifier is very susceptible to transient pick-up because of its high sensitivity and frequency response. In this example the sensitivity must be such that for 0.5 mV input a full output swing is required and the slewing rate of the output must be in excess of 10 V/ μ s.

Because of this effect, it has been stated that the dual ramp voltmeter is not very accurate about zero volts. One method which eliminates some of these problems is to wait at the end of the ramp-up period until the switching disturbance is over, and then make the polarity decision. Figure 3 shows a complete block diagram of a voltmeter which is organized to do this. Figure 4 shows the timing diagram for this instrument.

It should be noted that there are two delays, one at the end of the ramp-up period before the polarity decision is made, and a further delay after the polarity decision has been taken before the ramp-down is started. Bistable BS3 (Fig. 3) controls the ramp-down period. This bistable holds the gate G3 closed during the wait period generated by M2 at the end of ramp up. Hence, no spurious signal from the zero detect amplifier A3 can cause the gate to be prematurely opened when either S1 is switched, or when the polarity decision is being taken.

Some methods of switching the reference for polarity decision can cause very large current transients which are particularly difficult to isolate. In the author's opinion, one of the best methods of reference switching results in large current transients. The method is to use one Zener diode and to switch either the positive or negative end of the Zener to the 0 V rail to change polarity. This means that the Zener must be fed from a separate isolated supply (Fig. 5). The stray capacitance of the isolating transformer (a few tens of pF) has to be switched through the full reference voltage amplitude in a relatively short time period.

Thus the method of waiting after the ramp-up period has two advantages:

- to ensure that the wrong polarity decision is not taken due to switching transients;
- to ensure that the polarity switching does not prematurely operate the zero detect amplifier and switch the ramp-down bistable.

There is still a further problem when the reference voltage is switched in after the second delay, M2, in

that it could cause the ramp-down bistable to be triggered prematurely. The only possible solution to this would be to introduce a further half-bit delay before the ramp-down bistable is allowed to be switched. In practice it has been found unnecessary to take this further precaution.

2.4 Digital Synchronization of the Clock to the Ramp Waveform

If there is no synchronization between the ramp and the clock then the initial timing period can vary by up to one count, and this will cause an error in reading of one part in N_1 , where N_1 is the full-scale count, e.g. for a 2×10^4 full scale voltmeter an error of 0.005% can occur. If the technique described in Section 2.3 of waiting for a short period at the end of ramp up is used, the ramp-down period will not be synchronized to the clock either and a further error of one part in N_1 will occur, thus giving a total maximum error of 0.01% for a 2×10^4 full scale instrument. Because of this, it is essential for optimum performance to synchronize the start of ramp-up and the start of ramp-down to the clock frequency.

In Fig. 3 (the full block diagram of the d.v.m.) the synchronization is carried out by gates G1 and G4. G1 controls the start of the ramp-up period, one of the G1 gate inputs is connected via a monostable timer to bistable BS1, the other gate input is connected to the clock. The ramp-up bistable will not be fired until the monostable has fired and a clock pulse is present, therefore the beginning of the ramp-up period is synchronized to the clock waveform.

A similar logic action takes place with gate G4, and BS3 to synchronize the start of the ramp-down period to the clock.

2.5. Temperature Stability of the Reference Voltage

The reference voltage is almost always nowadays generated by the use of a compensated Zener diode. There are three main causes of drift in this sort of reference:

- The temperature coefficient of the Zener.
- Variation in the constant current feeding the Zener diode.
- Random variation of Zener voltage due to the supply being switched off and on.

2.5.1. Temperature co-efficient of the Zener

It is always dangerous to specify the best performance possible with this type of component because of the very wide range of manufacturers and the continued improvements which are becoming available. However, the best temperature coefficient that is valid at present, at a reasonable economic price, is about ± 10 parts in $10^6 \text{ deg}^{-1}\text{C}$. If the overall

temperature coefficient for the instrument is to be 10 parts in $10^6 \text{ deg}^{-1}\text{C}$ the total coefficient will be taken up with this one item alone.

Once a decision has been taken to control the temperature of the Zener then it is not expensive, or difficult, to control its temperature to within $\pm 1 \text{ degC}$, over an ambient change from 0°C to 50°C . Therefore a 10 parts in $10^6 \text{ deg}^{-1}\text{C}$ Zener would contribute a temperature coefficient of 0.2 parts in $10^6 \text{ deg}^{-1}\text{C}$.

2.5.2. Variation in the constant current source

The output impedance of these types of compensated reference Zener diodes is approximately 15Ω , therefore any variation of the constant current will develop an error voltage across the output impedance of the Zener. If the errors from this are to be kept to within 1 part in $10^6 \text{ deg}^{-1}\text{C}$ and the voltage of the Zener is 10 V, then

$$1 \text{ part in } 10^6 \text{ deg}^{-1}\text{C} = 10 \mu\text{V deg}^{-1}\text{C}.$$

Therefore the maximum error current

$$= \frac{10 \times 10^{-6}}{15} = 0.66 \mu\text{A deg}^{-1}\text{C}$$

Let the constant current for the Zener be 7.5 mA, then the temperature coefficient of the constant current must be less than

$$\frac{0.66 \times 10^{-6}}{7.5 \times 10^{-3}} \times 10^6 = 88 \text{ parts in } 10^6 \text{ deg}^{-1}\text{C}.$$

This temperature coefficient is quite easily obtainable with a number of well-known constant current sources.

2.5.3. Variations due to supply switching

This is solely a malfunction due to the manufacturing process of the Zener diode. Some manufacturers have developed processes which yield Zeners which have very much better characteristics in this respect than others. The effect results in the Zener settling to a different voltage after its supply has been switched off and switched on again. This can result in a variation of up to 0.02%. If this type of Zener were used, it would result in the instrument having to be re-calibrated every time it was disconnected from the mains supply.

2.6. Temperature Coefficient and Stability of the Gain Determining Resistors in the Input Amplifier and Attenuator

The gain of the input amplifier is determined by the feedback resistors as shown in Fig. 6. Any change in the ratio of these resistance values will cause a calibration error. For instance if RL6 is closed, then the gain of the amplifier will be:

$$A = \frac{R_3 + R_4 + R_5 + R_6}{R_6} \quad \dots\dots(18)$$

Therefore the ratio accuracy of R_6 to R_3 , R_4 and R_5 must be maintained for high accuracy.

For the attenuated ranges, i.e. the ranges above the maximum voltage swing of the input amplifier, R_1 and R_2 are switched in, which will add to the overall temperature coefficient of the instrument.

The important parameter of these resistors is how well the change of resistance with temperature and time track with each other. For values up to 100 k Ω it is possible to use a special type of metal film resistor which has a ratio temperature coefficient of not worse than ± 2.5 parts in $10^6 \text{ deg}^{-1}\text{C}$ and can be selected for a temperature coefficient of ratio of ± 1 part in $10^6 \text{ deg}^{-1}\text{C}$. The long-term stability on light load is better than 20 parts in 10^6 per year.

These resistors can be used for the gain determining resistors of the amplifier, therefore the temperature coefficient of the unattenuated ranges attributed to the input amplifier will be ± 2.5 parts in $10^6 \text{ deg}^{-1}\text{C}$ plus the voltage drift of the amplifier itself. For the attenuated ranges it is usual to offer an input resistance of 10 M Ω . Therefore these resistors will have to be precision wire-wound types for the best performance possible at present. The best ratio temperature coefficient for precision wire-wound resistors of 10 M Ω and 100 k Ω is about ± 5 parts in $10^6 \text{ deg}^{-1}\text{C}$ and this will have to be added for the attenuated ranges.

2.7. Short-term Stability of the Clock Frequency

Equation (5) for the final count N_2 shows that the voltmeter is independent of the clock frequency. This is assuming that the clock frequency remains constant throughout the digitization cycle. However, during this cycle the frequency must be stable to better than the overall accuracy required of the d.v.m., in the example taken to better than 0.01%. This is not very difficult over the short time of digitization, but care has to be taken that the oscillator is relatively insensitive to supply rail change. Further problems can arise when the oscillator frequency is deliberately adjusted to keep the ramp-up period to exactly one period of the mains in order to optimize the series

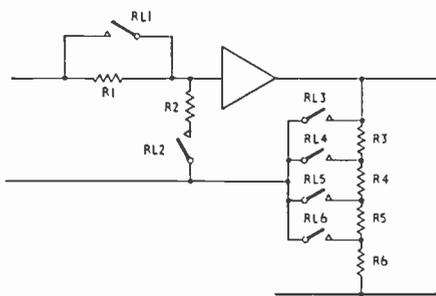


Fig. 6. Gain determining resistor network for input amplifier.

mode rejection of the instrument. This is dealt with in Section 2.9.

2.8. Linearity of the Ramp Generator

The linearity of the ramp generator is given by the formula

$$E = \frac{50T_1}{(1-A)RC} \left[(1-A) \frac{R}{R_L} + 1 \right] \quad \dots\dots(19)$$

where

E = worst case % error

T_1 = integration time

A = gain of the integration amplifier

R = integration resistor

C = integration capacitor

R_L = leakage resistance of the integrator capacitor.

From this it can be seen that the percentage error is directly proportional to the integration time. Therefore if the ramp-up time equals the ramp-down time the error will be equal and will cancel. At all other times the cancellation will be only partial. However, with a ramp of the speed required for this type of d.v.m. there is no real problem of achieving a linearity of considerably better than 0.01%.

If the constants in the example are applied to the formula, an amplifier gain of 80 dB and an R_L of $10^9 \Omega$ would theoretically provide a linearity of 0.001%.

2.9. Errors Caused by Series and Common Mode Signals

Plenty of papers have been written on the subject of series and common mode, therefore it is not intended in this paper to deal with the subject as such. The problem with common mode interference signals is that they develop a voltage across the unbalanced impedances in the signal lines and generate a series mode signal. The usual source of common and series interference signals is from the mains supply.

All types of integrating voltmeter have the property of offering, in theory, infinite rejection to the frequency whose period is equal to the integrating period. If the frequency, or the integrating period, alters by only 1%, then this constitutes a 1% series mode signal, which is not integrated out and the rejection will fall from infinity to 40 dB.

It can be seen that if the integration period is made always to equal one mains period, or a multiple of mains periods, in theory the instrument will always offer infinite rejection to mains series mode interference. The integration period is determined by the 'full scale' count of the display counter, therefore by adjusting the clock frequency the integration time

will be altered. Figure 7 shows a block diagram of a system for controlling the integration time. First the incoming mains frequency is squared and divided by two, to provide a square wave with a half-period equal to the mains period. This square wave is compared to a period which is generated by the full-scale count of the counter. If the counter period is either shorter or longer than the mains period, then the comparator will generate a logic signal to command the staircase voltage generator to alter one step, either positive or negative, dependent upon the direction of the error. The output of the staircase voltage generator provides the controlling voltage for the voltage controlled oscillator. This completes the servo loop.

It is possible for noise on the mains temporarily to alter the mark/space ratio of the mains-derived square wave. This will cause the digital comparator to generate a false signal. With the technique described, the maximum change in frequency for any one comparison is a fixed amount, usually about 0.01% per sample, therefore a false signal will alter the integration period by only a small percentage. If a technique is used where the frequency change is proportional to the error signal, then a false signal could alter the mains period by a large amount, with subsequent degrading of the series mode rejection for the next few readings.

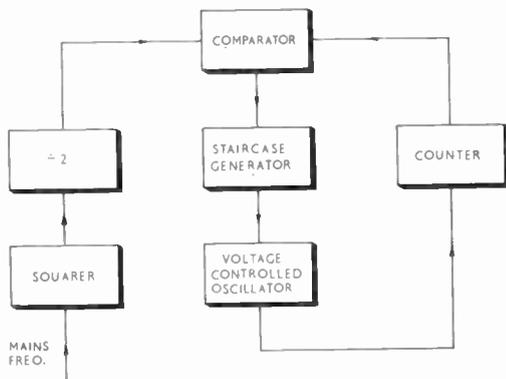


Fig. 7. Block diagram of system for controlling integration time in presence of mains interference.

Figure 7 is a simplified diagram which illustrates the technique. In practice the complete circuit is slightly more complex, the digital comparison is done at the end of the ramp-up, but the information is stored until the end of ramp-down, before the correction is applied.

This method of control has the advantage that no correction to the clock frequency is taking place during the integration period, therefore this technique will not introduce any error by causing the clock frequency to shift during a digitization cycle.

3. Conclusions

It is not possible to consider all of the factors which can cause errors in the dual ramp technique, but all the major items have been discussed. There is no one area which affects the overall accuracy; but a multiplicity of small items which, if attention to their detailed design is not given, can add up to spoil the overall specification. The switching of the input signal and the reference into the integrator has caused many problems in the past, and it is only recent improvements in semiconductor switch technology which have made it a practical proposition to implement the patent filed by R. W. Gilbert in 1957 and now held by Schlumberger.⁵

The input amplifier has been dealt with only briefly in this paper. The performance of all d.v.m.s is, regardless of the technique employed, very dependent upon the performance of its input amplifier.

A dual ramp voltmeter which has been designed using the techniques described, has proved that it is possible to meet the very exacting specification requirements laid down at the beginning of Section 2.

All of the first 100 instruments were tested for temperature coefficients of calibration over the temperature range 0–50°C. An average figure of 4 parts in $10^6 \text{ deg}^{-1}\text{C}$ was obtained from the batch on the unattenuated ranges with a worst case figure of 8 parts in $10^6 \text{ deg}^{-1}\text{C}$ and a best case figure of 2 parts in $10^6 \text{ deg}^{-1}\text{C}$.

The voltage zero drift averaged at $0.1 \mu\text{V deg}^{-1}\text{C}$, mainly due to the input amplifier. The linearity of the whole batch was better than 0.005% of reading.

4. Acknowledgement

The author thanks the Directors of the Solartron Electronic Group for giving permission to publish this paper.

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Of Current Interest

European Space Research—A Progress Report on Four Satellites

Iris, the first of the four satellites developed and launched by the European Space Research Organization completed its 10,000th orbit round the Earth on 25th March last. The satellite, which was launched on 17th May 1968,† was originally expected to have an operational life of six months and has now comfortably exceeded the two-year mark. The on-board instruments continue to function very satisfactorily, and the ground stations are successfully receiving their output data on solar electromagnetic and corpuscular radiation, thus providing much more extensive research material than was anticipated.

The satellite's remarkably long life is not only evidence of the efficiency of European co-operation in the space field, but is also providing scientific research with an unexpected bonus. Since the effort—in terms of money and manpower—devoted to the development and launching of a satellite is determined on the basis of the quantity and quality of the data required from it during its predicted life-time, *Iris*, by substantially exceeding expectations, is giving exceptionally good value.

Aurorae, the Organization's second satellite, launched a few months after *Iris*, on 3rd October 1968,‡ functioned satisfactorily until it re-entered the Earth's atmosphere on 26th June last. It transmitted highly valuable data on auroral phenomena to ground stations and long outlasted the design goal of six months' lifetime. Early in March, *Aurorae* completed its 7500th orbit, having executed more than 31 500 commands and transmitted more than 41 500 minutes of high-speed real-time data.

A special feature of *Aurorae* was the high-speed data transmission, which augmented the normal continuous telemetry. The high-speed information system was commanded into operation over the northern auroral zone, and enabled measurements to be made so rapidly that the fine structure of auroral phenomena could be surveyed with a precision of a few hundred metres.

HEOS-1, ESRO's third satellite was launched on 5th December 1968 into a highly eccentric orbit, with apogee at 225 000 km (i.e. almost two-thirds of the distance to the Moon) and a period of about 4½ days. Its scientific mission is the exploration of interplanetary space far beyond the vicinity of the Earth, with particular emphasis on the study of magnetic fields, cosmic radiation and the solar wind outside the magnetosphere and it is still fully operational. The data transmitted by this satellite—which can properly be called an interplanetary space observatory—are received almost continuously by the ground stations owing to the satellite's very great distance from the Earth during the most interesting part of its trajectory.

As regards *Boreas*—ESRO's fourth satellite, launched on 1st October 1969—its mission was to study auroral phenomena over the northern polar zone. The selected

orbit was a low one, inevitably leading to a relatively short lifetime. In the event, *Boreas* completed 835 orbits, re-entering the atmosphere after two months of intensive and successful operations. It had achieved its mission of extending and complementing the investigations started by *Aurorae*.

The fact that these four satellites—designed, developed and launched by ESRO—have successfully achieved their objectives, despite initial difficulties, proves that European co-operation can be worthwhile and efficient. On the basis of the experience gained during these first years, Europe is now envisaging a more extensive programme. This includes not only several scientific satellites, some of them far more ambitious than the four mentioned, in development or preparation, but also the development of application satellites, the need for which has become urgent in fields such as communications, meteorology and air traffic control.

Employment of Graduates

Data on the employment of graduates are given in the Statistical Supplement to the latest Annual Report of the University of London Appointments Board. A full breakdown of what graduates do is not possible owing to the sizeable number who do not obtain posts at once, but some interesting facts emerge from a review of the information presented.

In 1969, 275 students graduated in electrical engineering (which includes electronic engineering), the largest number in any technology. Only about 10% remained at the University to continue postgraduate study, which compares with nearly 40% of those obtaining degrees in physics and applied physics. Over 120 electrical engineering graduates were employed in industry and 30 in public service, while the field of employment of 55 was unknown or undecided at the end of 1969: it is difficult to understand why this figure should be a considerably higher proportion of those graduating than for other fields of technology (except chemical engineering, which is about the same).

In relation to the numbers of first degrees obtained at London, the number of higher degrees in electrical engineering and electronics—111 in 1969, is surprisingly large, even bearing in mind that the numbers graduating one or two years ago were larger. The proportion of overseas students for higher degrees—about 25%—is about four times as high as it is for the first degree and the constituent colleges of London University also obviously attract a substantial number of graduates from other British Universities. Predictably, a larger percentage of holders of higher degrees go to universities and technical colleges compared to industry than do first degree holders.

Few deductions can be drawn from the London University figures on the relative attractiveness to women of electrical engineering as compared to other branches of engineering—only two graduated in electrical engineering (both of whom have incidentally gone into industry) and only metallurgy recorded a greater figure, namely four. There were no higher degrees gained by women in 1969 in electrical engineering and only one in each of mechanical, civil and chemical engineering.

† 'The *Iris* satellite,' *The Radio and Electronic Engineer*, 33, No. 6, p. 342, June, 1968.

‡ 'British experiments in ESRO I (*Aurorae*) satellite' *The Radio and Electronic Engineer*, 36, No. 5, p. 272, November 1968.

Contributors to the Journal



A. R. Pratt graduated in 1964 from Birmingham University with a B.Sc. degree in electronic engineering and continued at the University until 1967 on post-graduate work concerned with radar signal processing. In 1967 he was appointed to be an assistant lecturer in electronic engineering at Loughborough University of Technology and was promoted to a lectureship a year later. Mr.

Pratt's main research interests include time statistics of non-stationary noise processes and radar and sonar signal processing.



G. G. Scarrott is Manager of Research and Advanced Development Organization of International Computers in Stevenage. Following graduation, he worked for several years at the Cavendish Laboratory, Cambridge, on nuclear physics instrumentation. In 1953 he joined the Computer Department of Ferranti Limited which was later incorporated into the International Computers and

Tabulators Limited and is now part of ICL. Mr. Scarrott has published papers on electronics for nuclear physics, including a pulse height analyser, and, since moving into the computer field, on wire-type acoustic delay lines, slave stores, efficient use of logical devices† and principles for system design.



D. Wheable joined the Post Office Research Station at Dollis Hill in 1943 and studied for H.N.C. and endorsements at Enfield Technical College. In 1957 he moved to Blackburn Electronics and became chief design engineer; later he was transferred to Hawker Siddeley Dynamics as chief circuit engineer in the Auto Check Out Division. Four years ago he joined Solartron

with digital instruments. While as group leader concerned at Blackburn Electronics, Mr. Wheable was joint author of a paper read before a meeting of the Institution in 1960 and subsequently published in the *Journal*, which described one of the earliest electronic digital voltmeters.†



Dr. J. A. Coekin graduated with a B.Sc. degree in electrical engineering from the University of Nottingham in 1956 and in 1960 he received his doctorate, also in electrical engineering, from the same University. From 1959 to 1962 he worked with Pye Limited, Cambridge, as applications engineer on digital systems. For the next three years he held a lectureship in electrical

engineering at the Ahmadu Bello University, Northern Nigeria, where he carried out research work on atmospheric electricity and tropical lightning. On returning to England in 1965, he was appointed a lecturer in the Department of Electronics at the University of Southampton and in March of this year, he received his present post of Senior Lecturer in Electrical Engineering at James Cook University, Townsville, Queensland, Australia. Here he is continuing his research interests of information displays and ultra high-speed pulse techniques.



M. L. Meade received the degree of B.Sc.Tech. in electrical engineering from Manchester University in 1964 and subsequently joined the Radiation Laboratory of E.M.I. Electronics Limited to work on the design of microwave modulators and aerial arrays. Since 1967 he has been engaged as a Research Assistant at the J. J. Thomson Laboratory, University of Reading, where

his research interests include the investigation of the noise properties of semiconductor devices and the design of active filters and signal recovery systems.

A biographical note on Dr. E. A. Faulkner was published in the July 1969 issue of the *Journal*.

† Fuchs, H. and Wheable, D., 'An all-electronic four digit digital voltmeter', *J. Brit. Instn. Radio Engrs.*, 20, No. 7, p. 541, July 1960.

‡ Scarrot, G. G. and Mitchell, R. W., 'High-speed logic circuits using a tunnel diode transistor feedback amplifier', *The Radio and Electronic Engineer*, 26, No. 6, p. 485, December 1963.

The University of Birmingham 6-metre Offset Cassegrain Aerial

A strong programme of research in the fields of microwave communications, design and control of reflector aerials and radar meteorology has been built up in the Department of Electronic and Electrical Engineering at the University of Birmingham over the last eight years. Projects have included microwave communication by Moon reflection, probing of the atmosphere by radar tracking of acoustic shock waves, investigations of a variety of aerial configurations for satellite communication and radio astronomy, and strategies for computer control of aerial positioning. The work has been supported in part by grants from the Science Research Council.

In 1966 it became clear that a fully steerable aerial as an adjunct to the new Electronic and Electrical Engineering Building then being planned would provide a flexible experimental facility which could be used in a number of projects of this nature and would stimulate new work. It was particularly desired to extend the research to the shorter wavelengths between 3 cm and 7 mm since it was evident that a combination of demand and technological advance was likely to lead to their useful application in the future. In the event the pace of development in this waveband seems likely to be even faster than anticipated.

Requirements of Aerial

A specification was drawn up for the aerial to achieve full performance at a wavelength of 1 cm. The reflector profile was to be within $1/30$ of a wavelength (r.m.s. error) of the true parabola, corresponding to 0.3 mm (0.01 in) over the 6 m (20 ft) diameter reflector. The position control was to be accurate to a small fraction of the beamwidth, which for this aperture and wavelength was 12 minutes of arc. Such an accuracy required digital control and the concept of an aerial mounted directly on the new building and integrated with an on-line control computer was adopted. This would also permit real-time processing experimental radio observations.

The roof-top location dictated a design of aerial with a very low pick-up of man-made and natural noise below its horizon. The short wavelength of operation implied high loss of power in waveguides from transmitter or receiver to the feed horn of the aerial, so that a design of aerial which would make the length of waveguide as short as possible was sought. Also for flexibility of experimental arrangements, a roomy cabin immediately adjacent to the waveguide access point was required. Last, but important, an economic design was dictated by the modest funds available.

Electrical Design

Together with the consultants, Husband and Company, (designers and consultants for the Jodrell Bank radio telescopes and the aerials at Goonhilly and other satellite Earth stations) the configuration selected to meet the above requirements was a reflector with an offset Cassegrain feed as shown in Fig. 1. The cabin can be rotated about the vertical axis to obtain azimuthal motion, while rotation of the reflector, support arm and small sub-

reflector about the slant axis rotates the beam around a cone centred on the slant axis. In Fig. 1 the beam is shown pointing vertically, while rotation of 180° about the slant axis brings the beam round to point slightly below the horizontal and to the right.

This geometry gives an exceptionally short run of waveguide to the feed-horn with no awkward tilting of this feed as the beam elevation is changed, making it possible to provide a roomy cabin with direct access to the feed waveguide and with no motion other than rotation.

The offset Cassegrain sub-reflector adopted is clear of the beam, thus avoiding blocking and scattering of radiation and achieving the characteristic of low-noise pick-up from beneath the horizon required.

For lower-frequency operation a front feed is preferable and the Cassegrain sub-reflector can then be swung out of the way to leave the prime focus available.

Research Programme

The aerial will be used to extend the programme of microwave Moon-reflexion communication and to make use of Moon-reflected signals to measure the absorption of

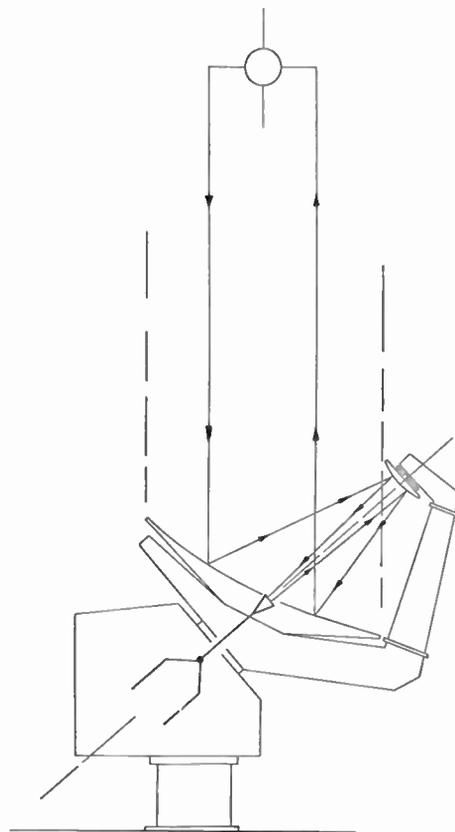


Fig. 1. Diagram of the Cassegrain configuration of the Birmingham University aerial.

radio waves in the atmosphere, particularly during heavy rain. This information is needed for both satellite and terrestrial radio links.

Radiometers for 3 cm and 1 cm wavelength will be used to map noise contributions from rain, cloud, terrain and man-made sources, so building up data which will show whether an urban satellite communication terminal is practical.

Measurements of satellite signals at 3 to 1 cm wavelength and the effects of weather upon them will be made in conjunction with the British Post Office, the European Space Research Organization (ESRO) and the Communication Satellite Corporation (COMSAT). Radio echoes from buildings will be studied in so far as they affect the radio telephone and television links between the Post Office towers in cities such as Birmingham and London.

The use of electronic techniques to compensate for mechanical imperfections in aerial reflectors, already under study, will be continued with the aim of extending the use of the 6 m reflector to even shorter wavelengths.

New control strategies using the real-time digital computer control will be employed to track communication satellites. The aim is to eliminate the present costly split-feed and twin-channel receivers needed for this and to employ instead the normal communication feed horn and receiver in conjunction with a 'hill-climbing' optimizing control programme.

The experiments on the positional control of the aerial and new tracking techniques will be under the direction of Professor H. A. Prime, Head of the Postgraduate School and the Industrial Division of the Department of Electronic and Electrical Engineering, and the radio experiments will be under the direction of Professor E. D. R. Shearman, Head of the Communications and Electronics Division. Before taking up his University post, Professor Prime had been responsible in industry for the design of the electrical control systems of the Jodrell Bank 250 ft and 85 ft radio telescopes and of the first Post Office Goonhilly satellite aerial.

Professor Shearman was formerly at the S.R.C. Radio and Space Research Station where he designed rocket and satellite instrumentation and carried out radio propagation research. He is an honorary member of the staff of the Royal Radar Establishment at Malvern, where he is planning with its radar meteorology group cooperative experiments with the new aerial.

This aerial is the first of its kind in Britain and, apart from a scale model, in the world. The Post Office is to have a similar one constructed for experimental use at its new research station at Martlesham in Essex. Both aeriels are being constructed by Markham & Company Ltd., of Chesterfield, with Portobello Fabrications Ltd., of Sheffield, as subcontractors for the main reflector bowl, aluminium support arm, ladders and platform.

Letters

Interconnexion of F.D.M. and T.D.M.

I have read with considerable interest the paper by Professor J. E. Flood and Mr. A. C. Ponder.† A statement on page 72 (last paragraph) leaves me somewhat puzzled however. For full flexibility between f.d.m. and t.d.m. systems, the highest carrier frequency is quoted as $2nf_r$. Should it not be n^2f_r , i.e. the n th harmonic of the lowest frequency nf_r ? I would be most grateful if the authors would kindly elucidate this point.

46 London Road,
Salisbury, Wiltshire.

R. J. BEST

12th May 1970

The analysis given in Section 8 of our paper deals with an ideal case. In order to transfer the amplitude modulation from a sinusoidal carrier to a pulse train, the latter should sample the former at its peaks (as shown in Fig. 8 of the paper). For an n -channel system, a peak of the wave of carrier frequency f_c should coincide with each channel

† 'Interconnexion of frequency-division and time-division multiplex transmission systems', *The Radio and Electronic Engineer*, 39, No. 2, pp. 67-76, February 1970.

pulse in the frame period ($1/f_r$) of the t.d.m. system. Thus, the minimum carrier frequency is nf_r . This condition will also be satisfied for $f_c = 2nf_r, 3nf_r, \dots, n^2f_r$. The highest carrier frequency would then be n^2f_r , as stated by Mr. Best. However, this would make very uneconomic use of bandwidth because the separation of the carriers is greater than necessary.

In practice, the received carriers will not be in phase at the start of the t.d.m. frame because of delay distortion in the transmission path. A means of adjusting carrier phases is required and this can also be used to align carrier peaks with pulse positions. If the carrier frequencies lie between nf_r and $2nf_r$, the adjustment required does not exceed 45° . The statement in Section 5 that all carriers must have peaks at every sampling time is then not strictly true, since some carriers are adjusted in phase to coincide with some sampling pulses.

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14th July 1970

Speech and Vocoders

By

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C.Eng., M.I.E.R.E.†

Presented at a meeting of the Institution's Communications Group in London on 8th January 1969 and at a meeting of the East Anglian Section in Hornchurch on 29th October 1969.

Speech signals are produced by relatively slow articulatory movements. This suggests that the information rate of the speech signal is much less than would be expected by considering the bandwidth of the acoustic signal. Vocoders attempt to exploit the redundancy in the speech waveform by extracting and transmitting the information bearing parameters of the speech signal. At the receiver, these parameters are used to control a speech synthesizer that reproduces the original signal without any serious loss of intelligibility but with some degradation of quality.

The paper describes speech production principles, and their application to speech synthesis; the operation of various types of vocoder and the problems of pitch extraction.

1. Introduction

Speech can be considered from many viewpoints: phoneticians, physiologists, and linguists would all consider the speech process in different ways. From the point of view of the communications engineer speech is an analogue signal whose bandwidth extends from the very low frequencies in the audio range (about 50 Hz) up to frequencies of the order of 8–10 kHz. Direct transmission of high quality speech therefore requires a bandwidth of this order. Fairly good quality speech can still however be transmitted with a bandwidth of approximately half that quoted because there are sufficient clues to perception in the lower half of the frequency band.

If the analogue speech signal is converted into digital form, for example by pulse code modulation, then an information rate of between 20,000 and 100,000 bits/second is needed for transmission depending on the subjective quality required for the received signal.

A speech signal is produced by relatively slow articulatory movements, and these movements can be described with sufficient accuracy by signals of much lower information rate than might be expected by considering the bandwidth of the acoustic signal. This implies that much of the waveform detail of speech is in fact redundant, and that a speech transmission system using some sort of model of the human speech production process at the receiver might require less transmission channel capacity than is needed for the speech waveform. Such systems are known as analysis-synthesis telephony systems or vocoders. Vocoders (the word comes from voice and coders) exploit the redundancy in the speech waveform. This is done by analysing the speech waveform and attempting to extract information-bearing parameters. These parameters can then be transmitted with a much lower bandwidth, or can be digitally encoded at a much lower bit rate, than the original signal. At the receiver,

† Joint Speech Research Unit, Post Office, Ruislip, Middlesex.

speech is then synthesized from the transmitted parameters, usually with some degradation of quality, but without any serious loss of intelligibility.

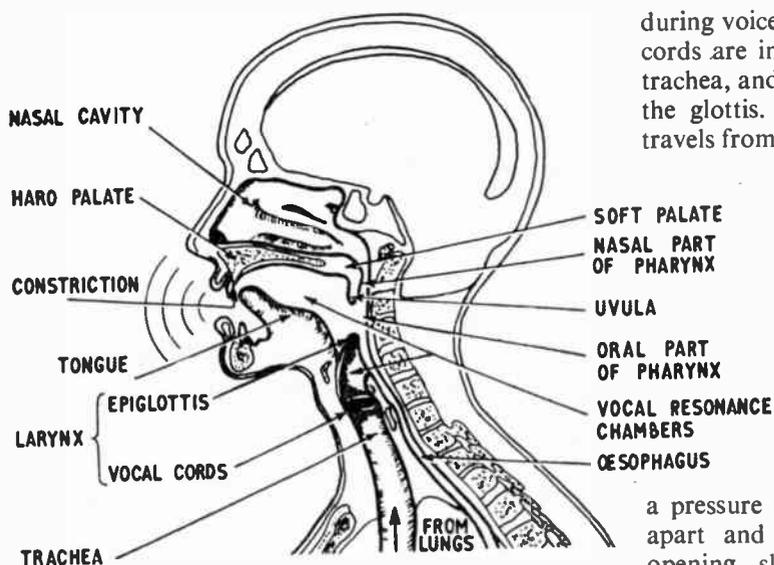
No practical vocoder has yet been developed which is sufficiently tolerant of a wide-range of speech material, and which gives a sufficiently high quality speech output signal, to be acceptable for commercial telephone use. The vocoder can still however provide a useful service in certain situations, for example on h.f. radio to provide narrow-band digital transmission with its immunity to noise, for space communications where the signalling rate must be kept low, and for military communications to provide speech privacy.¹ Another recent use for a suitably modified vocoder is to correct the change of quality of divers' speech caused by breathing a helium-oxygen mixture (the 'Donald Duck' effect).

The purpose of this paper is to explain the principles of speech production and vocoder operation, to discuss various types of vocoder and to illustrate the types of speech quality that can be obtained from vocoders.

2. Speech Production

Figure 1 shows in diagrammatic form the essential parts of the human vocal system. The vocal organs are the lungs, the trachea or windpipe, the larynx, the pharynx or throat, the nose, and the mouth. The part of this 'tube' extending from the larynx to the lips is known as the vocal tract, and in an adult male is about 17 cm long. The shape of this tract is varied extensively during speech production, by moving the lips, the tongue, and the jaw, i.e. the articulatory organs.

A typical speech waveform is shown in Fig. 2. This was produced from a recording made by a phonetician speaking in an anechoic chamber. The waveform was produced when the vowel 'ʌ' in the word comfort was being said. Generally, we can say three things



during voiced sounds they are drawn together. These cords are in fact folds of ligament at the top of the trachea, and the slit-like orifice between them is called the glottis. When we produce a voiced sound air travels from the lungs up to the trachea, and builds up

Fig. 1. Schematic diagram of the human vocal system.

about the waveform that are fairly obvious:

- (a) It looks a fairly complicated function of time.
- (b) Its form is consistent with its having been produced by some sort of resonant system.
- (c) It exhibits a periodicity.

Having examined a typical vowel waveform it is of interest to consider how some of the sounds of speech are produced.

Speech is produced by two basic types of sound source; the class of sounds known as voiced sounds are produced by puffs of air released during vocal cord vibration. Voiced sounds include all the vowels and such consonants as m, n, l and r. The other class of sounds, unvoiced sounds, are produced when turbulence is caused by air being forced through a narrow constriction somewhere in the vocal tract; examples of unvoiced sounds are f, s, p and t.

Some sounds, such as z and v, require the use of both types of sound source simultaneously.

2.1. Voiced Sounds

During ordinary breathing the vocal cords are in a relaxed condition and are held fairly wide apart, but

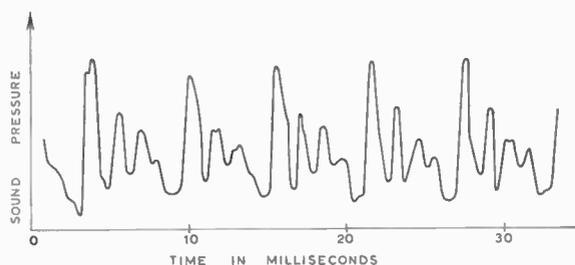


Fig. 2. The speech waveform corresponding to an occurrence of the vowel 'a' in the word k_Λmfət.

a pressure behind the vocal cords; these are pushed apart and air rushes through the narrow glottal opening, slowing down again when it reaches the wider pharynx above. By a combination of muscular tension in the cords and the lowering of pressure in the glottis due to the Bernoulli effect, the vocal cords are drawn back to their starting position and the air flow ceases. The sub-glottal pressure then forces the cords apart again and the whole cycle is repeated. It can be seen therefore that the vocal cords act as an intermittent barrier to the flow of air from the lungs, and in fact chop the air stream so that a discrete set of puffs is produced.

The vocal cord vibration period is a function of vocal cord mass, tension, and sub-glottal pressure. For normal male talkers these puffs of air are produced with a frequency typically in the range 50–250 Hz and extends to 500 Hz and higher for women and children. Although the frequency of vocal cord vibration is fairly high, its rate can be changed only slowly, by varying the sub-glottal pressure and tension of the vocal cords, both of which are under muscular control. These puffs of air constitute the basic generator for the voiced sounds of speech; typically the shape of a glottal puff (i.e. the volume velocity of air plotted against time) is approximately triangular, and since these puffs are quasi-periodic can be considered to have an approximation to a line spectrum.

Figure 3 shows the shape of some typical glottal puffs, and Fig. 4 the spectral envelope of a single puff. The repetition rate of the larynx puffs is closely related to the perceived pitch of the voice, and for this reason is known colloquially to vocoder engineers as the pitch frequency, or simply the pitch. In most electronic speech synthesizers electrical pulses are used to approximate the larynx puffs.

If we were able to listen directly to the larynx output it would sound rather buzzy as might be expected from the waveform shape. Normally how-

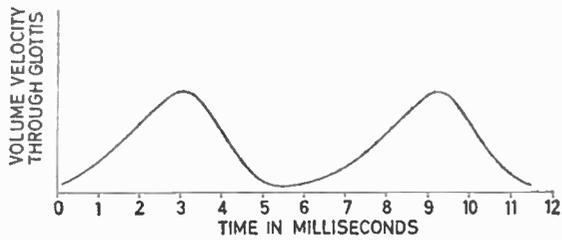


Fig. 3. The larynx waveform of two vocal periods corresponding to the speech waveform in Fig. 2.

ever the larynx output signal reaches our ears via the vocal tract, and it is the shape of the vocal tract that determines the speech sound quality we hear. The vocal tract can be shown to be quite closely analogous (up to fairly high audio frequencies) to a mismatched non-uniform transmission line, and for radio engineers is probably most easily considered from this viewpoint; it is therefore a resonant system that intensifies the energy of certain bands of frequencies. These resonances, whose frequencies can be changed by movement of the articulators, are given the name 'formants'. The formants superimpose their response on the vocal cord signal to produce the voiced sounds of speech. Voiced sounds are usually characterized by three or four formants in the frequency range up to 4000 Hz. For convenience the formants are numbered, the lowest one (F_1) corresponding to the lowest frequency, etc.

Figure 5 shows the spectrum of an occurrence of the vowel 'ʌ'. This clearly shows the line structure and the formants.

2.2. Unvoiced Sounds

Many of the sounds of speech come into the class of sounds known as unvoiced. During unvoiced sounds the vocal cords are held wide apart and the air stream from the lungs is forced through a constriction between the tongue and the teeth as in 's' or between the teeth and the lips as in 'f' causing

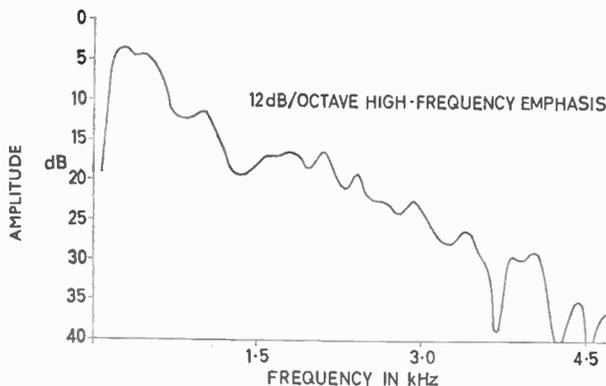


Fig. 4. Calculated amplitude spectrum for one of the vocal periods in the previous figure.

turbulence and producing the characteristic 'fricative' sounds of speech. The basic generator for this type of sound is the air stream whose source can be considered to be at the point of constriction. The unvoiced sounds do not exhibit the harmonic structure of the voiced sounds and the sound generator is probably best thought of in electrical terms as a random noise source. The energy of the fricative sounds is generally much lower than that of voiced sounds and the resonances in the system have greater bandwidth than for the voiced sounds.

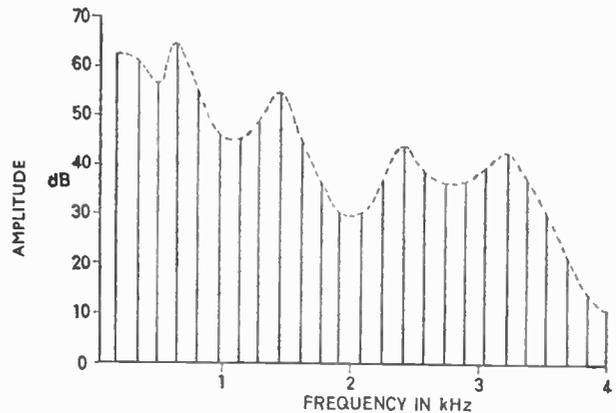


Fig. 5. Spectrum of an occurrence of the vowel 'ʌ' in the word kʌmfət.

2.3. Plosive Consonants

Another class of sounds that are produced fairly often in speech are the plosives or stop consonants; these sounds are produced by stopping the flow of air from the lungs by blocking the vocal tract at some point and then very quickly releasing the air pressure. The plosives therefore are always characterized by a silence preceding the burst of energy; they may be voiced, for example as in b and d, or unvoiced such as p and t.

The preceding description of speech production is not intended in any way to be exhaustive but it is hoped that a reasonable idea of the mechanism has been conveyed.

3. Speech Synthesis

Having considered how the various speech sounds are produced it is now possible to see how this knowledge can be applied to the synthesis of speech.

Most of the earliest speech synthesizers were mechanical, and in fact constructed as long ago as the 18th century. Alexander Bell in the late 1800s made a speech synthesizer by making a cast of a human skull and moulding the vocal tract and cords from rubber and similar materials. Mechanical models however are very difficult to control and

progress has only been made with the advent of electronic speech synthesizers.

Speech production can be thought of as a combination of two functions; firstly the generation of a sound source and secondly the modification of the sound from this source by the vocal tract. These two functions are given names taken from general circuit theory, and are known as the excitation and system functions respectively. Most speech synthesizers depend on this idea of the separation of the speech signal into an excitation function and a system function. The excitation function can be represented in

electrical terms as a generator producing a periodic waveform rich in harmonics during voiced sounds, and a random noise source during unvoiced sounds. Many synthesizers do not provide the mixed excitation signal required for the voiced fricative sounds such as z and v. In practice it is found that there are sufficient other perceptual clues for these sounds to be correctly identified in spite of this restriction. The system function (corresponding to the vocal tract) can be represented by a linear, time-varying four-terminal network terminated in a resistance representing the radiation resistance of the mouth. Control parameters

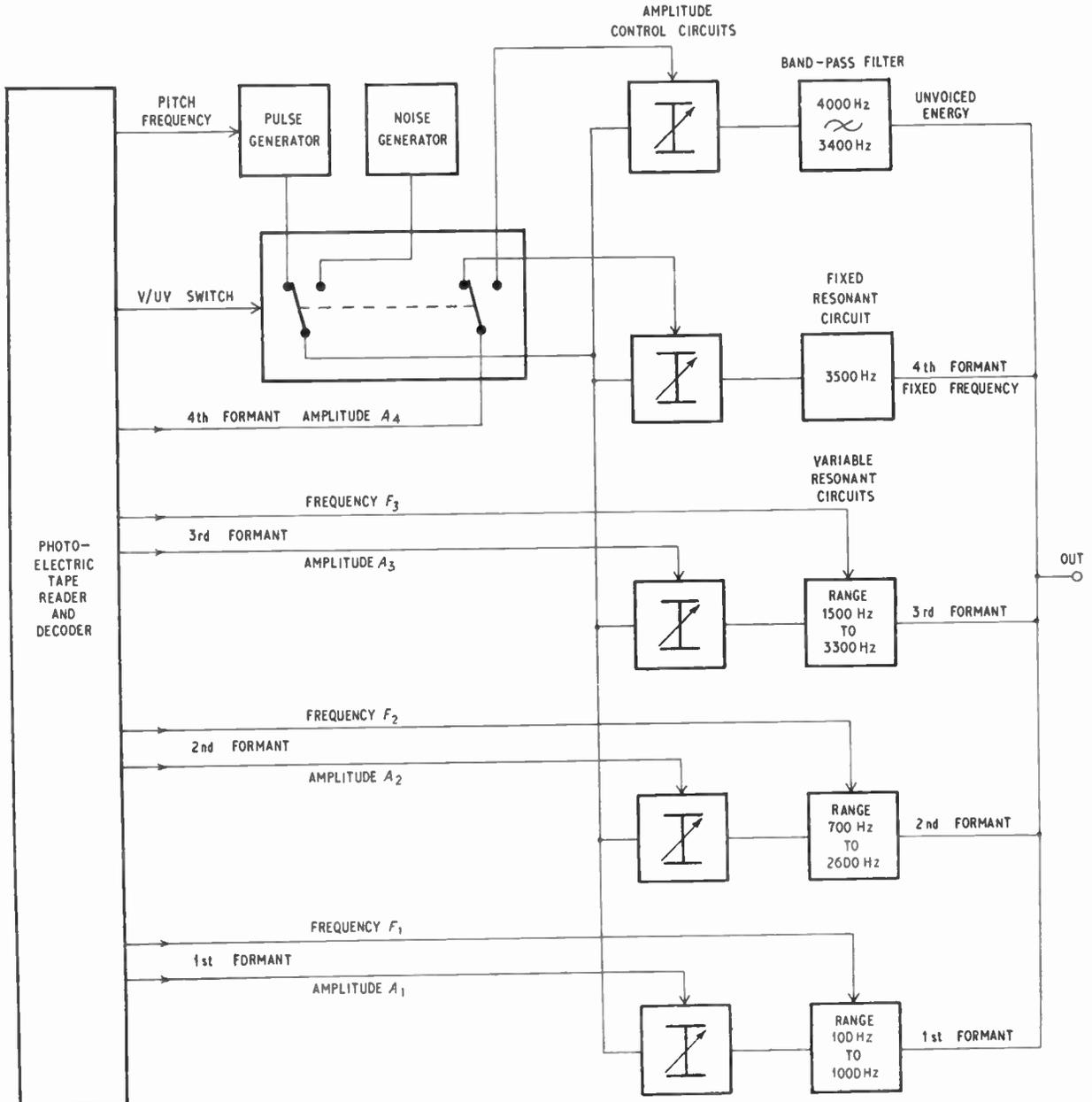


Fig. 6. Block diagram of a formant synthesizer. This synthesizer is normally controlled from a punched paper tape generated by a digital computer.

are used to vary the frequency response or spectral envelope of the four-terminal network to produce synthetic speech. Speech synthesizers of this type have been known for some time and in some instances the network corresponding to the vocal tract has been a lumped approximation to an electrical transmission line with parameters corresponding to the dimensions of the vocal tract that can be varied. Other types of synthesizer have been simpler in that they approximate the response of the vocal tract with simple resonant circuits. These are usually given the name formant synthesizers. A block diagram of a typical formant synthesizer is shown in Fig. 6.

4. Formant Synthesizer

Formant synthesizers are of two basic types. Those in which the resonant circuits are cascaded are called serial synthesizers, and those whose outputs are combined in parallel are known as parallel synthesizers. The particular synthesizer shown in Fig. 6 was designed to operate from a punched paper tape generated by a digital computer. It is a parallel-type formant synthesizer and single tuned circuits are used to generate the formant peaks. The centre frequencies of three of the tuned circuits corresponding to F_1 , F_2 , and F_3 are dynamically controlled by voltage analogues generated by the computer. The bandwidths of the tuned circuits are fixed and are set to a bandwidth of between 60 and 120 Hz. Analogue signals representing the amplitude of the formants are fed to the amplitude control circuits and at the same time a common excitation signal (pulses or random noise) is fed to all controllers. The output signal from the amplitude control circuits is therefore a wide-band (4000 Hz) excitation signal whose amplitude has been determined by the appropriate analogue voltage A_1 , A_2 , or A_3 . These signals are fed to the required voltage controlled tuned circuit. These variable frequency resonators select the band of the excitation signal in the region of their centre frequency and their outputs are simply added to provide the synthetic speech. Two other circuits are added as a refinement to this basic synthesizer; one is a fixed resonant circuit corresponding to a fourth formant, and the second a further wide-band filter that is switched on only during unvoiced sounds to provide a further improvement in quality. The excitation signal is provided either from a 'voltage-controlled' pulse generator during voiced sounds or from a random noise source during unvoiced sounds.

The speech quality obtainable from a synthesizer of this type can be very good indeed if great care is taken to provide the correct control parameters.

5. Vocoders

A vocoder is a device that performs measurements on speech that are in some way related to the short-

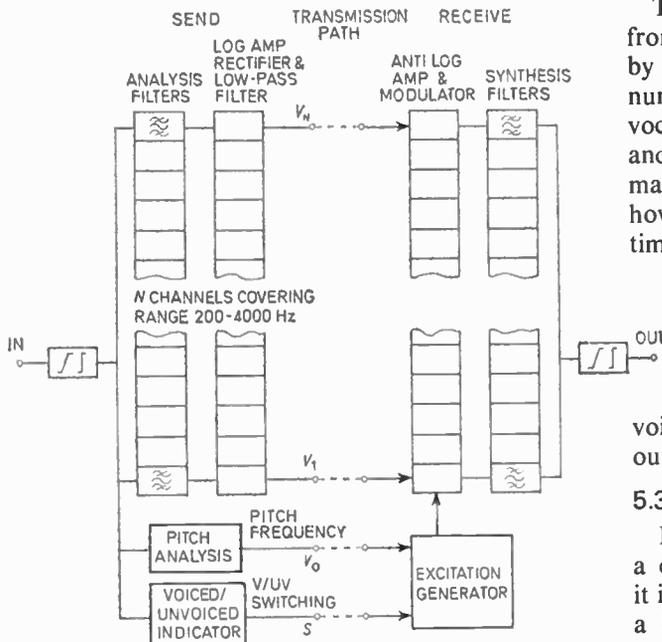
term spectral envelope of the speech signal. Parameters are extracted from these measurements, and transmitted with considerably less bandwidth than required by the speech signal. At the receiver speech is synthesized from the transmitted parameters. It is the choice of these parameters that distinguishes the different types of vocoders.

5.1. Formant Vocoders

The formant vocoder analyser attempts to measure in real time the frequency and amplitude of spectral peaks of the speech signal (the formants) and transmits these measurements as parameters to control a synthesizer similar to the one shown in Fig. 6. At the same time the larynx vibration rate and the decision as to whether the speech is voiced or unvoiced must also be measured and transmitted as additional control parameters. It is generally recognized that good speech quality can be obtained from a formant synthesizer, but when connected to a real time analyser the results obtained to date have always been inferior. This is mainly because of the difficulty in measuring formant frequencies accurately. Figure 6 shows that the formant frequency ranges overlap and this tends even more to aggravate the problem. The advantage of the formant vocoder over most other types of vocoder is the bandwidth compression that can be obtained. Typically a formant vocoder needs eight parameters and each of these can be band-limited to about 25 Hz so that a bandwidth reduction of the order of 20:1 can be obtained relative to the original speech band. It should however be emphasized that all the difficulties of real-time formant analysis are not yet solved and that until they are it is unlikely that formant vocoders will prove to be very useful.

5.2. Spectrum Channel Vocoder

This is the original channel vocoder invented in the 1930s by Homer Dudley of Bell Telephone Laboratories. In the channel vocoder the short-term spectral envelope of the speech signal is represented by samples spread across the frequency axis. These samples are usually obtained from a bank of band-pass filters whose centre frequencies are spaced across the speech band. Normally between 10 and 20 filters are used to give a corresponding number of samples of the spectral envelope; the outputs of the filters are then rectified and smoothed by low-pass filters to give a set of time varying average signals representing the short term spectral envelope. As in the formant vocoder, pitch analysis has also to be performed. The resulting signals are then transmitted to the vocoder synthesizer and are used to control the frequency response of what is in effect a time-varying band-pass filter that is fed with a spectrally flat excitation signal. The block diagram of a typical



The bandwidth reduction or compression obtained from a channel vocoder is not as large as that achieved by a formant vocoder. Typically it needs twice the number of control parameters required by a formant vocoder at about the same bandwidth per parameter and gives therefore a bandwidth reduction of approximately 10 to 1. The channel vocoder has the advantage however over the formant vocoder that at the present time it will tolerate a larger proportion of speakers'

Fig. 7. Block diagram of a typical channel vocoder.

voices and gives a more satisfactory synthetic speech output signal than the formant vocoder.

5.3. Vocoders for Digitized Speech Transmission

Most modern vocoders have been designed to use a digital transmission path. One reason for this is it is much easier to transmit a serial digit stream than a large number of low-frequency analogue signals, and another is that it is much easier to introduce privacy into a 'digital' speech link.¹ To digitize a channel vocoder both spectrum and pitch channels have first to be multiplexed and then coded by some digital coding scheme such as delta-modulation or pulse code modulation. At the receiver a digital-to-analogue converter reconstitutes the signals into an analogue form; these analogue signals are then filtered by low-pass filters with cut-off frequencies of the order of 25 Hz and are then used as control signals to the synthesizer. Digit rates used are in the order of 2000–3000 bits/s, which can be transmitted by modern data-modems over normal 3 kHz lines.

5.4. Pitch Extraction

It has been tacitly assumed so far that measurement of larynx vibration rate, or 'pitch extraction' is a fairly insignificant part of the vocoding process. In fact accurate measurement of pitch is probably one of the most difficult vocoder operations. In principle this measurement of the vocal cord vibration rate is simple, since by passing the speech waveform through a low-pass filter the fundamental frequency component can be extracted. In practice things are not so easy; the range that the pitch frequency can occupy is large (3 or 4 octaves), pitch inflexions can be rapid, and on some circuits where vocoders might be useful the fundamental component in the available input signal is weak or missing altogether. Errors that occur during pitch extraction can cause very objectionable effects in the resultant synthetic speech; for example, if the pitch extractor occasionally measures the second harmonic instead of the fundamental frequency (a common fault), the sudden 'squeak' that occurs sounds extremely unnatural.

channel vocoder is shown in Fig. 7. The analyser includes the filter bank, rectifiers and low-pass filters mentioned previously. The purpose of the equalizer shown at the analyser input is to ensure that the band-pass filters in the analyser have comparable signal levels over the whole frequency range. Another feature shown here is the presence of the logarithmic amplifiers (one per channel) connected to the outputs of the band-pass filters and whose purpose is to increase the dynamic range of the vocoder.

The low-pass filtered signals obtained at the output of the analyser are connected to the transmission path, and at the receiver are used to dynamically control the gain of each synthesis filter. The synthesis filters are supplied with a wide-band (pulse or random noise) excitation signal controlled from the analyser and each synthesis filter selects a band of excitation about its centre frequency. Summation of the output signals from the synthesis filters and further equalization yields the synthetic speech.

The vocoder, in common with other speech processing systems, must have provision for handling the wide dynamic range of the input speech signal. An experimental vocoder at the Joint Speech Research Unit has been designed to operate over a range of 50 dB; this is achieved by employing logarithmic amplifiers immediately following each analysis filter and a corresponding anti-logarithmic device preceding the synthesis filters. It is thought that for most purposes this dynamic range is sufficient, but in cases where a larger spread of input signal level is expected a Voice Operated Gain Adjusting Device (VOGAD) can be used to provide an extra dynamic range of about 20 dB.

Many schemes have been devised for the measurement of fundamental frequency. Early vocoders used a simple low-pass filter to extract the fundamental component from the speech waveform; a frequency meter was then used to derive an analogue of the pitch frequency. With a high-quality input signal and carefully spoken speech fairly good results could be obtained. However more elaborate schemes are necessary for greater tolerance to various input signals. One such scheme that gives better results uses a 'tracking' low-pass, or band-pass filter to follow changes in the fundamental frequency of the speech waveform. Pitch extractors of this type can work well over a restricted frequency range if they have as their input a high quality speech signal of good signal/ noise ratio. In the last decade autocorrelation techniques have been used to measure voice pitch. Simple autocorrelation of the speech waveform, however, does not give particularly good results during fast pitch inflexions, or during rapid formant transitions. Some pre-processing of the signal before autocorrelation may be expected to improve measurements, and Gill² has reported a successful pitch extractor that autocorrelates 'envelopes' derived from the speech waveform. More recently Sondhi³ has suggested the autocorrelation of centre clipped speech. Centre clipping the speech removes most of the zero crossings from the waveform and this has the effect of compressing or flattening the spectral envelope. This centre-clipped or spectrally-flattened speech is then fed to an autocorrelator whose task is made much easier than a correlator that has to operate directly on the speech waveform.

A different approach to fundamental frequency extraction has been described by Noll⁴ and is known as 'cepstrum'. The cepstrum is the Fourier transform of the logarithm of the power spectrum of a signal. Because a speech waveform is nearly periodic during a voiced sound it has an approximation to a line-spectrum; this spectrum has periodic ripples in it at the 'line' spacing, which corresponds to the fundamental frequency. Taking the logarithm of this spectrum compresses the peaks due to the formants and gives an equal weight to the ripples in the high-energy and low-energy regions. The Fourier transform of this log-spectrum exhibits a large peak corresponding to the pitch frequency.

In a practical cepstrum pitch extractor, log-spectra are continuously generated and the Fourier transforms calculated; the positions of the resultant cepstral peaks are then used to provide an estimate of the pitch period. Since the logarithmic spectrum will have a strong periodic component even when the fundamental is missing, this method will give good results for telephone quality speech. Cepstrum pitch extractors, although apparently successful, are ex-

tremely complicated to implement, and something simpler is desirable.

Yet another approach to pitch extraction uses the philosophy that one simple measurement of pitch is unlikely to be satisfactory, but by combining the answers of several fairly simple measurements and taking a majority vote on the results a useful pitch extractor can be obtained. Systems using this principle have been described by Gold⁵ of the MIT Lincoln Laboratory and also by Gill⁶ of J.S.R.U.

In Gill's system, measurements are made of the interval between major peaks in the speech waveform in four frequency bands in the range 0-600 Hz. These period measurements are converted into voltage analogues, each analogue representing the logarithm of the larynx vibration frequency over a range from 37.5-600 Hz. The analogue voltages of the four channels are compared with each other and also with the last transmitted measurement three times every 20 ms. If during this time a good measure of agreement is obtained the majority measurement is used for the next transmitted value of pitch. If good agreement is not obtained but the sound is judged to be voiced then a 'no-confidence' signal is transmitted to the receiver causing it to continue using the previous measurement.

The voiced-unvoiced decision is made by comparing the energy in the band 200-600 Hz with that in the range 5000-7000 Hz. A high ratio of low frequency to high frequency energy indicates a voiced sound and a high ratio of high frequency to low frequency energy an unvoiced sound. This decision is influenced by the pitch measurement; when a confident pitch measurement is obtained the low frequency/high frequency ratio can be much lower and still interpreted as a voiced sound.

At the receiver the pitch analogue signal is passed through a 10 Hz low-pass filter, the cut-off frequency of this filter is switched to a higher value at the beginning of each voiced sound in order to get the analogue signal to its desired new value quickly.

This pitch extractor will operate with very few gross pitch errors or errors of voicing detection, on rapid conversational speech.

The formant vocoder and the channel vocoder are the most well known of the analysis/synthesis systems but several other types of vocoder do exist and a few of these will now be described.

5.5. Voice Excited Vocoders

Many workers have attempted to improve the speech quality of the channel vocoder. One useful solution which considerably eases the problem of pitch extraction and at the same time improves the received speech quality is the voice excited vocoder.

The voice excited vocoder differs from the channel vocoder in two main ways. Firstly a baseband of natural speech is transmitted (usually in the range 200–800 Hz), and is used at the receiver instead of synthetic speech for the lower part of the spectrum, and secondly at the receiver the baseband is used after some processing to provide the vocoder excitation signal for the frequency region above the baseband. Channel signals for the band above the baseband are transmitted as in a normal channel vocoder.

Generation of a suitable flat spectrum excitation signal from the baseband requires some form of non-linear processing, often called spectral flattening, and there are several methods for doing this. A simple example of spectral flattening is half-wave rectification followed by a suitable weighting network to produce a spectrum which is approximately flat in the required band.

Generally speaking voice-excited vocoders provide much better quality speech than simple channel vocoders. Some of this improvement is obviously due to the fact that part of the natural speech is being directly transmitted, but the rest of the improvement obtained is probably due to the inherently more accurate excitation signal obtained from the spectral flattening process. The voice excited vocoder will give a bandwidth compression of about 3:1 compared with the 10:1 obtained from a typical channel vocoder.

5.6. Autocorrelation Vocoder

The autocorrelation function of a signal is the Fourier transform of its power spectrum. It therefore follows that the short term spectrum of a speech signal can be represented by a set of short term autocorrelation functions. This is the basis of the autocorrelation vocoder, originated by Schroeder.⁷ In this type of vocoder, analysis is performed by multiplying the signal by itself at various delays (delay increments must be less than half the reciprocal of the required speech bandwidth). The multiplier output signals are then low-pass filtered to about 25 Hz to produce the short term autocorrelation channel signals. These signals after transmission are used to control a speech synthesizer. The synthesizer is a time varying filter, whose impulse response is a replica of the short term autocorrelation function for the range of delays used in the analyser. Synthesis is achieved by 'sampling' the channel signals with the excitation signal in a set of multipliers, and then assembling the samples in the correct time relationship by use of a delay line. Since the correlation function is even it will be symmetrical about the centre sample. This means in practice that a delay line that gives a complete reflexion at one end can be used, with a consequent halving of the delay-line length.

Synthetic speech produced by the autocorrelation vocoder will have a spectral envelope corresponding to the square of the spectrum of the original speech signal, since in the analyser the speech signal was multiplied by itself. This results in making the spectral peaks more pronounced than they normally would be, and gives the synthetic speech a rather unnatural 'bouncy' quality. This defect may be overcome by using a complex equalizer which effectively takes the square root of the input signal spectrum. Figure 8 shows a block diagram of an autocorrelation vocoder.

5.7. Harmonic Vocoder

The harmonic vocoder or harmoniphone first reported by Pirogov⁸ is yet another variation of the vocoder principle. In this vocoder the short term spectral envelope of the speech signal is expanded into a Fourier series; there are several methods for achieving this, but a fairly straightforward way is by means of a resistor matrix connected to the low-pass filtered channel signals of a normal channel-vocoder analyser. The time-varying Fourier coefficients obtained from this matrix are then used after transmission to control a speech synthesizer.

The synthesizer takes the form of a set of wide-band non-dispersive, interconnected delay lines. This network has frequency responses of $\sin nD\omega$, $\cos nD\omega$, available at a series of tapping points; n takes the value 0, 1, 2, etc., and D is the delay of one delay section. The excitation signal is connected to this network, and the signals at the tapping points are multiplied by the appropriate Fourier coefficient and are then summed in order to produce the synthetic speech.

An interesting feature of this type of vocoder is that each single coefficient affects the whole spectrum of the synthetic speech, and not just a portion of it as in the channel vocoder. It is found in practice that the higher coefficients have a progressively smaller effect on the spectrum than the lower ones. This would generally mean that errors in transmission of coefficients near the spectrum fundamental would have a more serious effect on the speech quality, and therefore one would expect that these coefficients would need to be transmitted more accurately than those corresponding to the higher harmonics of the spectrum shape.

5.8. Chebyshev Vocoder

A further refinement of the harmonic vocoder principle is the Chebyshev vocoder first reported by Kulya.⁹ This vocoder expands the short term spectral envelope of the speech into an orthogonal series in the same way as the harmoniphone; in this case however the functions used in the expansion are not

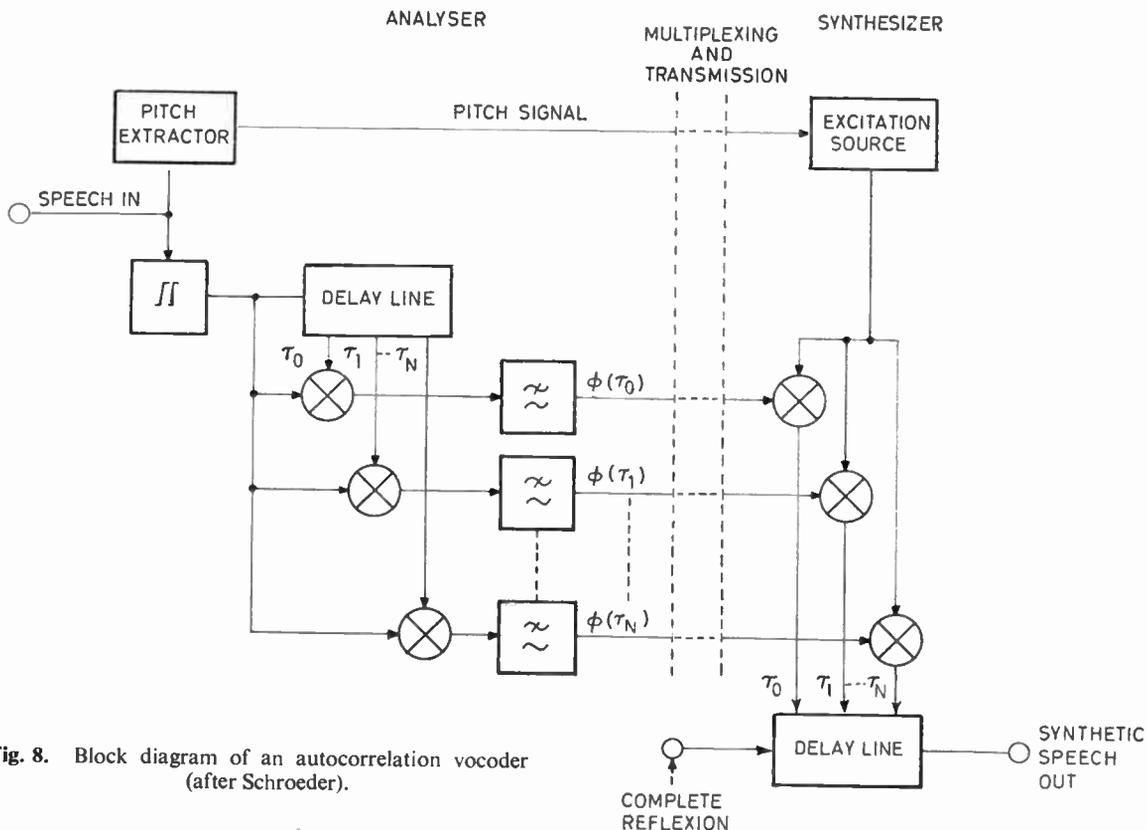


Fig. 8. Block diagram of an autocorrelation vocoder (after Schroeder).

sine and cosine waves but transformed Chebyshev functions. This may at first sight appear to be an unnecessary complication but it does in fact have two main advantages. Firstly, the Chebyshev functions scan the short-term spectral envelope in such a way that more accuracy is obtained at the low-frequency end of the spectrum. Since the resolution of the ear is believed to be distributed in a similar manner, this greater low-frequency accuracy would seem a desirable feature. Secondly, we have seen that synthesis in terms of Fourier series requires the use of wide-band non-dispersive delay lines which are bulky and difficult to design; it can be shown, however, that the use of Chebyshev functions requires a set of dispersive

delay networks that can be simply constructed from RC networks known as Laguerre networks. Speech quality from a Chebyshev vocoder can be at least as good as from a channel vocoder with the same number of control parameters, and there is some evidence to show that if we are restricted to a small number of coefficients (e.g. five) that the speech quality can be better than for the corresponding channel vocoder.

Figure 9 shows a conventional 18-channel vocoder analyser, modified by the use of a matrix of 180 resistors, that converts it into a Chebyshev vocoder analyser. Figure 10 shows the block diagram of the corresponding 10-coefficient synthesizer.

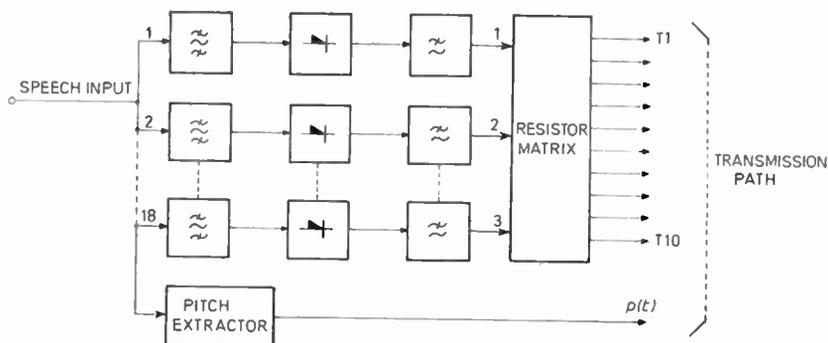


Fig. 9. Block diagram of a 10-coefficient Chebyshev vocoder. This was obtained by modifying an 18 channel vocoder analyser with the resistor matrix shown.

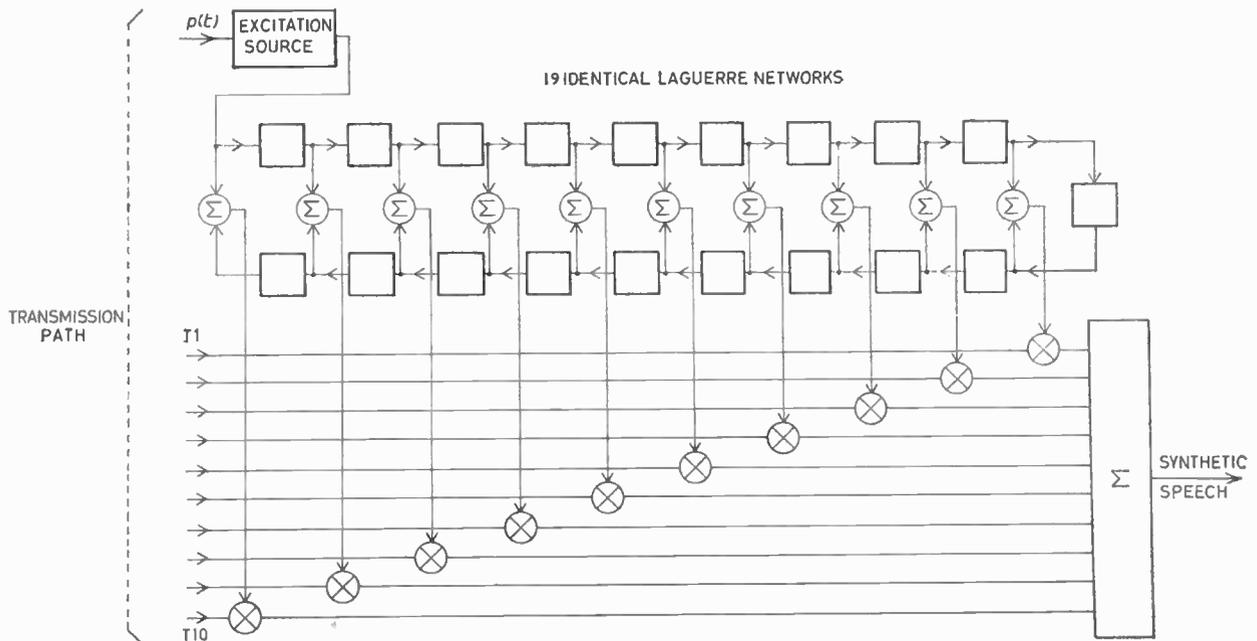


Fig. 10. Block diagram of a 10-coefficient Chebyshev vocoder synthesizer.

6. Conclusions

Since the channel vocoder was invented in the 1930s progress in vocoder research has been relatively slow. This is mainly due to the complicated nature of speech, the complexity of electronic circuitry required for a vocoder terminal, and the large amount of prototype construction and testing needed to optimize the design of the analysis and synthesis filters. With the improvements available in circuit techniques it is, however, now quite practicable to construct a vocoder terminal in a volume of about a cubic foot, and further reductions of size are likely.

Present and future research on vocoders is essentially a task for digital computers. It is now possible to simulate complete vocoders on a digital computer. Recordings of natural speech may be processed by a simulated vocoder, whose design parameters may be quickly optimized; for example the parameters corresponding to complete analysis or synthesis filter banks can be changed simply by altering a few numbers on a data tape. This simulation process obviously outstrips anything that can be done in the laboratory with hardware, but has one main disadvantage; a typical vocoder simulation program takes something like 300 times real time on a moderately fast computer. This obviously limits the amount of speech that can be processed so that one still needs hardware available for final checking of an optimized design on a large volume of speech material. Vocoders are now reaching the stage where they are becoming practical communication devices.

The next generation of vocoders may well be of the formant type, when the existing problems of formant tracking have been solved. These vocoders should provide improved speech quality at a lower transmission rate than present day channel vocoders.

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Fast Pseudo-Random Number Generators for Computers

By

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Available techniques for the production of random numbers for digital computers rely on software which is relatively slow. Simple techniques for generating these numbers are available in hardware form which allows the full speed of the computer to be used to full advantage in simulation problems. Several possible practical arrangements are considered and their performances compared.

1. Introduction

Random number generators are necessary in digital computers for simulation experiments in both scientific and business studies. Simulation experiments cannot be avoided since they arise from problems to which no analytical solution can be found or from problems which cannot even be expressed in tractable equation form. Often the mathematical approach is so difficult that the program has largely to reproduce the real physical situation. Intuitively it is clear that a program of this nature will be extremely time-consuming in comparison with programs which have only to determine numerical solutions to partly-solved mathematical equations.

Digital computers are becoming employed to full capacity, i.e. 24 hours a day, and while it is unlikely that computer manufacturers will allow the demand for machines to outstrip their production capability or fail to develop even faster machines than are at present available, it is to be hoped that they will ensure the most efficient use of their products to be made. By this is meant that in the majority of circumstances, compilers should be able to produce the most efficient set of machine instructions from the source program as is possible.

Manufacturers have gone some way towards achieving this situation by the introduction of hardware packages which perform basic arithmetic operations such as multiplication and division. Some typical hardware devices may be several orders of magnitude faster than their software counterparts. This is, however, only the beginning of the story for many of the more often used functions could be converted to hardware package form with consequent improvements in overall programming time and, therefore, more efficient computer utilization. The author knows of at least one company at present examining the possibility of incorporating the basic external functions into a hardware (as opposed to software) package.

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This paper aims to show how the random number generator—possibly one of the most time-consuming functions incorporated in simulation programs—may be easily implemented in hardware form. The improvement in time to produce a random number appears to be several orders of magnitude.

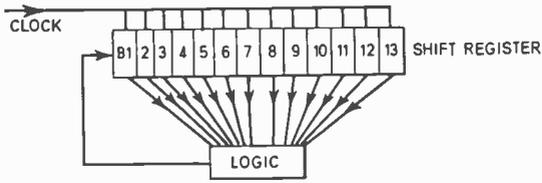
2. Development of a Random Number Generator

At the present time, the majority of software random number generators for use in digital computers are of the multiplicative or congruential types.^{1,7} In general, the form of such generators is:

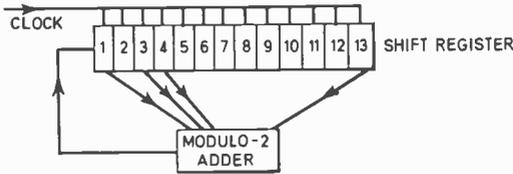
$$x_{i+1} = (kx_i + c) \bmod (m), \quad i = 0, 1, 2, \dots \dots (1)$$

These clearly rely on suitable choices for k and c to yield numbers which have the required probability density function and also generate long sequences. An obvious advantage is that this sort of calculation can be performed relatively quickly (perhaps 500 computer cycles). Unfortunately, values for k and c may have to be found by a tedious Monte Carlo technique to ensure that the statistical properties of such generators are entirely satisfactory.

Another method for producing a sequence of random numbers is provided by a shift register with feedback. Feedback shift registers have been extensively examined in the literature^{2,3,4} and it is only necessary here to give a brief outline of their operation. A shift register basically consists of a set of bistable multivibrators connected together so that state of each element is transferred to the next element on reception of a clock pulse. Thus information appearing at the input to the first bistable element becomes stored, and the information stored on the last element of the shift register is lost. The shift register can remain active if some binary sequence is fed to the shift register input while it is being clocked. A simple way of ensuring that an input is always present is to derive such a sequence by performing some logic function on the combined states of several of the elements in the register as shown in Fig. 1(a). In particular, when the logic function in the feedback



(a) Connexion of feedback shift registers to control logical functions.



(b) Feedback shift registers connected to perform modulo-2 addition.

Fig. 1.

path is modulo-2 addition of the states of an even number of elements, as in Fig. 1(b), it is possible to show theoretically that the resulting sequence has many interesting and useful properties. This form of feedback is called linear and the whole shift register with feedback is called a linear feedback shift-register (linear f.s.r.). It is possible to characterize the different possible feedback configuration by a simple equation of the form:

$$f(x) = 1 - \sum_j x^j$$

where the summation is taken over all elements, j , from which an output is taken to the modulo-2 adder.

3. Properties of a Linear F.S.R.

By choosing suitable connections to the modulo-2 adder, it can be shown that the ensuing sequence has a period of $2^n - 1$. The polynomial $f(x)$ then divides $x^L - 1$ for $L = 2^n - 1$ but for no L smaller. The equation $f(x) = 0$ is then primitive modulo-2. The shift register therefore must contain at least n stages. If use is to be made of the whole shift register, then clearly one of the feedback inputs to the modulo-2 adder must come from the end of the shift register. This indicates that the minimum shift register must be composed of exactly n bistable elements, and a sequence length of $2^n - 1$ is known as a maximum length sequence.

The number of ways of choosing n binary digits is exactly 2^n so that every binary pattern but one occurs exactly once in the shift register sequence. The pattern which does not occur is the one comprising all 0's.

This is self-perpetuating since the output of a modulo-2 adder whose inputs are always zero is also zero. The pattern of 1's and 0's in the shift register can be considered as a binary number with the leftmost digit the most significant and the rightmost digit the least significant. In this situation, we can say that every integer between 1 and $2^n - 1$ occurs exactly once, that is, there is a uniform probability density function for the binary number appearing in the register. Thus shift register pseudo-random number generators satisfy one of the requirements of computer random number generators.

The other requirement for random number generators is that there should be no serial correlation between random numbers at least up to the tenth and hopefully more. The shift register in its standard form does not fulfil this condition as can easily be seen from the following argument. If the binary number in the shift register at any one time is A , and if the feedback input is to the leftmost bistable multi-vibrator, the next number appearing in the register is:

$$\begin{matrix} 1 \\ \text{g.i.l.}(A/2) + (\text{or}) \times 2^{n-1} \text{ (g.i.l.: greatest integer less than)} \\ 0 \end{matrix}$$

Thus the correlation coefficient relating this term to the last is approximately 0.5. In general, under these circumstances, the correlation coefficient relating one number to the m th number later is less than $\frac{1}{2}^m$ provided that $m < n$. When $m > n$ there will be some correlation and this is exactly $-3/(2^n - 2)$. The Appendix develops exact formulae for the auto-correlation of binary weighted words in a maximum length sequence. One way to ensure that a number from the shift register sequence is not seriously correlated with the next number in the sequence is by clocking the shift register until all the digits comprising the first number are removed and a completely new set of digits fills the register. The minimum number of clock pulses required to do this is equal to the number of elements in the shift register. To see how this affects the properties of the sequence, it is convenient to visualize the complete sequence of binary numbers as a state diagram. Each state has a unique predecessor and a unique successor and the state diagram is a circle. This is illustrated in Fig. 2.

Clocking the register n times between the output of each random number is then equivalent to jumping n states on the state diagram. If the length of the original sequence $2^n - 1$, has a factor, n , jumping n states between each number will effectively shorten the available sequence because we return to the original point after only one cycle of the state diagram. In general we can therefore say that the sequence length is unaltered if n and $2^n - 1$ have no factors in common, i.e. if they are relatively prime. If this is

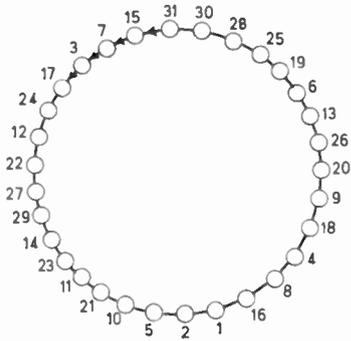


Fig. 2. State diagram for a 5-stage linear f.s.r

not the case, the sequence length is:

$$\frac{\text{l.c.m.}(n, 2^n - 1)}{n}$$

However, when the sequence is shortened in this way, we cannot be sure of the uniformity of the output probability density function as we were in the case of a maximum length sequence.

When $2^n - 1$ and n are relatively prime, the number of random numbers available before the output sequence comes within n digits of the starting point on the state diagram is nearly $(2^n - 1)/n$. In other words no serious correlation of random numbers occurs until approximately the $(2^n - 1)/n$ one. Word lengths in computers range between 23 and 48 bits, so even with $n = 23$, serial correlation will be small until the 250,000th random number. This is adequate for most computer applications.

The whole object of the concept of hardware implementation of often used functions is primarily a decrease in average program time. There are two timing aspects of importance. Firstly, in this case, to ensure that a new random number is available as quickly as possible and secondly, to minimize program access time to this function. The second problem is one of internal computer organization and will not therefore be discussed here. The first one, however, is very relevant to the design of a hardware number generator and it will now be shown that it is not difficult to arrange a shift register in such a way that it jumps n states on the reception of a single clock pulse.

4. A Fast Random Number Generator

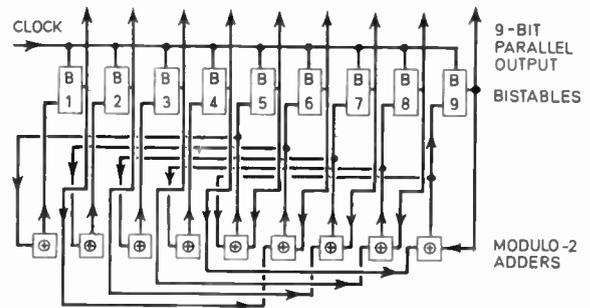
At any one time, we can label the states of the bistable elements comprising the shift register of Fig. 1(b) such that the leftmost digit is a_1 and the rightmost is a_n .

When the register is clocked, these states will propagate down the register from left to right. In particular, after the reception of one clock pulse, the

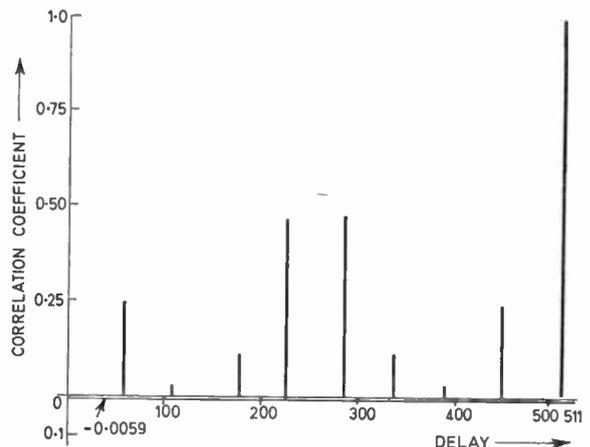
new states of each element will be a_0 in the leftmost or first bistable, a_1 in the second bistable, and so on with a_{n-1} in the n th bistable. Thus before the reception of any clock pulses, the output of the modulo-2 adder or the input to the leftmost bistable is a_0 .

After n clock pulses, the states of the elements of the shift register are $a_{1-n}, a_{2-n}, a_{3-n} \dots a_0$ reading from left to right, i.e. with a_0 in the n th bistable element. It is not difficult to include extra modulo-2 adders operating on the digits a_0 to a_n to form the new digits a_{1-n} to a_0 . These new digits are fed into the register in parallel on the reception of a single clock pulse so that a_0 now occupies the position which a_n previously occupied and so on. Thus the register has jumped n positions on the state diagram on the reception of a single clock pulse.

There is clearly no requirement for the bistable states to propagate down the register and the final form of the number generator is a collection of bistable multivibrators joined together by modulo-2 adders. The circuit diagram of a 9-bit random number generator and the autocorrelation of its output sequence are shown in Figs. 3(a) and 3(b). There is



(a) 9-bit pseudo-random number generator block diagram.



(b) Autocorrelation of output sequence of circuit in Fig. 3(a).

Fig. 3.

no difficulty in extending the principle to longer random number generators. It can also be seen from Fig. 3(a) what logic delays are involved in the calculation of a new random number. The delay is just that due to one bistable element and two exclusive OR gates. Since the introduction of integrated circuit logic which can operate at 400 MHz, it seems plausible to expect that a new random number could be calculated every 10 or 20 ns. Since this is a much shorter time than that in which the majority of computers can even perform a simple instruction it seems probable that the ultimate limitation upon the speed of this random number generator will be imposed by other considerations. For example, low-speed logic could be used to reduce cost, or the speed could be limited by purely organizational problems within the computer.

In conclusion, it has been shown that a pseudo-random number generator with well-controlled correlation and probability properties is readily available as a particular logic connexion of computer type elements. Perhaps its major advantage is one of speed—it should be able to operate, given satisfactory conditions, several orders of magnitude faster than software routines for producing random numbers. Computing machines with such hardware random number generators should therefore be capable of processing simulation or Monte Carlo synthesis programs much more rapidly than at present.

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6. Appendix: Autocorrelation of Weighted Words from Maximum Length Pseudo-random Sequences

We assume that the maximum length sequence is generated by an n -stage binary shift register so that the sequence length is $2^n - 1$. We also assume, in the general case, that there is an additional binary shift register of p stages acting as an extension of the sequence generating register. We will denote the

weight of the i th stage of the total shift register as W_i . Outputs are taken from q stages only.

$$1 \leq i \leq q$$

and

$$q \leq n + p$$

$$p \geq 0.$$

The output from the weighting network after the shift register has received the j th clock pulse is x_j so that

$$x_j = \sum_{i=1}^q a_{i+j} W_i \quad \dots\dots(1)$$

where a_{i+j} represents the state of the i th bistable after the reception of the j th clock pulse from some arbitrary point in time. The normalized autocorrelation function may now be written as

$$C(k) = \frac{\sum_{j=1}^{2^n-1} (x_j - \bar{x})(x_{j+k} - \bar{x})}{\sum_{j=1}^{2^n-1} (x_j - \bar{x})^2} \quad \dots\dots(2)$$

where

$$\bar{x} = \frac{1}{2^n-1} \sum_{j=1}^{2^n-1} (x_j).$$

We will first of all consider \bar{x} . This may be rewritten as

$$\bar{x} = \frac{1}{2^n-1} \sum_{j=1}^{2^n-1} \sum_{i=1}^q a_{i+j} W_i \quad \dots\dots(3)$$

Interchanging the order of the summations and allowing a_{i+j} to take the values 0 or 1 we find

$$\bar{x} = \frac{1}{2^n-1} \sum_{i=1}^q W_i \sum_{j=1}^{2^n-1} a_{i+j} \quad \dots\dots(4)$$

The summations are independent since the starting point of the sequence is irrelevant. Since there are exactly 2^{n-1} 1's in a maximum length sequence and $(2^{n-1} - 1)$ 0's, the final form for \bar{x} is

$$\bar{x} = \frac{2^{n-1}}{2^n-1} \sum_{i=1}^q W_i \quad \dots\dots(5)$$

If the W_i are binary coefficients with $W_i = 2^{q-i}$, then \bar{x} is given by

$$\bar{x} = \frac{2^{n-1}(2^q - 1)}{2^n - 1} \quad \dots\dots(6)$$

We will now turn our attention to the denominator of equation (2). This may be rewritten:

$$\begin{aligned} & \sum_{j=1}^{2^n-1} (x_j^2 - 2x_j\bar{x} + \bar{x}^2) \\ &= \sum_{j=1}^{2^n-1} x_j^2 - (2^n-1)\bar{x}^2 \\ &= \sum_{j=1}^{2^n-1} \left[\sum_{i=1}^q a_{i+j} W_i \right]^2 - (2^n-1)\bar{x}^2 \quad \dots\dots(7) \end{aligned}$$

The bracketed and squared summation may be multiplied out to yield all the squared terms and the cross-products. After doing this the equation may be written:

$$\sum_{j=1}^{2^n-1} \left[\sum_{i=1}^q a_{i+j}^2 W_i^2 + 2 \sum_{i=1}^q \sum_{l=i+1}^q a_{i+j} a_{l+j} W_i W_l \right] - (2^n-1)\bar{x}^2. \quad \dots\dots(8)$$

Clearly there are as many 1's in the sequence a_{i+j} as there are in a_{i+j}^2 , that is 2^{n-1} 1's. There are only 2^{n-2} 1's in the sequence $a_{i+j} a_{l+j}$ when i is not equal to l . The order of the summations may again be interchanged so that equation (8) becomes

$$\begin{aligned} & 2^{n-1} \sum_{i=1}^q W_i^2 + 2^{n-1} \sum_{i=1}^q \sum_{l=i+1}^q W_i W_l - \\ & - \frac{2^{n-2}}{2^n-1} \left[\sum_{i=1}^q W_i^2 + 2 \sum_{i=1}^q \sum_{l=i+1}^q W_i W_l \right] \\ & = 2^{n-2} \sum_{i=1}^q W_i^2 + \left[2^{n-2} - \frac{2^{2n-2}}{2^n-1} \right] \times \\ & \quad \times \left[\sum_{i=1}^q W_i^2 + 2 \sum_{i=1}^q \sum_{l=i+1}^q W_i W_l \right] \\ & = 2^{n-2} \sum_{i=1}^q W_i^2 - \frac{2^{n-2}}{2^n-1} \left[\sum_{i=1}^q W_i \right]^2. \quad \dots\dots(9) \end{aligned}$$

Again, if the W_i are binary coefficients with $W_i = 2^{q-i}$ then equation (9) becomes

$$2^{n-2} \sum_{i=1}^q 2^{2q-2i} - \frac{2^{n-2}}{2^n-1} \left[\sum_{i=1}^q 2^{q-i} \right]^2. \quad \dots\dots(10)$$

A proof by induction shows that

$$\sum_{i=1}^q 2^{2q-2i} = \frac{2^{2q}-1}{3}$$

since

$$\begin{aligned} \sum_{i=1}^{q+1} 2^{2(q+1)-2i} &= \sum_{i=1}^q 2^{2q-2i} + 2^{2q} = \frac{4 \times 2^{2q}-1}{3} \\ &= \frac{2^{2(q+1)}-1}{3} \end{aligned}$$

Thus equation (10) becomes

$$2^{n-2} \frac{(2^{2q}-1)}{3} - \frac{2^{n-2}}{2^n-1} (2^q-1)^2$$

When $q = n$, this contracts to

$$\frac{[(2^n-1)^2-1](2^n-1)}{12}$$

As a matter of interest the equations (7) to (10) represent 2^n-1 times the mean squared deviation from the mean of the random process, x_j , the mean value of which is \bar{x} . Let us digress for a moment and consider the meaning of these equations in relation to the crest factor. This may be defined as the ratio of the peak deviation from the mean to the root mean

squared deviation from the mean of the random process. It is somewhat difficult to define the peak deviation from the mean in all cases because it is dependent on the exact structure of the sequence when q is greater than n . However, when q is equal to or less than n , the peak deviation is obtained when all the relevant a_i are 0 or 1. Thus the crest factor k is

$$k = \frac{\sum_{i=1}^q W_i - \frac{2^{n-1}}{2^n-1} \sum_{i=1}^q W_i}{\left[\frac{2^{n-2}}{2^n-1} \sum_{i=1}^q W_i^2 - \frac{2^{n-2}}{(2^n-1)^2} \left[\sum_{i=1}^q W_i \right]^2 \right]^{\frac{1}{2}}} \quad n \geq q.$$

When n is large, this approximates to

$$k \approx \frac{\sum_{i=1}^q W_i}{\left[\sum_{i=1}^q W_i^2 \right]^{\frac{1}{2}}}$$

For clarity, we will make the assumption that

$$\sum_{i=1}^q W_i = 1$$

then

$$k \approx \frac{1}{\left[\sum_{i=1}^q W_i^2 \right]^{\frac{1}{2}}}$$

The minimum value of the denominator of this expression occurs when all the W_i are equal to $1/q$ and the maximum attainable crest factor then becomes

$$k \approx \sqrt{q}$$

From the noise generator point of view, uniform weighting of the outputs from the shift register does not yield a very satisfactory spectrum and the more usual $(\sin x/x)$ weighting will worsen the crest factor. It is also interesting to note that the crest factor only increases approximately as the square root of the number of shift register outputs that have been used.

To continue with the main theme, we now have to evaluate the numerator of equation (2). Multiplying out, we find

$$\sum_{j=1}^{2^n-1} (x_j - \bar{x})(x_{j+k} - \bar{x}) = \sum_{j=1}^{2^n-1} x_j x_{j+k} - (2^n-1)\bar{x}^2 \quad \dots\dots(11)$$

since the average of x_j or x_{j+k} is just \bar{x} .

Substitution of equation (1) in (11) gives

$$\begin{aligned} \sum_{j=1}^{2^n-1} x_j x_{j+k} - (2^n-1)\bar{x}^2 &= \sum_{j=1}^{2^n-1} \left[\sum_{i=1}^q a_{i+j} W_i \right] \times \\ &\times \left[\sum_{i=1}^q a_{i+j+k} W_i \right] - (2^n-1)\bar{x}^2. \quad \dots\dots(12) \end{aligned}$$

When the sequence a_{i+j} is identical with a_{i+j+k} , their product sequence contains exactly 2^{n-1} 1's. This

occurs for those values of i and l which satisfy the following equation:

$$i = l + k. \quad \dots(13)$$

When the sequences a_{i+j} and a_{l+j+k} are not identical, that is for values of i and l which do not satisfy equation (13), their product sequence contains exactly 2^{n-2} 1's.

Rewriting the right-hand side of equation (12) by expanding the bracketed portions and separating out the identical and non-identical sequences, we find that equation (12) becomes

$$\begin{aligned} &\sum_{j=1}^{2^n-1} \left[\sum_{l=1}^{q-k} a_{l+j}^2 W_l W_{l+k} + \sum_{i=1}^k \sum_{l=1}^q a_{i+j} a_{l+j+k} W_i W_l + \right. \\ &\quad + \sum_{i=k+1}^q \sum_{l=q-k+1}^q a_{i+j} a_{l+j+k} W_i W_l + \\ &\quad + \sum_{i=k+1}^{q-1} \sum_{l=i-k+1}^{q-k} a_{i+j} a_{l+j+k} W_i W_l + \\ &\quad \left. + \sum_{i=k+2}^q \sum_{l=i-k-1}^{q-k-1} a_{i+j} a_{l+j+k} W_i W_l \right] - \\ &\quad - (2^n-1)\bar{x}^2. \quad \dots(14) \end{aligned}$$

Clearly when $k = q$ the first term disappears because x_i and x_{i+k} contain no common parts of the shift register sequence. Interchanging the order of the summations and performing the summation over j , we find that equation (14) becomes:

$$\begin{aligned} &2^{n-2} \sum_{l=1}^{q-k} W_l W_{l+k} - 2^{n-2} \sum_{i=1}^q \sum_{l=1}^q W_i W_l - (2^n-1)\bar{x}^2 \\ &= 2^{n-2} \sum_{l=1}^{q-k} W_l W_{l+k} + 2^{n-2} \left[\sum_{i=1}^q W_i \right]^2 - \\ &\quad - \frac{2^{2n-2}}{(2^n-1)} \left[\sum_{i=1}^q W_i \right]^2. \quad \dots(15) \end{aligned}$$

The massive contraction between equations (14) and (15) may look a little unbelievable at first, however, we have taken 0.5 of the first term inside the brackets of (14) and added it to the other four terms to obtain the second term on either side of equation (15). Thus equation (2) becomes:

$$C(k) = \frac{\sum_{i=1}^{q-k} W_i W_{i+k} - \frac{1}{2^n-1} \left[\sum_{i=1}^q W_i \right]^2}{\sum_{i=1}^q W_i^2 - \frac{1}{2^n-1} \left[\sum_{i=1}^q W_i \right]^2} \quad \dots(16)$$

which is conveniently symmetrical.

We will now insert binary weighted coefficients for W_i in equation (16) and obtain the particular normalized autocorrelation function for binary weighted words of pseudo-random sequences.

$$\begin{aligned} C(k) &= \frac{2^k \sum_{i=1}^{q-k} 2^{2(q-i-k)} - \frac{(2^q-1)^2}{2^n-1}}{\sum_{i=1}^q 2^{2(q-i)} - \frac{(2^q-1)^2}{2^n-1}} \\ &= \frac{(2^n-1)2^k(2^{2q-2k}-1) - 3(2^q-1)^2}{(2^{2q}-1)(2^n-1) - 3(2^q-1)^2}. \end{aligned}$$

Often the binary word which represents the random number will comprise all the outputs from the shift register, then $q = n$ and the autocorrelation function becomes

$$C(k) = \frac{2^{2n-k} - 3 \times 2^n - 2^k + 3}{2^{2n} - 3 \times 2^n + 2}.$$

When n is large and k is small, this equation may be approximated by $C(k) = 2^{-k}$.

When $k = n$, we obtain the out-of-phase correlation coefficient which holds for all values of k in the shown range:

$$\begin{aligned} C(k) &= \frac{-3(2^n-1)}{2^{2n}-3 \times 2^n+2} \quad n \leq k \leq 2^n-1-n \\ &= \frac{-3(2^n-1)}{2^n(2^n-1)-2(2^n-1)} \\ &= \frac{-3}{2^n-2}. \end{aligned}$$

Negative values of k may be accommodated by writing $-k$ for k everywhere since the autocorrelation of real functions is always even.

As a final note, it is worth mentioning that the general result in equation (16) may be applied to many situations involving the weighted outputs of binary shift registers. It is beyond the scope of this paper to discuss even some of them. Clearly the above equations may be further generalized using the same techniques to deal with weighted outputs from m -ary shift registers.

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Optical Storage

By

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Presented at a Joint I.E.R.E.-I.E.E. Computer Groups Colloquium on 'Optical Storage Techniques' held in London on 10th April 1969.

The paper discusses the requirements for information processing system components, drawing attention to the parallel manner in which systems and components evolve. A short survey follows of optical techniques which may be applied to digital storage, and of systems which have been produced in the past. Finally, suggestions are made as to optical storage systems which should have very desirable characteristics.

1. Introduction

Optical stores are not used in present-day computing systems. Accordingly, this paper will deal with the following:

- (1) Requirements for new components of information processing systems;
- (2) Optical techniques available for satisfying these requirements;
- (3) Some proposals for optical stores which have already been published; and
- (4) Some proposals which seem to be most promising.

2. Requirements

A computing system is a complex assembly of components such as logical networks, storage devices, input and output equipment, and programs. The assembly is designed by systems engineers to make the best use of the components available and to carry out a range of information processing tasks which are believed to be commercially viable. It follows from this that the evolution of computing systems is somewhat analogous to the growth of a living organism adapting itself to its environment. In this case the environment is the range of devices available to the systems engineer who always adapts his design to those components which he believes to be available and most suitable for his purpose. It is not possible in this shifting scene to recognize a clear requirement for a new component which would improve the productivity/cost ratio of a computing system, so that we come to the dismal conclusion that there is no clearly recognizable requirement for optical logical or storage devices in information processing systems. Neither was there any requirement for disk stores, magnetic tape stores, or even core stores, since at various times in the evolution of the computer industry viable information processing systems have been built and used without incorporating these components which we now regard as essential. We

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must, therefore, look more closely into the parallel symbiotic processes whereby computing devices and computing systems evolve. These processes can be described as the ever more accurate matching of the properties of computing and storage devices to the way in which they are used in a computing system. If therefore we study the negotiable parameters of computing and storage devices and of information processing and storage tasks within a computing system, it is possible that we may find a potential match between them which has not yet been exploited in the present generation of computing systems. By this process we may create at the same time both a requirement for a new device and a method of profitably incorporating it in a computing system.

Even using this method of seeking new requirements it is still very difficult to see an opening for optical engineering in the information *processing* field. Elegant optical information processing devices have been proposed but they are essentially inflexible analogue processors which can carry out a single process such as a Fourier transformation by virtue of the fact that certain optical phenomena, in this case wave diffraction, obey laws which can be easily described in terms of the required function.

A programmed general-purpose processor on the other hand is necessarily an assembly of amplifying devices and it is very difficult to devise a practical amplifying device which uses light to convey the input and output information signals.

Considering now the information *storage* problem we can list the relevant negotiable parameters of stores and their methods of use in a system such as:

- (i) Size;
- (ii) Read activity;
- (iii) Write activity;
- (iv) Is the access pattern systematic?
- (v) If the random access pattern involves short groups of consecutive words, how large are such groups?

In a typical present system we already include diverse storage devices which are matched to the requirements usually with the assumption that items

(ii) and (iii) above, the read activity and write activity, are equal. The evolution of computing systems is directed by their designers to match them to their human users by making them more versatile in the sense that they can respond in an intelligent way to an ever-growing variety of human requests. This is done by building into the systems very large system programs which must be accessible for reading in a computer time scale, e.g. at a microsecond's notice, but need not be written or changed unless a human designer has discovered the necessity for it. There is, therefore, a large quantity of information handled in present-day computing systems which has a very great asymmetry in its read and write activities, and this class of information may be suitable for storage in an optical store in which, as will be seen later, engineering problems which are at present difficult or insoluble may be bypassed by allowing the writing process to be slow.

The most difficult design problem in an information store is typically the address selection device, and for this purpose amplification is not required so that passive optical devices such as lenses and mirrors may be used to advantage.

Present-day computing systems commonly incorporate virtual stores which are accessed by indirect methods using a directory which has to translate from a symbolic identifier used in the program to the physical location of the item of information required. It is, therefore, not difficult in such a system to provide the illusion that a store with no write facilities at all can in effect be written into by causing the directory to point to an erasable store instead of the fixed store. The up-dating information can then be written into the erasable store leaving the out-of-date information still in the fixed store but with no means of reference to it.

3. Some Digital Optical Techniques

It is useful to list the optical techniques which may be useful for incorporation into information processing systems, and describe them in information processing terms.

3.1. Geometrical Optical Devices

Many conventional optical instruments have been devised with the intention of transforming and processing a pictorial image. For this purpose scattered light is to be deplored, but can often be tolerated since it simply causes the images to suffer a loss of contrast.¹ If we seek, however, to use geometrical optical devices to transfer digital information, the loss of contrast due to scattered light (veiling glare) can be very serious, since, for example, 50% scattered light would render zeros and ones indistinguishable.

With the proviso, therefore, that proper precautions

are taken to minimize scattered light a lens is a very powerful information transferring device. It may be considered to comprise 10^6 – 10^8 separable parallel information channels (the number of resolved points in the field of view), each with a very high potential data rate. Thus, for example, if the colour correction of the lens is such that its resolving power is achieved over a bandwidth of 10 nanometres, then the potential information rate for each channel is in the region of 10^{13} bits per second and the total information rate may be 10^{20} bits per second. Although these formidable figures indicate the potential information handling capacity of commercially available lenses, it is not easy to use a lens in a digital optical system anywhere near its limit of information handling capacity.

3.2. A Matrix of Lenses

A matrix of lenses can make a matrix of images of a single object and such a matrix was used by Bell Telephone Laboratories in their optical store.²

A matrix of large complex lenses is very difficult to make since it is required that all the images should be in the same plane and this implies that all the lenses should have precisely the same focal length, a requirement which is very difficult to satisfy.

3.3. Lenticular Plate

If each lens in a matrix of lenses is made very small, then it may be very simple with a single curved refracting surface, and nevertheless can have a resolving power which is diffraction limited.³ This is because geometrical errors of a simple lens can be reduced simply by making the lens smaller, whereas the diffraction mechanism is independent of the size of the lens. If therefore each lens is made very small (of the order of 0.5 mm diameter) then the entire matrix of lenses can be moulded in a suitable transparent plastic to make a lenticular plate. It is not difficult to devise ways of using such a lenticular plate to increase the selection power of an optical store.

3.4. Optical Tunnel

A useful device related to the well-known kaleidoscope can be made using four mirrors assembled accurately to make an optical tunnel of square cross-section.⁴ An optical tunnel together with a single lens can make a matrix of images of a single object, that is, achieve much the same result as the matrix of lenses which was used in the Bell Telephone Laboratories' optical store. However, since only a single lens is involved, the problem of manufacturing many lenses with the same focal length is avoided. There is, however, a problem in ensuring that the reflectivity of the mirrors is high enough.

At first sight a tunnel using total internal reflexion would appear to solve this problem, but there are difficulties in such a construction in achieving

adequate mechanical stability for the tunnel and adequate optical uniformity for the tunnel material.

4. Optical Selection Device

A cathode-ray tube can be regarded as an optical selection device. The resolving power of a cathode-ray tube, defined as the number of resolved spot positions within its field of view, can easily be as high as 10^6 , but it is very difficult to exploit this high resolution on account of the problems involved in generating accurate electrical selection signals.⁵ It is possible to avoid this difficulty by the use of a servo system. In such a servo system some of the light from the cathode-ray tube is projected on to a coded scale which causes a series of electrical signals to be generated which correspond to the position of the spot. The spot position signals are compared with external selection signals to generate a correction signal which causes the demanded spot position, and the actual spot position, to coincide.

A servo-controlled c.r.t. was used in the Bell Telephone Laboratories' store⁶ and also in the ICL optical store.^{7,8}

4.1. C.R.T. with Compound Selection Device

A very elegant method of achieving both high resolving power and high accuracy in a cathode-ray tube has been demonstrated by Newberry.⁹ Newberry uses an electron lens analogue of the optical lenticular plate. An ordinary deflexion system causes the electron beam to be directed at one of the sub-lenses in the matrix, and a second deflexion system then causes the exit beam from the selected lens to be deflected by a small amount to the final spot position. Thus, for example, if each of the electrical signals has adequate precision to select one from 32 possible positions, then the whole device has a resolving power corresponding to 32^4 positions, i.e. 10^6 . This is all achieved without requiring that any electrical signal be more precise than about 1 in 200. Moreover, the selection process can be very fast.

4.2. Mechanical Deflexion Devices

It is quite easy to devise an electro-mechanical deflecting device similar to a mirror galvanometer. Moreover, it is very practicable to connect such a device in a servo system so that it can have high accuracy. Such a mechanical deflexion device is however inevitably slow by electronic standards, and it would be difficult to make a mechanical selector work in less than 1 ms.

4.3. Deflexion of Laser Beam by an Ultrasonic Stress Wave

A plane ultrasonic stress wave in a suitable transparent material causes small periodic variations in the optical properties of the material which can cause an

incident beam to be deflected by a process exactly analogous to that involved in X-ray reflexion by a crystal.¹⁰ The resolution of the exit beam evidently depends upon the number of waves in the reflecting region so that to achieve a high resolving power many waves are required which take a correspondingly long time to establish. It follows from this that an acoustic deflexion device takes a time proportional to the resolving power. Typically a few microseconds is required in practical cases. The exit beam energy, however, is a very large proportion of the input beam.

5. Storage Media

5.1. Silver Halide (Photography)

The photographic process is now so versatile and so well understood that many computer engineers have sought to exploit photography in a store. It is important to realize that the writing agent in photography is not the light but the chemical reducing agent used as a developer, since the physical change in the storage medium is initiated by light but effected by the chemical energy available from the developer. It is, therefore, inevitable that stores exploiting the photographic process must have poor writing facilities in comparison with those of the classes of store already in use in information processing systems, and this conclusion is quite independent of the way in which the photographic plate is used, e.g. to store information directly in the form of a pattern, or indirectly as a hologram.

5.2. Holographic Information Storage

To make a hologram requires a photographic plate with a very high resolving power. The reconstructed image, however, can be created by the use of any small part of the hologram so that we can consider that the information implicit in the reconstructed image is spread uniformly all over the hologram. In effect, therefore, the very high resolving power and high information storage capacity of the photographic plate is used to store a comparatively small amount of information in the reconstructed image by the use of a very redundant code in the process of making the hologram. It follows that a holographic store has the considerable advantage that a very perfect storage medium is not required, and the corresponding disadvantage that since the stored information is mixed up and spread all over the hologram, selective erasure or correction of the stored information is impossible.

5.3. Phase-modulation Photography

If a photographic plate is used to make a hologram then the energy in the reconstructed image is only a small proportion of the incident energy since the photographic emulsion works by absorbing some of the radiation. The energy loss can be avoided by

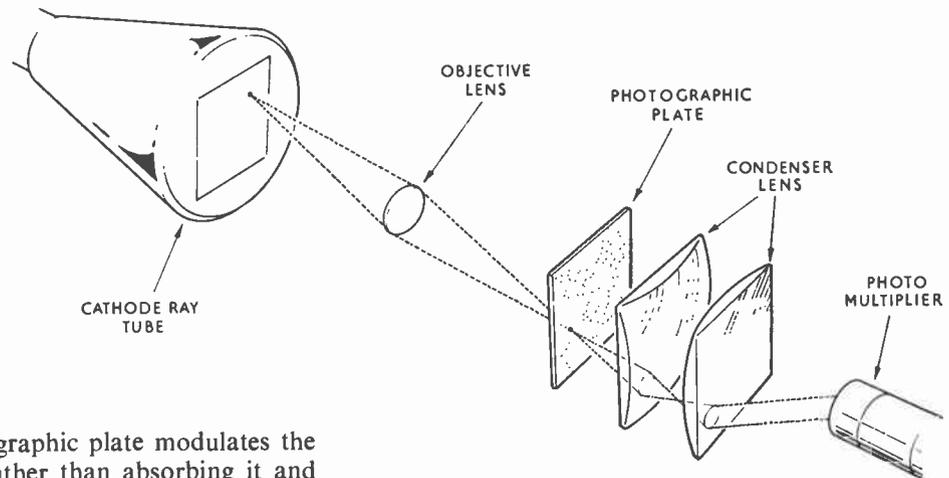


Fig. 1.
Bell Laboratory store
(After *Bell Syst. Tech. J.*,
38, No. 2, p. 404, March
1959).

arranging that the photographic plate modulates the phase of the radiation rather than absorbing it and various processes have been worked out for achieving this.

5.4. Photochromic Materials

A number of materials are known which after exposure to light become coloured. In order to detect whether the material has been coloured it is necessary to expose it to light of a wavelength which is absorbed by the coloured photochromic material and in most cases the interrogating light causes the coloured material to be bleached and revert to its normal state. Materials of this kind can in principle be used to make an erasable optical store, but there are very considerable practical difficulties since the light intensity required to change the photochromic material to its coloured state is very high. However, now that high intensity laser beams are available and methods are known for deflecting such beams, it is quite possible that a photochromic store may become practical in the next few years. It is perhaps worth noting that it may eventually be possible to design and make photochromic material which can be changed from one state to another by photons of energy E_1 and changed back again by photons of energy E_2 . The two states are distinguishable from one another by the fact that the material absorbs, or does not, photons of energy E_3 , but E_3 photons cannot affect the state of the material. The author knows of no material which satisfies these criteria, but is not aware of any law of physics which forbids its existence.

5.5. Curie Point Writing on Magnetic Materials

Certain materials such as manganese bismuthide, and gadolinium iron garnet are magnetic with a high coercivity and a low Curie point.¹¹ If, therefore, a very small region of material is heated above the Curie point by a focused laser beam, the magnetic state of the material after cooling may be changed. Moreover, the magnetic state may be detected by magneto-optic methods. These materials, therefore, provide

possible optical storage media which are at present being actively studied in many laboratories.

6. Proposals for Optical Stores

Most of the proposals which have already been published have been aimed at semi-fixed stores exploiting photographic plates as a storage media.

6.1. The Bell Laboratories Store (Fig. 1)

This was developed a decade ago for use in an experimental electronic telephone exchange. It exploited geometrical optical methods throughout, and incorporated a matrix of lenses to generate multiple images of the light spot from a cathode-ray tube. It is understood that it gave entirely satisfactory service, but was not used in production electronic telephone exchanges on account of the inconvenience of keeping the information up-to-date.

6.2. The ICL Optical Store (Fig. 2)

This is logically similar to the Bell Laboratories store, but differs in two details:

- (a) the lens matrix is replaced by an optical tunnel; and
- (b) the optical servo system used to locate the light spot accurately has a much larger optical aperture than that employed in the Bell Telephone Laboratories store in order to make the spot locating servo system operate as fast as possible.

6.3. Hologram Store (Fig. 3)

Anderson¹² has proposed a store in which the information is stored on a photographic plate in hologram form, and is read by an acoustically deflected laser beam. There appears to be every reason to expect that this type of store should work, since most of the techniques involved in it have been individually demonstrated to be viable.

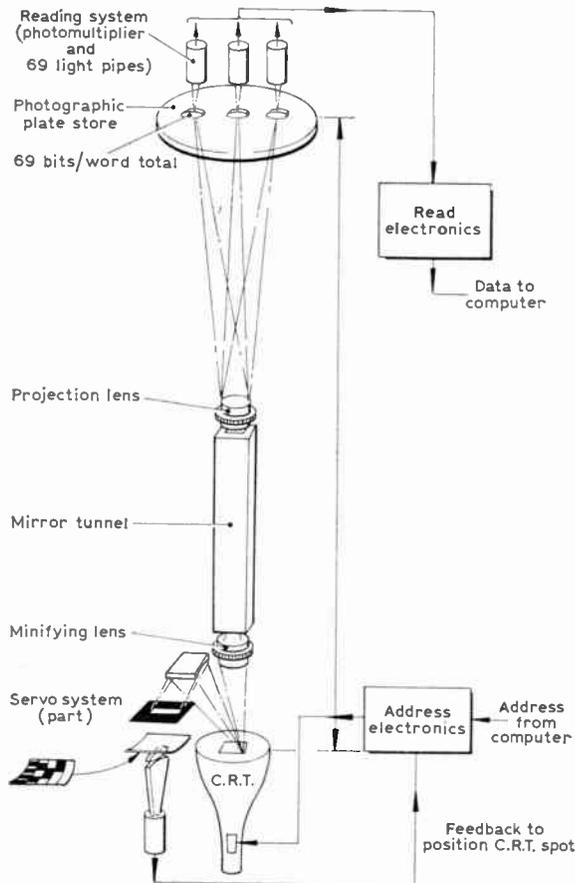


Fig. 2. ICL optical store.⁸

6.4. A Partly Variable Store (Fig. 4)

Stores have been proposed and built in which the information is represented by evaporated holes in a metallized surface on plastic tape. The holes are punched by a pulsed laser beam, and by this means it is possible to achieve about 1.5×10^5 bits per square centimetre. The store can be read at a very high information rate on account of the high information density on the tape. This class of store has the curious property that although holes in the tape cannot be deleted, extra holes can be added so that the information can be erased by a method similar to that involved in the use of an erase character on paper tape. It is, therefore, possible to use a store of this kind for information which is being changed fairly slowly, since out-of-date information can be erased and replaced by a new block at the end of the tape.

6.5. Mass File Store

A mass file store has been built¹³ in which the information is recorded on small pieces of photographic material which can then be mechanically selected from a very large collection. The access time for information in this mass file store is dominated by the mechanical access time to select the right chip of photographic material since the subsequent selection of the information on the chip by electro-optic methods is very fast.

7. Some Promising Possible Proposals

It is probably not very profitable at the present time to make new proposals for semi-fixed stores since work is already in hand on this class of store and the difficulties are as much associated with the problem of fitting semi-fixed stores into the information processing system as with the engineering

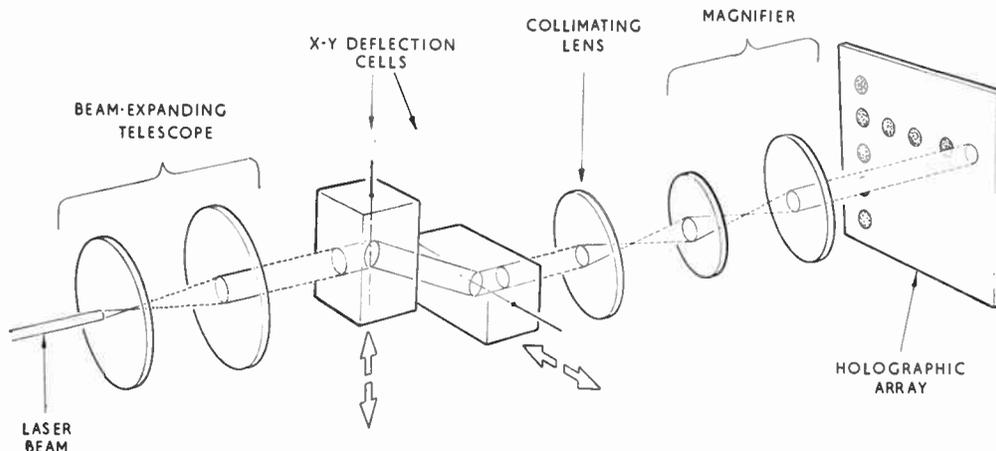


Fig. 3. IBM holographic store.¹²

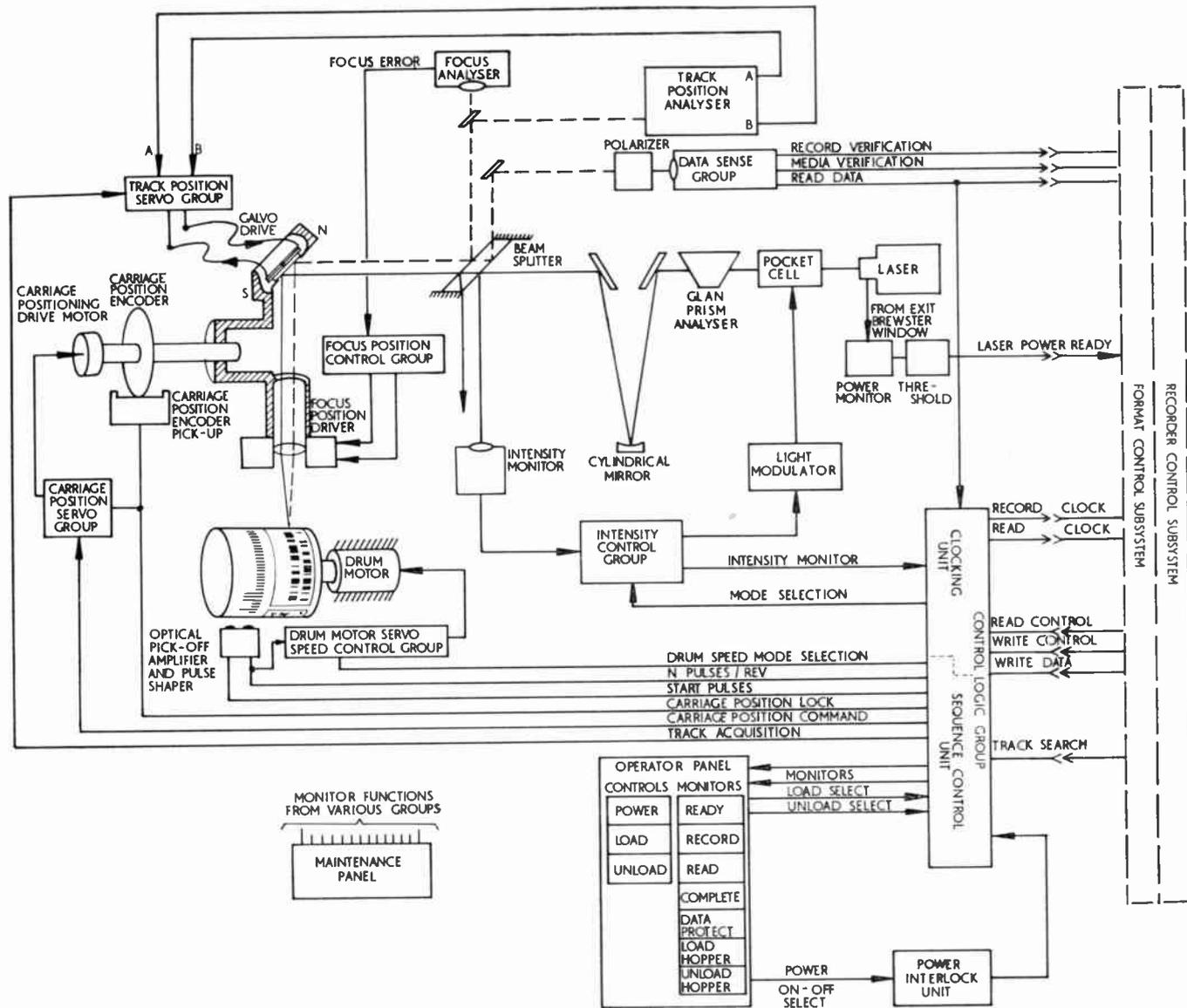


Fig. 4. Unicon store (Laser recording unit, Precision Alto, California). From *Proc. Fall Joint Computer Conference 1968*, 33, Part 2, p. 1373).

difficulties of making the store itself work. Accordingly a few proposals are mentioned here for stores with full read/write properties.

7.1. Magneto-optic Film Store

If we assume that one of the magneto-optic materials with Curie point writing will be used as the storage medium, then it seems unlikely that such a medium will be used with holographic techniques since it is in the nature of a hologram that selective re-writing of information is impossible.

It follows that the stored information will have to be represented directly on the film, and therefore a very perfect storage medium will be required. The development and manufacture of such storage media may very well be a crucial problem. If, however, this problem of making a very perfect medium is solved, it is quite easy to devise various combinations of electrical and mechanical selection which would lead to a store which is logically similar to magnetic disk stores as they are at present known, but with an access time perhaps ten times faster. There could be little doubt that such a store would be a most valuable addition to the system engineers' armoury.

7.2. Photochromic Store

It is not possible to make a plausible proposal for a photochromic store based on presently known photochromic materials since they suffer from fatigue effects such that after a large number of write/erase cycles the physical change required for writing becomes difficult to effect. However, with the deep understanding we now have of the mechanism of the photochromic effect in terms of energy levels, and the very powerful methods now available for growing crystals, it is quite conceivable that a program of materials research into photochromic materials could lead to a material which could be used as a reversible information storage medium.

At the same time work on the development of lasers is also making rapid progress, largely by a similar process of designing materials with appropriate energy level schemes so that an ever widening range of laser photon energy is becoming practicable.

One may, therefore, speculate that sooner or later, photochromic and laser materials will become available whose properties fit together in the way required to make an optically accessed reversible store.

It does not necessarily follow that such a store will have performance and cost parameters to justify its development, but since storage technology is of fundamental importance and the potential information density of an optically selected store may approach 10^7 bits/cm², i.e. 100 times the information density of the most ambitious magnetic store we can now

conceive, the case for undertaking a program of research into photochromic materials would seem to be worth considering.

8. Conclusions

Although optical stores are not at present used on a commercially significant scale, they offer two important potential advantages over magnetic storage.

(1) The information density on the storage medium (bits/cm²) can be higher by a factor of about 100.

(2) Various electro-optic and acoustic/optic techniques are available to achieve selection without moving parts or with few small moving parts.

At present these potential advantages have not been exploited so that the exploration of optical information storage techniques would appear to be likely to be a fruitful field for applied research.

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Of Current Interest

Changes to the British Post Office's Long-range Telephone Service

A reorganized long-range radiotelephone service for ships came into operation on 26th April, centred on Portishead, Somerset, and by-passing the Baldock, Herts., radio station. With the new system, the ship's operator calling a United Kingdom number from anywhere in the world may contact Portishead and, as soon as a radiotelephone frequency is available, he will then be connected into the inland network direct—not through the International telephone exchange as at present.

The shore subscriber calling a ship has previously made his call through the International exchange whence it went through Baldock radio station to the transmitter at Rugby and then to the radio operator aboard. The customer now asks his local telephone operator for the Ships' Telephone Service, Portishead Radio. The vessel will be called in special traffic lists, broadcast at two-hourly intervals (the time dictated by international radio regulations). This can mean faster connection with the ship.

Nearly 1,500 ships at present use the existing service. It is expected that with the new procedure traffic will increase rapidly. The long-range radiotelephone service is available to ships which are equipped to receive and transmit on this system in any part of the world. In addition, the Short-range Radiotelephone Service is used by shipping within 300 miles of the 11 coastal radio stations around Britain for ship-to-shore communication.

High-speed Scanning Electron Microscopy

Recent developments at the National Physical Laboratory, Teddington, have extended the use of the Scanning Electron Microscope, making it possible to observe dynamic phenomena at high magnification. It is now possible at the Laboratory to observe continuously the changes taking place in materials subjected to stress. Carbon fibre composites are among the materials to have been observed in this way. A straining stage has been specially developed for this purpose to fit the 'Stereoscan' microscope made by Cambridge Scientific Instruments Ltd. Using a conventional video tape recorder, it is possible to record the results of such experiments on tape and to re-examine subsequently a single frame or a sequence of frames.

In the scanning electron microscope an electron beam is scanned over the surface of the specimen in synchronism with a cathode-ray tube. Electrons leaving the specimen are collected and the resultant current is amplified and used to control the brightness of the tube. Since the number of electrons leaving the specimen is dependent on its topography, an image of the surface is displayed on the tube. One of the most striking features of the scanning electron microscope is its great depth of focus and because of this it produces images having a realistic three-dimensional appearance.

Hitherto, the electron image display has had to be built up slowly, like a radar display, on a long persistence screen. In the new N.P.L. high-speed scanning system a bright, flicker-free image is produced on a television monitor

tube. The advance has been made possible by improvements in the electron detection system and in the performance of the scanning amplifiers. These improvements can be added without modification to the basic instrument. The instrument has been used at magnifications up to 20 000 times, corresponding to a scanned area of about 5 micrometres square.

Implants of Nuclear-powered Heart Pacemakers

Heart pacemakers powered by a nuclear battery developed at the Atomic Energy Research Establishment at Harwell have successfully been implanted in patients at the National Heart Hospital in London. These operations follow the successful implantations of this type of pacemaker in dogs reported earlier this year. The new pacemaker is designed to maintain the heart-beat of patients suffering from 'heart-block' for at least 10 years before the need for a fresh implant arises. This compares with the period of about two years between implant operations which is all patients using conventional, chemically-powered pacemakers now have. About 1,300 conventional pacemakers are implanted in new patients in Great Britain every year.

The nuclear battery, which was developed at Harwell, utilizes the heat from the radioactive decay of a small quantity of plutonium-238 to generate electricity from a miniature semiconductor thermo-electric converter. The complete battery is 5 cm long and about 1.25 cm across (2 in \times $\frac{1}{2}$ in). It weighs about 28 g (1 oz) and is designed to last for a minimum of 10 years. There is no radiation hazard to the patient, or to anyone else, from the small quantity of plutonium used and the battery is fully encapsulated to prevent the escape of radioactive material or attack from body fluids. The capsule containing the plutonium is immensely strong, and has been designed to withstand the worst kind of foreseeable accidents that could happen to it, including intense heat and a crushing force of two tons. The pacemakers used in the trials are special units coupled to the Harwell battery through a voltage-changing circuit developed at Aldermaston.

The nuclear battery was developed by A.E.R.E. in close collaboration with the Institute of Cardiology, the National Heart Hospital, A.W.R.E., Aldermaston, Devices Implants Ltd., and the Department of Health who are giving substantial financial support. Active discussions are going on with a British firm about the manufacture and marketing of the battery at home and overseas.

Heart pacemakers have been used for over 10 years to maintain the heartbeat of patients suffering from 'heart-block'.† This disease, the failure of a bundle of nerves in the heart (the Bundle of His), can be overcome by using a pacemaker to provide the minute rhythmic electrical pulses normally transmitted through the nerve bundle. The devices are powered by small chemical batteries which need replacement at approximately two-year intervals which means repeated surgery in hospital for the patient.

† See, for instance, Davies, J. G., 'Cardiac pacemakers', *J. Brit. I.R.E.*, 24, p. 453, December 1962.

An Oscilloscope Polar Coordinate Display for Multi-dimensional Data

By

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The paper describes in detail a simple electronic system which processes up to eight simultaneous analogue signals and displays them in the form of a polar coordinate diagram on a commercial oscilloscope with identical X and Y amplifiers. Some preliminary subjective tests are briefly described, which confirm that the human observer can quickly detect several changes in the display and classify it on the basis of particular features or overall shape.

1. Introduction

Greatly improved techniques in transducing, electronic processing and automation have made it both possible and necessary to monitor an increasing number of parameters in many different types of systems. Whether the system consists of heavy industrial plant or the human body, for instance, the increasing amount of available data while apparently of great value, can also be an embarrassment because it may not be absolutely clear what to do with it.

In particular, difficulties arise in presenting the information to human operators, who may need specific numerical data or just an overall picture of the state of the system, or a combination of both. The numerical information can be provided with a wide range of instruments and displays, from meters, pen recorders and number tubes to computer print-out and graphic terminals.

However, the other form—the qualitative or ‘overall picture’ information—is more difficult to provide, increasingly so as the number of parameters increases. The operator often has to interpret the state of a system from a collection of instruments or a list of numbers. Although with practice the human operator can become very proficient at this, it ought to be possible to provide an aid to quick appraisal of the situation.

The need seems to be, therefore, for a method of displaying many parameters simultaneously in a form in which they can easily be related to one another and interpreted as a whole. One suggested solution to this problem is the polar coordinate type of display which is widely thought to have much to commend it.^{1,2}

In principle the display is simple. A number of parameters, eight say, may be represented on a two-dimensional medium by plotting their amplitudes outwards along radials from a common zero position.

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In the case of eight parameters this is most sensibly achieved using the rectangular axes and diagonals as shown in Fig. 1. The plotted points may be joined up to produce a closed shape, as shown. This diagram is not the only solution to the problem, but the fact that it forms a closed shape is a very important feature. The ‘bar-chart’ or ‘histogram’ types of display also represent many parameters in two dimensions, but they do not have ‘shape’ in the same sense as the polar coordinate display. This feature is important because the human being is good at interpreting and remembering shapes, albeit this is a completely subconscious process. We can rapidly sum-up a scene; we can recognize a face after 20 years.

On the polar diagram, the parameters need not be plotted directly, but scaled so that under normal conditions the points lie on a ‘normal circle’—represented on the eight-point diagram as an octagon. Abnormalities now show as deviations from the normal circle or octagon. The diagram is not limited to eight variables, of course, but in the case of an electronic display the processing must necessarily become more complicated as the number of parameters increases. In any particular application, experience might indicate how best to distribute the variables around the diagram.

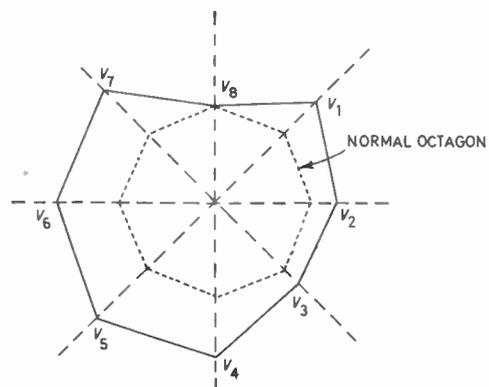


Fig. 1. The eight-parameter polar coordinate diagram.

It has been suggested¹ that this type of display might be very useful in the representation of clinical data where there is the need to make a judgement on the total state of a patient or on the total effectiveness of treatment. In engineering systems where the parameters are often interdependent, it may be very useful to see the *total* result of a single adjustment. It has also been suggested that the display would greatly facilitate the examination of multiple sets of data in order to select the parameters that undergo the largest variations over all the sets.

To study this type of display an electronic system has been designed and constructed, which accepts up to eight simultaneous positive d.c. levels, samples them cyclically and displays the polar coordinate diagram on a commercial oscilloscope with identical *X* and *Y* amplifiers.

2. The Electronic System

The electronic plotting of the eight amplitudes is not difficult provided that each of the input channels is assigned to a particular radial of the display (see Fig. 1). Clearly the parameters assigned to the rectangular axes (inputs 2, 4, 6, 8) can simply be applied to the oscilloscope *X* or *Y* inputs without modification, except that inputs 4 and 6 must be inverted. From Fig. 1 it can be seen that the diagonal points are all obtained by multiplying the appropriate input amplitude (*V*) by the sine or cosine of 45 degrees. The points are given as follows:

- 1 $X = Y = +0.71V_1$
- 3 $X = +0.71V_3$ $Y = -0.71V_3$
- 5 $X = -0.71V_5$ $Y = -0.71V_5$
- 7 $X = -0.71V_7$ $Y = +0.71V_7$

The electronic processing must therefore provide the inputs 1, 3, 5 and 7 multiplied by 0.71, with 3, 5 and 7 also inverted.

The various functions have been achieved by combining a number of standard electronic circuits and techniques; the block schematic is shown in Fig. 2. There can be considerable variation in the form of

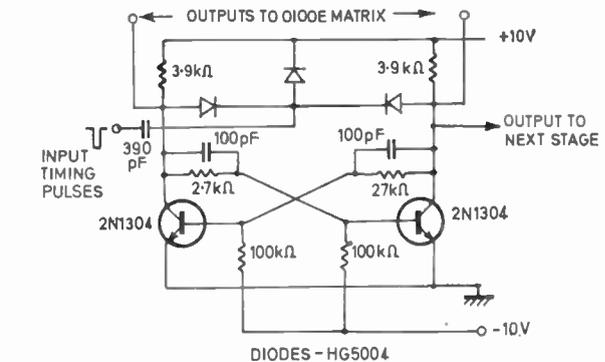


Fig. 3. The first bistable circuit of the cyclic control.

the system, depending ultimately on the number of input channels, and other requirements and restraints.

The processing of the input signals is controlled by a cyclic sequence of eight sample pulses derived from the cyclic control circuits (Fig. 2) and used to open the input and output gates in the correct order. It was found that the best input p.r.f. is approximately 7 kHz, since at this frequency the display has uniform brightness. At lower frequencies the plotted points become brighter, at the expense of the lines joining them.

The cyclic control consists simply of three collector-triggered bistable circuits (the first is shown in Fig. 3) connected in the normal way to count up to eight. The outputs of the bistables are decoded with a diode matrix giving the sequence of eight sample pulses. The width of these pulses is, of course, the same as that of the output pulse of the first bistable circuit.

To reduce the loading on the diode matrix, each sample pulse is first used to switch an emitter follower buffer stage using the 2N1304 transistor which has a current gain of about 50. As shown in Fig. 4 the emitter follower output controls a transistor switch in series with the input. A silicon transistor is used here because its relatively high V_{BE} will provide some immunity from spurious signals in the absence of the

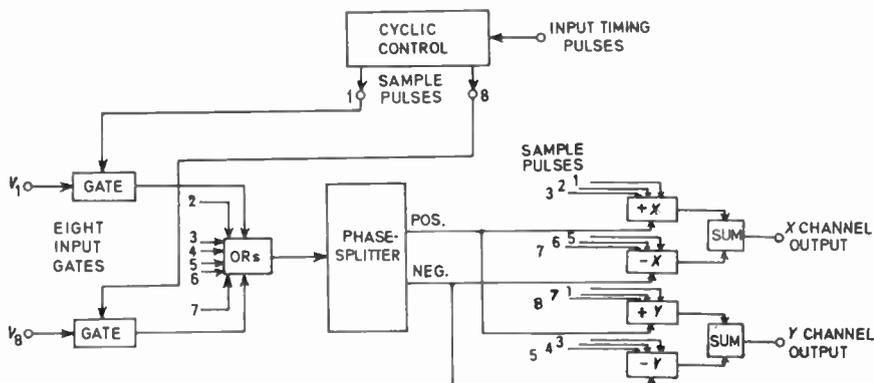
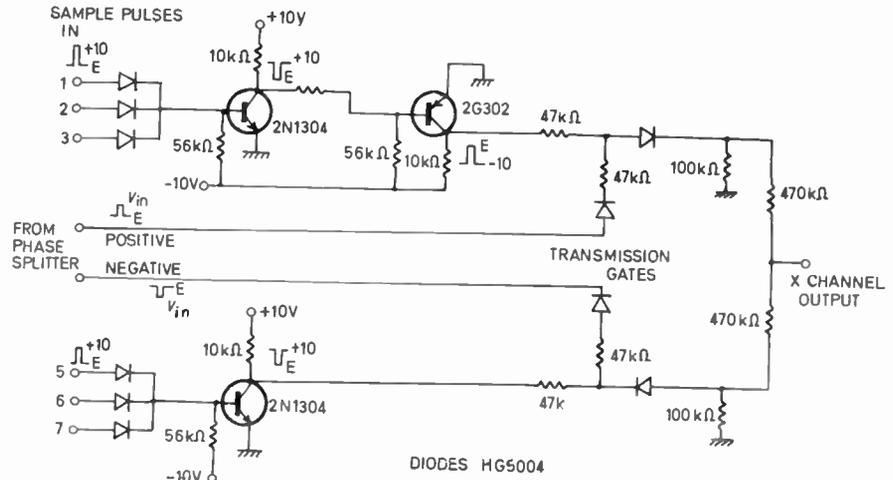


Fig. 2. Block schematic of the electronic system.

Fig. 5. The output gates of the X channel.



sample pulse. The eight series switches are connected in two groups through diode ORs to the phase-splitter. One group consisting of the inputs 1, 3, 5 and 7 is first attenuated by a potential divider to give $0.71V_{in}$. The remainder are connected directly to the next stage. The phase-splitter consists of a 2G302 transistor—with low $V_{CE(sat)}$ —supplied from +10V and -10V, to give positive and negative outputs at the emitter and collector respectively.

It can be seen from Fig. 2 that the X and Y outputs must each consist of positive and negative components which are controlled by the appropriate sample pulses. Figure 5 shows the arrangement for the X channel, in which the positive and negative gates are opened by pulses 1, 2, 3 and 5, 6, 7 respectively. Correspondingly for the Y channel the positive gate is opened by pulses 1, 7 and 8, and the negative one by pulses 3, 4 and 5.

The waveform obtained at the X-output for the normal octagon, when there is no oscilloscope connected is shown by the solid line in Fig. 6. The Y-output is identical in shape but displaced by three intervals. When the X and Y outputs are connected to an oscilloscope the amplifier input capacitances

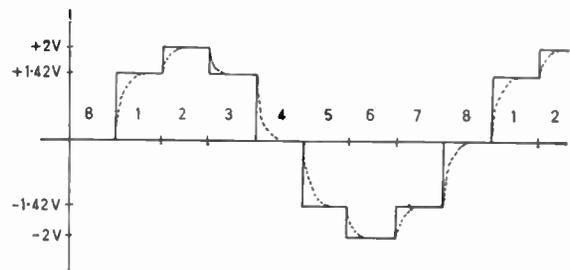


Fig. 6. Output waveform of the X channel.

form low-pass filters with the output impedances of the X and Y channels and the waveform is rounded, as shown by the dotted lines in Fig. 6. However, this is not unwanted distortion but the basis for the straight lines connecting the points of the diagram. Tektronix and Solartron oscilloscopes have been used

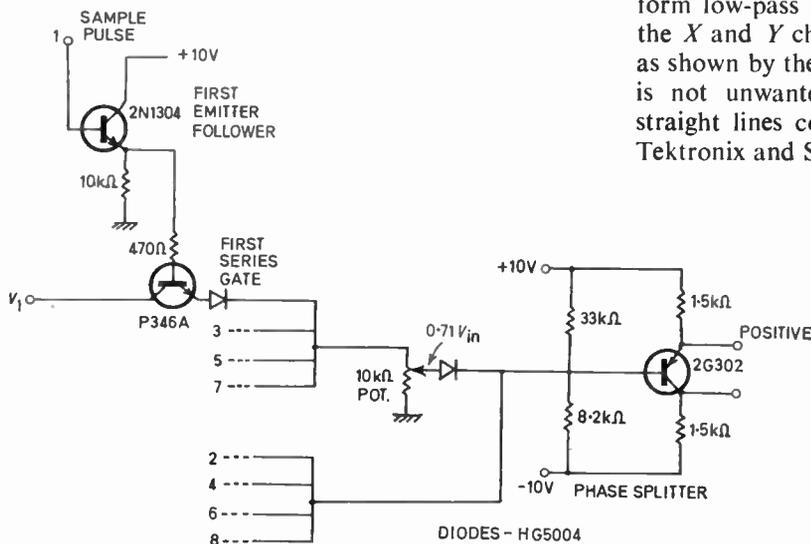


Fig. 4. The input gates, diode ORs and phase splitter.

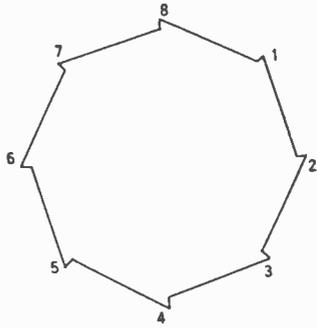


Fig. 7. The display with insufficient load capacitance.

with the present system and their amplifiers had input capacitances of 47pF (Tektronix Type H) and 35pF (Solartron CX1441). An input capacitance of less than this value does not hold the output and the display has the form shown in Fig. 7. Capacitance must be added to the channel outputs to correct this, so some 'tailoring' to the oscilloscope may be required, but experience here indicates that this will not be difficult.

The polar display obtained with the present system is shown in Fig. 8. The brightness is variable of course, and the dots can be accentuated by decreasing the sampling rate. The light 'flecks' adjacent to each point are reflexions in the circular graticule placed over the c.r.t. face.

Further development in progress is to increase the number of parameters displayed to sixteen using a somewhat different process and to provide long- and short-term memory so that at any time the display

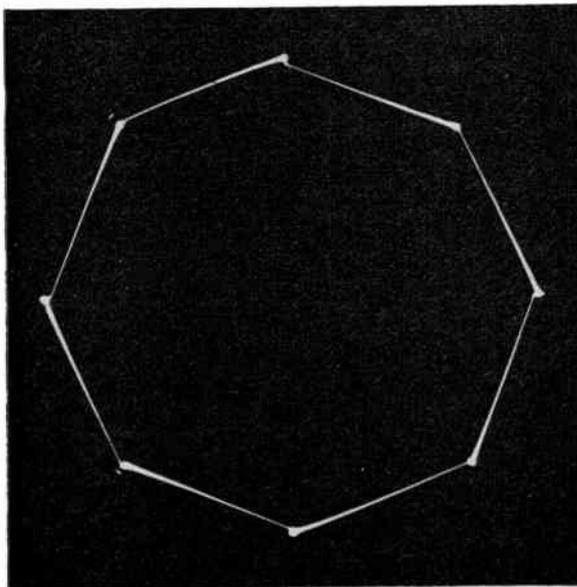


Fig. 8. The actual polar display obtained.

shows the present values compared with those of a known period earlier. The processing does not have to be electronic, especially for very slowly changing parameters. Servo systems can be used³ and have the advantage that it is comparatively simple to introduce different, and possibly non-linear, laws into particular parameter channels.

3. Some Preliminary Subjective Tests

To determine the potential of this type of display it was decided to investigate the ability of human observers (a) to detect slight but simultaneous changes in two or more parameters, (b) to ascertain the category of the complete diagram, and (c) to detect particular features. Only five observers were tested but the pattern of results was clearly established. After a few minutes of explanation each observer was given four tests in which the 'normal octagon' of Fig. 1 was always shown first. The *actual* c.r.t. display, illustrated in Fig. 8 was used throughout these tests.

Test 1. A series of ten diagrams was shown in which there was either no change at all from the normal octagon, or there was an increase of 10% in one parameter *and* a decrease of 10% in another. In this series only, the observers were told when they were correct.

Although this was only the first trial, the subjects detected 85% of the changes correctly and only took 7 to 10 seconds per diagram. On a second series the same accuracy was achieved at the increased rate of one diagram every 7 or 8 seconds. Better than 80% was obtained at times of 5 to 7 seconds. One observer found every change at 4.5 seconds per diagram; another two found 88% at 5 seconds. It must be emphasized that the normal octagon was not shown between test diagrams.

Test 2. A series of 10 complex figures was shown; examples are given in Fig. 9. The subjects were asked to detect whenever *both* parameters 1 and 2 changed (increase or decrease) *or*, *both* 4 and 6 changed. Note that one pair are adjacent to one another, the others are not. All the observers got all the answers correct at only 3.7 to 5 seconds per diagram which is quite fast considering the amount of work to be done in the time.



Fig. 9. Examples of the complex diagrams used in Test 2.

Test 3. Another series of complex figures was presented and the subjects were asked to describe the area in each quadrant, particularly which were accentuated and which depressed, compared with the norm. Everyone classified the diagrams accurately and with great detail at a speed limited only by the time to speak the answer.

Test 4. In a series of ten diagrams the observers were asked to detect the occurrence of one or more of a set of features shown on a card next to the display. The features were, for example, a horizontal line between points 4 and 5 or 3 and 4; a right-angle formed by the lines joining 6, 7 and 8. The observers found 97% of the features at 8 seconds per diagram.

The observers' task of detecting deviations from the 'norm' was considerably more difficult than would be the case in practice, where it would be quite simple to provide a reference diagram on a transparent overlay on the c.r.t. face.⁴

4. Conclusions

The need for the simultaneous display of many parameters has been explained and the polar co-

ordinate type of display advanced as a possible solution. An electronic system for the display of eight parameters has been described in detail and shown to be possible using established and cheap circuit techniques. The results of preliminary subjective tests demonstrate that the diagram can be used efficiently by observers and suggest that the display may have wide application.

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2. Coekin, J. A., 'A versatile presentation of parameters for rapid recognition of total state', Proceedings of the I.E.E.E. International Symposium on Man-Machine Systems, Cambridge, 8th-12th September 1969.
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STANDARD FREQUENCY TRANSMISSIONS—July 1970

(Communication from the National Physical Laboratory)

July 1970	Deviation from nominal frequency in parts in 10 ¹¹ (24-hour mean centred on 0300 UT)			Relative phase readings in microseconds N.P.L.—Station (Readings at 1500 UT)		July 1970	Deviation from nominal frequency in parts in 10 ¹¹ (24-hour mean centred on 0300 UT)			Relative phase readings in microseconds N.P.L.—Station (Readings at 1500 UT)	
	GBR 16 kHz	MSF 60 kHz	Droitwich 200 kHz	*GBR 16 kHz	†MSF 60 kHz		GBR 16 kHz	MSF 60 kHz	Droitwich 200 kHz	*GBR 16 kHz	†MSF 60 kHz
1	-300.1	-0.2	0	615	595.0	17	-300.0	-0.2	+0.1	616	605.0
2	-300.1	-0.1	0	616	596.0	18	-300.1	-0.1	+0.1	617	606.0
3	-300.1	-0.1	0	617	596.6	19	-300.1	-0.1	0	618	607.4
4	-300.0	-0.1	0	617	597.6	20	-300.2	-0.2	+0.1	620	609.4
5	-300.0	-0.1	0	617	598.2	21	-300.1	-0.2	+0.1	621	611.0
6	-300.0	-0.2	+0.1	617	600.0	22	-300.0	-0.1	0	621	611.9
7	-300.0	-0.1	+0.1	617	601.0	23	-300.0	-0.1	+0.1	621	612.4
8	-300.0	+0.3	+0.1	617	597.8	24	-300.1	-0.1	0	622	613.4
9	-300.0	-0.1	+0.1	617	599.2	25	-300.0	0	+0.1	622	613.4
10	-299.8	0	+0.1	615	598.8	26	-300.1	0	0	623	613.8
11	-299.9	0	+0.1	614	599.0	27	-300.1	-0.1	0	624	614.6
12	-300.0	0	+0.1	614	599.4	28	-299.9	0	+0.1	623	614.8
13	-300.0	-0.1	+0.1	614	600.2	29	-299.9	0	+0.1	622	615.0
14	-300.0	-0.1	+0.1	614	601.6	30	-300.0	0	+0.1	622	615.2
15	-300.1	-0.1	0	615	602.2	31	-300.0	-0.1	+0.1	622	616.0
16	-300.1	-0.1	+0.1	616	603.2						

All measurements in terms of H.P. Caesium Standard No. 334, which agrees with the N.P.L. Caesium Standard to 1 part in 10¹¹.

* Relative to UTC Scale; (UTC_{NPL} - Station) = + 500 at 1500 UT 31st December 1968.

† Relative to AT Scale; (AT_{NPL} - Station) = + 468.6 at 1500 UT 31st December 1968.

Noise Considerations in the Design of Current Sources

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and

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The inclusion of a current source in the input stage of an amplifier may seriously degrade the overall noise figure. Some commonly used input circuits incorporating bipolar transistors and j.f.e.t.s. are analysed, and simple criteria are developed for low-noise design of the current source.

The d.c. current source is widely used in modern circuit designs. In it, a transistor-resistor combination is used not as an amplifier but to simulate one terminal of a high value resistor (of the order of 1 MΩ) with a high d.c. voltage (of the order of 1 kV) applied to its remote terminal. The most familiar application of the current source is in the balanced input circuit illustrated in Fig. 1(a).

In the Figure the generator v_1 in conjunction with resistance R_1 represents the Thévenin equivalent of a low value (of the order of 1 V) d.c. voltage source which customarily consists of a resistive potential divider, a series of forward-biased p-n junctions, a reverse-biased Zener diode or an R-C combination; in all cases energization is from the supply rails. The base of transistor Q_C is thus maintained at a constant potential and Q_C supplies constant current to the signal transistors Q_A and Q_B .

A number of cases have been brought to our attention where a current source in the input stage causes severe degradation of the noise figure of the amplifier. We shall derive here some simple criteria for the low-noise design of such sources.

In Fig. 1(b) we have redrawn Fig. 1(a) to include the important noise generators. These are the current and voltage noise generators in each of the three transistors and the noise generators of the two resistances R_1 and R_2 . We shall carry out a low-frequency analysis on the assumption that all the devices have a high d.c. gain ($\beta_0 > 100$) and that R_1 and R_2 have been reasonably chosen in relation to the d.c. stability requirements, implying that

$$R_2 \gg r_{eC}, \quad R_1 \ll \beta_0 R_2 \quad \dots\dots(1)$$

where r_{eC} is the reciprocal of the transconductance of transistor Q_C at the given operating current.

If we assume that the operating current is sufficiently low (say $< 300 \mu\text{A}$) then the effect of base resistance

will be negligible and we may write the following expressions for the mean-square values of the transistor noise current and noise voltage generators.‡

$$\overline{i_{NC}^2} = \frac{N(1 + \omega_F/\omega)}{2\beta_0 r_{eC}}, \quad \overline{v_{NC}^2} = Nr_{eC}/2 \quad \dots\dots(2)$$

where N is used as an abbreviation for $4kT\Delta f$. We have included a flicker-noise factor with the current generator, its effect being formally equivalent to a decrease in β_0 as the frequency falls below the flicker-noise characteristic frequency ω_F . The other two noise generators have mean-square values given by the Johnson noise expressions,

$$\overline{v_{N1}^2} = NR_1, \quad \overline{v_{N2}^2} = NR_2. \quad \dots\dots(3)$$

Taking account of the conditions in (1) and carrying out a straight-forward circuit calculation from Fig. 1(b) we obtain the following result for the mean-square value of the noise current in the collector lead:

$$\overline{i_N^2} = N \left[\frac{r_{eC}}{2R_2^2} + \frac{1}{R_2} + \frac{R_1}{R_2^2} + \frac{(1 + \omega_F/\omega)(R_1 + R_2)^2}{2r_{eC}\beta_0 R_2^2} \right]. \quad \dots\dots(4)$$

In deriving equation (4) we have made the simplifying assumption that the small-signal a.c. gain h_{fe} is equal to β_0 , which is very well justified in the case of modern low-level bipolar transistors.

The effect of the noise current from the current source on the overall amplifier noise figure will depend upon the configuration of the second stage. In some circuit designs, particularly those for the low-level d.c. operation, the second stage is differential and is not affected by the current source noise which appears as a common-mode voltage if the two transistors Q_A and Q_B are well matched. However in a wide range of applications it is preferable to connect the second stage to the collector of only one of the signal input transistors Q_A or Q_B . In this situation it is necessary to calculate the contribution made by

‡ Faulkner, E. A., 'The design of low-noise audio-frequency amplifiers', *The Radio and Electronic Engineer*, 36, pp. 17-30, July 1968.

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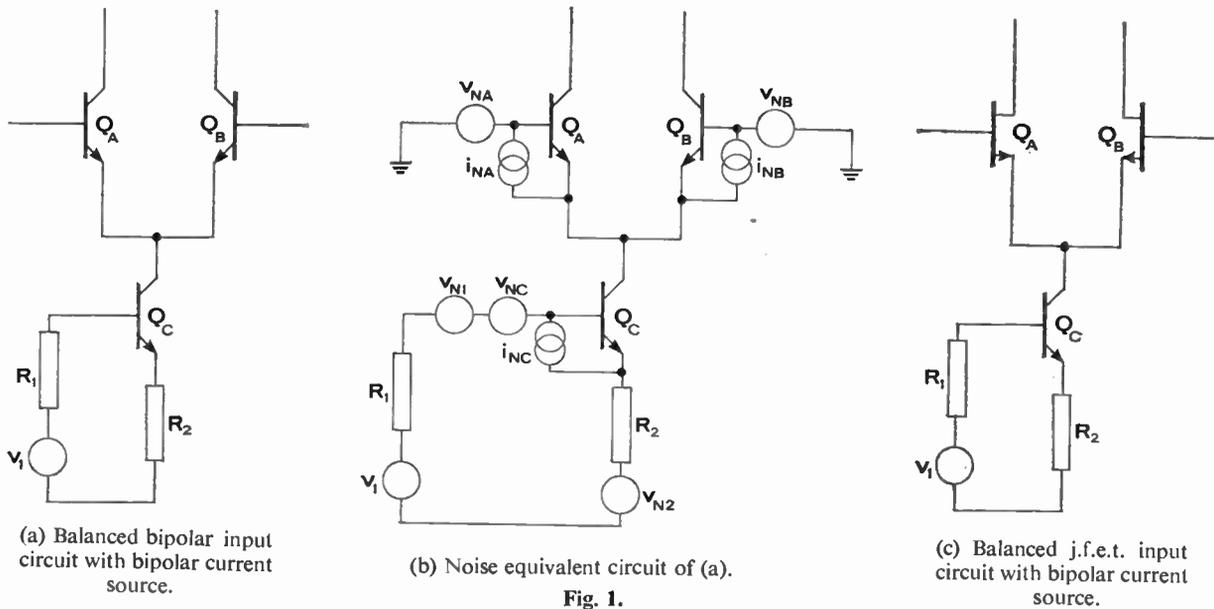


Fig. 1.

the current source to the noise current at the collectors of \$Q_A\$ and \$Q_B\$ and now the criterion for low noise design of the current source is that this contribution should be negligible compared with the contribution of transistors \$Q_A\$ and \$Q_B\$ themselves. The most stringent operating condition is that in which the least noise is contributed by \$Q_A\$ and \$Q_B\$, that is when the impedances between base and earth are zero.

We shall assume that the current from \$Q_C\$ divides equally between the signal input transistors \$Q_A\$ and \$Q_B\$. The reciprocals of the transconductances are then \$r_{eA} = r_{eB} = 2r_{eC}\$, and the equivalent noise generators of \$Q_A\$ and \$Q_B\$ will be equal in magnitude but of course uncorrelated. Solving the circuit of Fig. 1(b) we find that transistors \$Q_A\$ and \$Q_B\$ make the contribution

$$\frac{N}{4r_{eA}} \left[1 + \frac{(1 + \omega_F/\omega)}{\beta_0} \right]$$

to the mean-square noise current in the collector of \$Q_A\$. The mean-square contribution of the current source noise is one quarter of the value given by equation (4). The condition for this contribution to be negligible can now be written as follows,

$$\left[1 + \frac{(1 + \omega_F/\omega)}{\beta_0} \right] \geq 2 \left[\frac{r_{eC}^2}{2R_2^2} + \frac{r_{eC}}{R_2} + \frac{R_1 r_{eC}}{R_2^2} + \frac{(1 + \omega_F/\omega)(R_1 + R_2)^2}{2\beta_0 R_2^2} \right] \dots\dots(5)$$

The parameters \$\beta_0\$ and \$\omega_F\$ are assumed to be the same for all three transistors.

The first two terms on the right-hand side of equation (5), which arise from the voltage generator noise in \$Q_C\$ and the Johnson noise in \$R_2\$ respectively, are seen to be negligibly small because of the initial assumption that \$R_2 \gg r_{eC}\$. The third term, arising from Johnson noise in \$R_1\$, is likely to be negligible in any normal design where because of the d.c. stability requirements we may well have \$R_2 = 50r_{eC}\$, \$R_1 < 10R_2\$. It is the fourth term which proves to be the most troublesome in practice, the contribution arising from current noise (including flicker-noise) in \$Q_C\$.

Considering first the frequency range where flicker-noise is negligible, we have \$\omega \gg \omega_F\$ and find the requirement that \$R_1\$ must not be more than a few times greater than \$R_2\$. In the flicker-noise region the best possible results are obtained when \$R_1 \ll R_2\$, but for frequencies well below \$\omega_F\$ the current source will always make a flicker-noise contribution equal to that of \$Q_A\$. In this region a differential second stage is necessary if the best results are to be obtained.

It is not normally practicable to establish the condition \$R_1 < R_2\$ by using a potential divider network carrying a heavy current. The most practicable solution is to use two or more forward biased p-n diodes to provide \$v_1\$; if the diode chain carries a current equal to the current in \$Q_C\$ its resistance will be a few times \$r_{eC}\$ and the required conditions will be satisfied. The shot-noise generated by this diode combination will actually be less than the thermal noise in a resistor with the same impedance, so its contribution to the total noise will be negligible. Note that Zener diodes are normally much too noisy for this application.

Although an R-C combination is sometimes used as a means of reducing the impedance at the base of Q_C there seems, generally speaking, little to justify this method as a means of improving the performance in the flicker-noise region because of the bulk and cost of the capacitor that would be required and because of the inconveniently long time-constant involved.

Let us now consider the input circuit of Fig. 1(c) which is the same as 1(a) except that the input transistors are now junction f.e.t.s rather than bipolars. The principal differences which arise in this case compared with 1(a) can be summarized as follows:

(a) The transconductance of a j.f.e.t. at a typical operating current, say 1 mA, is less than one-tenth of the transconductance of a bipolar device operated at the same current.

(b) Flicker-noise effects in the j.f.e.t. appear in the noise voltage generator rather than the current generator, which is small enough to be negligible for our present purposes. In the absence of flicker-noise effects the mean-square value of the noise voltage generator is approximately $\frac{1}{2}Nr_s$ where r_s is the reciprocal of the transconductance; this value is approximately the same as a bipolar operating with the same transconductance. The mean-square noise contribution from transistors Q_A and Q_B is now

$$\frac{N}{4r_s} [1 + \omega_{FA}/\omega].$$

Now we may set up an equation analogous to (5) by making some assumption about the ratio between the transconductance of the bipolar and the j.f.e.t. A typical value for this ratio may be 20 and we obtain the condition,

$$(1 + \omega_{FA}/\omega) \gg 20 \left[\frac{r_{eC}^2}{2R_2^2} + \frac{r_{eC}}{R_2} + \frac{R_1 r_{eC}}{R_2^2} + \frac{(1 + \omega_{FC}/\omega)(R_1 + R_2)^2}{2\beta_0 R_2^2} \right]. \quad \dots\dots(6)$$

The first point which arises when (6) is compared with (5) is that the use of a j.f.e.t. places much more severe restrictions on the value of R_2 which must now have a value of at least $100r_{eC}$ if the Johnson noise in R_2 is not to have a detrimental effect. In practice this means a drop of at least 2.5 V across R_2 , and, when this condition is satisfied, the contribution from the third term will also be negligible if $R_1 < R_2$.

We now come to the fourth term on the right-hand side of (6) which represents the flicker-noise contribution of the current source. Putting $R_1 < R_2$, we find to a good approximation the condition

$$(1 + \omega_{FA}/\omega) \gg \frac{10(1 + \omega_{FC}/\omega)}{\beta_0}. \quad \dots\dots(7)$$

For $\beta_0 > 100$ this condition will be satisfied provided that $\omega_{FC} \leq \omega_{FA}$, which is the requirement that the flicker-noise characteristic frequency of the bipolar does not exceed that of the j.f.e.t.

We see that the design requirements in the flicker-noise regime are more easily met with a j.f.e.t. input stage, resulting from the fact that in this regime a bipolar is less noisy than a j.f.e.t. when the signal source is of low impedance.

The approach outlined above can be applied to other basic circuit configurations by suitable modification of the noise equivalent circuit (Fig. 1 (b)). In particular we find that unity-gain circuits incorporating a bipolar or j.f.e.t. signal transistor and a bipolar current source are governed by design criteria (5) and (6) as before.

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