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CONTRIBUTIONS TO THE JOURNAL

ONE of the ways in which the Institution fulfils its object of "promoting the advancement of radio and electronics and kindred subjects by the exchange of information" is through the publication of the *Journal*. Other ways are, of course, through the holding of meetings, but many of the subjects discussed as papers or reports are eventually published in the *Journal*.

Papers published in the *Journal* can be classified under several main groups. First and foremost is the important original contribution, which may be a description of a new circuit or component development or the presentation of an original piece of theoretical investigation.

The second group consists of review or survey papers; these contributions, usually written by an engineer with long experience in the particular field, fulfil a particularly useful purpose in bringing together information on the widest aspects of a problem.

The third category of papers was discussed in an article in the April 1955 *Journal* which particularly described an "Engineering Development" paper. This type of paper examines a complete equipment or project, demonstrates the application of familiar engineering principles and shows the way in which "know how" enters into the engineer's work.

The preparation of a technical paper for publication in the journal of a professional Institution is never an easy task, and calls for an appreciable amount of time and effort. Accordingly much work which would be of interest and of potential value to other engineers is seldom available outside the restricted field of an internal report: engineering knowledge is thereby the poorer. Such gaps in technical literature can well lead to repetition of work which has already been carried out.

With the object of helping the engineer who does not have time to prepare a full paper, the Council wishes to make more widely known that short papers are accepted for publication in the *Journal*. These may be of any length, from two or three hundred words up to one or two thousand words, and the subject may cover almost any aspect of radio and electronic engineering. An obvious example of material suitable for presentation as a short paper is the development of new circuits which, while not a major advance on existing practice, does represent useful ideas for other engineers. There will also be interest in the description of new types of instruments, with special reference to any novel development, and brief notes on investigations into such important aspects as propagation, reliability and subjective testing, to name only three. A particularly useful type of short paper would be a description of production techniques which have been evolved to meet special problems.

A short contribution incorporating discussion on papers already printed is another particularly valuable way in which the implications of published work can be brought out. It is the Council's view that such discussion between authors and other engineers is best dealt with in this form rather than as correspondence.

The Council hopes, therefore, that all members will consider whether any of their recent work can be described in the form of a short paper, and the Papers Committee will be pleased to consider and advise on any suggestions.

In order to encourage the shorter paper the Council has under consideration the establishment of a special premium for such papers.

INSTITUTION DINNER - MAY 1st, 1958

Members will be aware that it is *not* the Institution's practice to hold an annual dinner in London since it is believed that on the whole members prefer that such functions should take place only when the Institution is not holding a Convention in one of the major universities. This year's dinner, at the Savoy Hotel, London, W.C.1, will therefore be the first main Institution dinner to be held in the capital since 1951.

There has, therefore, been particular interest in the dinner to be held on May 1st, under the Chairmanship of the First Sea Lord, Admiral of the Fleet The Earl Mountbatten of Burma, K.G., P.C., G.C.B., G.C.S.I., G.C.I.E., G.C.V.O., D.S.O. The following list includes the principal guests who have already intimated their intention of being present:

Agnew, W. G., C.V.O. (*Clerk of the Privy Council*)
Alexander, Dr. W. P., Ed.B., M.A., B.Sc. (*Chairman, Association of Education Committees*)
Allen, Sir George V., C.B.E. (*Secretary, British Association for the Advancement of Science*)
Armstrong, Dr. V. (*Chairman, British Commonwealth of Nations Scientific Liaison Office*)
Bachtel, Major C. L. (*Advisory Group, U.S. Army*)
Barnett, Sir Ben, K.B.E., C.B., M.C. (*Chairman, Commonwealth Telecommunications Board*)
Best, Commander K. B., M.V.O., R.N. (Retd.) (*Director of Communications, Home Office*)
Brewell, S. H., M.B.E.
Brockman, Captain R. V., C.S.I., C.I.E., C.B.E., R.N.
Brundrett, Sir Frederick, K.C.B., K.B.E. (*Chairman, Defence Research Policy Committee and Scientific Adviser to the Minister of Defence*)
Cadell, Air Commodore C., C.B.E.
Clarke, Rear Admiral Sir Philip, K.B.E., C.B., D.S.O. (*Past President*)
Curtis, H. A. (*Chairman, Radio Trades Examination Board*)
Dyson, A. A., O.B.E. (*A Trustee of the Benevolent Fund*)
Emmerson, Sir Harold, G.C.B., K.C.V.O. (*Permanent Secretary, Ministry of Labour*)
Emson, Air Commodore R. H. E., C.B.E., A.F.C. (*Director of Air Armaments, Ministry of Supply*)
Evans, H. (*Relay Services Association*)
Gamage, L., M.C., M.A.
Gardner, Dr. G. W. H., C.B., C.B.E., B.Sc. (*Director, Royal Aircraft Establishment, Farnborough*)
Goudime, P., M.A. (*President, Scientific Instrument Manufacturers' Association*)
Hart, Air Marshal Sir Raymund, K.B.E., C.B., M.C.
Hawkins, J. F.
Hinshelwood, Sir Cyril, P.R.S., M.A., D.Sc. (*President, The Royal Society*)
Hives, The Rt. Hon. Lord, C.H., M.B.E., D.Sc., LL.D. (*Chairman, National Council for Technological Awards*)
Horton, P. (*Secretary, Reed's School*)
Hubback, Vice-Admiral Sir Gordon, K.B.E., C.B. (*Fourth Sea Lord*)
Inwood, H., O.B.E.
Jervis, C. (*Press Association*)
Johnson, P., O.B.E., M.A. (*Scientific Adviser to the Army Council*)
Jones, Lt. Col. C. M. Inigo
Joslin, Major-General S. W., C.B., C.B.E. (*Works General Manager, Dounreay Works, U.K.A.E.A.*)
Jourdain, E. D. T. (*Head of Scientific Secretariat, Advisory Council on Scientific Policy*)
Kermode, Air Vice-Marshal A. C., C.B.E., M.A. (*Director of Educational Services, R.A.F.*)
Kirkland, Lt. Colonel G. W., M.B.E. (*Vice-President, The Institution of Structural Engineers*)
Maitland, Major R. F., O.B.E. (*Secretary, The Institution of Structural Engineers*)
Makgill, Sir Donald.
Malloch, J. G., Ph.D. (*National Research Council of Canada*)
Mann, Air Commodore W. E. G., C.B., C.B.E., D.F.C. (*Director General, Civil Aviation Navigational Services, Ministry of Transport and Civil Aviation*)
Marriott, G. A., B.A. (*President of the Institution*)
Marson, Air Vice-Marshal J., C.B., C.B.E. (*Director General Technical Services, Air Ministry*)
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Part, A. A., M.B.E. (*Under Secretary for Further Education, Ministry of Education*)
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Powell, Commander C. (*Secretary, Parliamentary and Scientific Committee*)
Rothschild, Colonel D. D., O.B.E., T.D.
Sims, A. J., O.B.E. (*Director General Ships, The Admiralty*)
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Sturley, K. R., Ph.D. (*Head of Engineering Training Dept., B.B.C.*)
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Uttley, A. M., Ph.D. (*Superintendent, Control Mechanisms and Electronics Division, National Physical Laboratory*)
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Wallis, C. E., M.B.E.
Wansbrough-Jones, Sir Owen, K.B.E., C.B., Ph.D. (*Chief Scientist, Ministry of Supply*)
Wardlaw, Professor W., C.B.E., D.Sc. (*President, The Royal Institute of Chemistry*)
Weston, Air Vice-Marshal J. G. W., C.B., O.B.E. (*Assistant Chief of Air Staff (Signals), Air Ministry*)
Zepler, Professor E. E., Ph.D. (*Vice-President of the Institution*)

TRANSISTOR NOISE
ITS ORIGIN, MEASUREMENT AND BEHAVIOUR†

by

B. L. H. Wilson, M.A.‡

SUMMARY

The sources of noise in semi-conductors and the mathematical techniques needed in their discussion are indicated in order to survey the theory of noise in transistor amplifiers and to consider methods of measurement. The variation of transistor noise with operating point and frequency is discussed and a comparison is made of noise levels in audio amplifiers using transistors and valves respectively.

1. Introduction

When the transistor was introduced it seemed that noise might seriously limit its general application. The junction transistor showed a marked improvement over the point-contact type, but at first nearly all of these units were 10–30 db noisier than the theoretical limit at 1 kc/s—a frequency which is widely adopted as a reference in this work. Moreover, in many units the noise increased with time, especially before good encapsulation techniques were discovered. One or two of the early units were reported to have noise only a few decibels over the thermal noise in the input generator resistance, as was predicted from consideration of shot and thermal noise only. With the improvement in lifetime and dislocation density in germanium, and the development of techniques of etching and aging the germanium surface and of encapsulating the transistor, such transistors are becoming available commercially.

Much of sections 2 and 3 may be omitted by the reader principally interested in applications.

2. Mathematical Background

Suppose $x_1(t)$ represents a noise current or voltage as a function of time. Then the Fourier spectrum over a time T is

$$X_1(f)=\left(\frac{2}{T}\right)^{\frac{1}{2}}\int_0^T x_1(t)\exp(-\mathrm{j}2\pi ft)\mathrm{d}t \quad \text{.........(1)}$$

The interval T is taken so large that the transients at the beginning and end of the interval do not significantly affect the magnitude of $X_1(f)$. The mean square value of X_1 is called the power spectrum

$$W(f)=\overline{X_1{}^*X_1} \quad \text{.........(2)}$$

where the bar denotes the average and the star the complex conjugate. This is the power dissipated if the noise current or voltage is applied to one ohm.

The cross spectrum W_{12} between currents x_1 and x_2 is defined by

$$W_{12}(f)=\overline{X_1{}^*X_2}=W_{21}{}^*(f) \quad \text{.........(3)}$$

Unless x_1 and x_2 are coherent so that there is a systematic phase difference between X_1 and X_2, W_{12} will be zero. A normalized cross spectrum may be defined by

$$W_{12}'=W_{12}/(W_1W_2)^{\frac{1}{2}} \quad \text{.........(4)}$$

The correlation between two noise currents $y_1(t)$ and $y_2(t)$ is defined as

$$r_{12}=\overline{y_1y_2}/\overline{y_1{}^2}\,\overline{y_2{}^2})^{\frac{1}{2}} \quad \text{.........(5)}$$

If y_1 and y_2 are the currents resulting from passing x_1 and x_2 through identical narrow band filters it may be shown that

$$r_{12}=\mathfrak{R}\,(W_{12}') \quad \text{.........(6)}$$

where $\mathfrak{R}$ indicates the real part.

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We shall now show how these quantities may be used in circuit analysis in a way analogous to ordinary steady state methods. Suppose

$$x_3 = x_1 + x_2 \quad \text{.........(7)}$$

Then from (1)

$$X_3 = X_1 + X_2 \quad \text{.........(8)}$$

and multiplying (7) by its complex conjugate

$$\begin{aligned} W_3 &= W_1 + W_2 + W_{12} + W_{21} \\ &= W_1 + W_2 + 2\mathcal{R}(W_{12}) \\ &= W_1 + W_2 + 2r_{12}(W_1 W_2)^{\frac{1}{2}} \quad \text{.........(9)} \end{aligned}$$

The Fourier spectrum of eqn. (1) may be multiplied by impedance, admittance or transfer operators. If $x_1(t)$ is passed through a linear network having a transfer function $A(f)$, and $x_2(t)$ is the resultant voltage or current,

$$X_2(f) = X_1(f) A(f) \quad \text{.........(10)}$$

If a system of noise currents and/or voltages X is related by

$$\begin{aligned} X_A &= A_1 X_1 + A_2 X_2 + \ldots \quad \text{.........(11)} \\ X_B &= B_1 X_1 + B_2 X_2 + \ldots \end{aligned}$$

the corresponding power spectra follow from the definitions and equation (10) in a similar way to the special case given in equations (7), (8) and (9):

$$\begin{aligned} W_A &= |A_1|^2 W_1 + |A_2|^2 W_2 + \mathcal{R}(A_1^* A_2 W_{12}) \quad \text{.........(12)} \\ W_{AB} &= A_1^* B_1 W_1 + A_2^* B_2 W_2 + \\ &\quad + A_2^* B_1 W_{21} + A_1^* B_2 W_{12} \end{aligned}$$

Where X_1 and X_2 are independent we have simply

$$\begin{aligned} W_A &= |A_1|^2 W_1 + |A_2|^2 W_2 \quad \text{......(13)} \\ W_{AB} &= A_1^* B_1 W_1 + A_2^* B_2 W_2 \end{aligned}$$

We now suppose the quantity x to be redefined so that $\bar{x} = 0$. With this restriction we define the auto-correlation function,

$$\overline{x(t)\, x(t+u)},$$

the average being taken over the variable t so that

$$\overline{x(t)\, x(t+u)} = \lim_{T \to \infty} \frac{1}{T} \int_0^T x(t)\, x(t+u) \mathrm{d}t \quad \text{.........(13)}$$

A normalized form of the auto-correlation function,

$$\frac{\overline{x(t)\, x(t+u)}}{\overline{x^2(t)}}$$

is often useful.

The power spectrum may be calculated by means of the auto-correlation function by the Wiener-Khitchine theorem

$$W(f) = 4 \int_0^\infty \overline{x(t)\, x(t+u)} \cos 2\pi f u \, du \quad \text{.........(14)}$$

The inverse theorem is

$$\overline{x(t)\, x(t+u)} = \int_0^\infty W(f) \cos 2\pi f u \, df \quad \text{.........(15)}$$

In addition to the above, a theorem known as Peterson's equivalent generator theorem is implicit in the development below.

"The noise of a linear active four-pole requires four real parameters for its specification in addition to the four complex parameters describing its signal behaviour. These may conveniently be voltage or current generators at the input and output terminals according to whether the open circuit or short circuit signal parameters are used. The additional parameters may be the power spectra of the two generators and their (complex) cross spectrum."

The theorem is proved by writing down the mesh equations of the four-pole when four additional parameters are required to determine the statistical behaviour of the noise fully.

3. Types of Noise Encountered in Semi-conductors

3.1. *Johnson Noise*

A voltage is generated as a result of fluctuations from thermodynamic equilibrium in series with any impedance Z in equilibruim with its surroundings. The value of these fluctuations may be calculated from thermodynamic arguments alone, without considering the detailed processes which produce the fluctuations. The voltage is given by

$$\overline{E^2}(f)\Delta f = 4kT\mathcal{R}(Z)\Delta f \quad \text{.........(16)}$$

where k = Boltzmann's constant ($1{\cdot}347 \times 10^{-23}$ W/°C.)

T = temperature in degrees absolute

Δf = bandwidth in which the fluctuations are measured.

Alternatively the fluctuations may be described by a current generator in parallel with the admittance Y such that the current I is given by

$$\overline{I^2}(f)\Delta f = 4\,kT\mathcal{R}(Y)\Delta f \qquad \text{.........(17)}$$

This noise is known as the thermal or Johnson noise. The formulae may be proved by considering the equipartition of energy between the impedance Z and an L-C resonant circuit or a transmission line. This proof and much other background work are given by van der Ziel[4] and by Pierce[5]. It is found experimentally that the formulae hold for wire resistances even under bias.

3.2. *Shot Noise*

When a direct current i flows between the electrodes of a thermionic diode, noise is generated owing to the finite size of the electronic charge, e. If the electrons flow independently the noise current at frequencies much less than the reciprocal of the transit time is given by

$$\overline{I^2}(f)\Delta f = 2e\bar{i}\Delta f \qquad \text{.........(18)}$$

The derivation of this result is fully treated by Rice.[1]

The shot noise equation (18) is directly applicable to the temperature-limited thermionic diode with a tungsten filament (down to about 10 c/s). In other cases the requirement that the electrons should flow independently is not fulfilled. In the space charge limited diode, less noise is observed than that predicted, while if an oxide coated cathode is used, increased noise is observed at low frequencies. The same analysis may be applied to noise in p-n junctions at low frequencies. Shot noise in semi-conductors is discussed in Section 3.4.

3.3 *Partition Noise*

A further consequence of the discreteness of the electronic charge is partition noise, which arises chiefly in multi-electrode devices Suppose two independent courses are open to an assembly of n electrons and the probabilities that an electron takes one or other of the two courses are p and $1-p$. The statistics of such an assembly are governed by the Poisson distribution, for which it may be shown that if m_1 electrons take the first course and m_2 take the second, the mean and variance of m_1 and m_2 are given by

$$\left.\begin{aligned}\overline{m_1} &= np\\ \overline{m_2} &= n(1-p)\end{aligned}\right\} \qquad \text{.........(19)}$$

$$\overline{(m_1-\overline{m_1})^2} = \overline{(m_2-\overline{m_2})^2} = np(1-p) \qquad \text{.........(20)}$$

The calculation of the corresponding partition noise current follows the lines of the next section. Care must be taken in applying the formula to ensure that there are only two choices open to each electron and that all the electrons are equivalent. We shall not use this method to calculate the noise in transistors.

3.4. *Shot and Partition Noise in Semi-conductors*

Specialized types of shot and partition noise important in semi-conductors arise when the carriers are trapped or when two carriers of opposite sign recombine. The spectrum is most easily obtained from the Wiener-Khitchine theorem (14). Consider, for example, a uniform prism of material of length l and constant cross-section containing one type of carrier decaying exponentially with a lifetime τ and suppose the time to drift under an applied field E between two electrodes on opposite faces is τ_0. Let $i(t)$ and $\bar{i}$ be the instantaneous and mean current, $n(t)$ and $\bar{n}$ the instantaneous and mean number of carriers in the prism.

Suppose that the fluctuation $n-\bar{n}$ in the number of carriers has a value $\Delta n(t)$ at time t. At a time $t+s$ the fluctuation has on average decreased to a value $\Delta n(t)\exp(-s/\tau)$ owing to decay, and to $\Delta n(t)(1-s/\tau_0)$, if $s<\tau_0$, owing to the sweeping out of carriers at an electrode. The combination of these effects gives

$$\begin{aligned}\overline{\Delta n(t+s)} &= \Delta n(t)\exp(-s/\tau)\,(1-s/\tau_0), && s<\tau_0\\ &= 0, && s>\tau_0\end{aligned} \qquad \text{......(21)}$$

The bar here indicates an average over values of t.

The auto-correlation function for n is then

$$\begin{aligned}\overline{\Delta n(t+s)\Delta n(t)} &= \overline{(\Delta n)^2}\exp(-s/\tau)\,(1-s/\tau_0), && s<\tau_0\\ &= 0 && s>\tau_0\end{aligned} \qquad \text{......(22)}$$

Now the current in an external circuit depends instantaneously on the sum of the contributions of every carrier in the material at each instant.

Under a uniform field the current is given by

$$i = neE\mu/l, \qquad \ldots\ldots(23)$$

where μ is the mobility.

So we have analogously for the current fluctuation Δi

$$\overline{\Delta i(t+s)\Delta i(t)} = \overline{(\Delta i)^2}\exp(-s/\tau)(1-s/\tau_0), s<\tau_0$$
$$= 0 \qquad s>\tau_0 \qquad \ldots\ldots(24)$$

The quantity $\overline{\Delta n^2}$ is determined by the statistics of recombination. For Poisson statistics, i.e. where the probability of decay of each electron is independent,

$$\overline{(\Delta n)^2} = \bar{n} \qquad \ldots\ldots(25)$$

Under this restriction the problem has been solved by Davydov and Gurevitch[6]. Two special cases arise; when $\tau \gg \tau_0$ and when $\tau \ll \tau_0$. In the former case,

$$\overline{\Delta i(t+s)\ \ \Delta i(t)} = \overline{(\Delta i)^2}(1-s/\tau_0),\ s<\tau_0 \qquad \ldots(22a)$$

Then applying the Wiener-Khitchine theorom and using (23 and 25),

$$\overline{I^2}(f) = 2e\bar{i}\frac{\sin^2\omega\tau_0/2}{(\omega\tau_0/2)^2} \qquad \ldots\ldots(26)$$

This reduces to the usual shot noise formula at low frequencies; the high frequency variation alone differs for different electrode geometries. The application of (25) is justifiable here if the electrons are removed independently at the electrode. This is the case relevant to the analysis of shot noise in transistors. For the case $\tau \ll \tau_0$, where

$$\overline{\Delta i(t+s)\Delta i(t)} = \overline{(\Delta i)^2}\exp(-s/\tau) \ldots\ldots(22b)$$

there is obtained

$$\overline{I^2}(f) = \left(\frac{Ee}{l}\right)^2 \overline{(\Delta n)^2}\frac{4\tau}{1+\omega^2\tau^2} \qquad \ldots\ldots(27)$$

The variance, $\overline{\Delta n^2}$, depends on the type of statistics applicable to the decay process. Its value has been determined by Burgess[7,8] for a semi-conductor with N donor levels by two methods: one, purely thermodynamic, involves an expansion of the free energy in terms of the fluctuations Δn, Δp in numbers of holes and electrons; the other is based on the statistics governing the generation and recombination of holes and electrons and the ionization and condensation of electrons at impurity levels. For the case of strongly extrinsic n-type material, equivalent to the present case, he obtains

$$\overline{\Delta n^2} = \left(\frac{2}{\bar{n}} + \frac{1}{N-\bar{n}}\right)^{-1} \qquad \ldots\ldots(28)$$

This is the formula applicable to n-type transistor and diode material in bulk. Earlier analyses based on an immediate application of Poisson statistics gave, as in eqn. (20),

$$\overline{\Delta n^2} = Np(1-p) \text{ where } p = \frac{n}{N}$$
$$= \left(\frac{1}{\bar{n}} + \frac{1}{N-\bar{n}}\right)^{-1} \qquad \ldots\ldots(29)$$

More complicated cases may be analysed with the aid of the expressions given by Burgess for the variances $\overline{(\Delta n)^2}$, $\overline{(\Delta p)^2}$ and $\overline{\Delta n \Delta p}$. The frequency dependence in cases where a fluctuation decays exponentially must however be that of eqn. (27).

3.5. *Flicker Noise*

While frequency dependence governed by a single relaxation time τ as described above has been detected, for example by Herzog and van der Ziel[9] in single crystal filaments of germanium, the experimental observations of frequency dependent noise are dominated by noise whose power spectrum varies as f^{-m} with $1<m<1{\cdot}5$, and frequently $m \cong 1$. This is widely referred to as "flicker" or "$1/f$ noise". Furthermore, the magnitude of the noise is often much greater than would be predicted by the type of mechanism suggested above whose statistics do not allow $\overline{(\Delta n)^2} \gg n$. Noise of the types discussed in section 3.4 is known as electronic noise, being due to the generation, scattering and decay of the current carriers themselves. It is believed that the flicker component in noise is due to the modulation of carrier flow by charged centres at potential barriers. These barriers occur at $p-n$ junctions in the bulk material or at the surface of the semi-conductor. In single crystal germanium and silicon, flicker noise has been found to be extremely dependent on surface conditions, particularly on the presence of water vapour. For example Pearson *et al.*[10] found that no noise other than shot noise was present in a silicon junction down to 80 c/s under dry conditions, but that under 100 per cent relative humidity the noise showed an $f^{-1{\cdot}2}$ dependence and the current noise power at 100 c/s was greater than that of the shot noise by a factor of 3×10^5. The change was reversible and intermediate effects were observed at lower humidities.

Flicker noise has been found in nearly all semi-conductors and in metal films. The persistence of an $f^{-1.0}$ dependence in noise at low frequencies has attracted interest as it implies an infinite noise power at zero frequency. Rollin and Templeton[11, 12] investigating the range at 7·5 c/s to $2{\cdot}5 \times 10^{-4}$ c/s found an f^{-1} spectrum for noise from carbon resistors and an $f^{-1.35}$ spectrum with germanium filaments. Hyde has fitted a composite spectrum consisting of white, f^{-1} and two discrete relaxation time components to the noise from a point contact rectifier over the range 0·117 c/s to 14 Mc/s, identifying an f^{-1} component over 7 decades. He has also fitted a similar spectrum, with one discrete relaxation time, to noise from a junction rectifier. At high frequencies the $1/f$ component is eventually swamped by the white noise or discrete relaxation time components.

Many authors have discussed the problem of deriving an f^{-1} spectrum from theory. The early workers were largely concerned with the oxide-coated cathode. (For a review of the earlier work with references, see van der Ziel[15].) It seems necessary to postulate a spread g(τ) of relaxation times. If for each relaxation time (22b) and (27) apply independently, integration over τ yields

$$\overline{I^2(f)} \propto \overline{(\Delta n)^2} \int \frac{g(\tau)\,\tau}{1+\omega^2\tau^2}\,d\tau \qquad \ldots\ldots(30)$$

Any sufficiently slow variation in the numerator can be shown to give an f^{-1} spectrum. The spectrum in τ has been attributed to a diffusion mechanism, but detailed models based on diffusion alone do not readily give an f^{-1} spectrum. Van der Ziel[15] and du Pré independently gave a plausible origin of g(τ) by referring the distribution in τ back to a distribution of activation energies E, such that

$$\tau = \tau_0 \exp E/kT \qquad \ldots\ldots(31)$$

If E is uniformly distributed between the limits E_1 and E_2, that is

$$\left.\begin{aligned} g(E) &= dE/(E_1 - E_2) \qquad & E_2 < E < E_1 \\ &= 0 & \text{otherwise} \end{aligned}\right\} \ldots\ldots(32)$$

then $$g(\tau) = \frac{d\tau}{\log_e \tau_2/\tau_1} \quad \tau_2 < \tau < \tau_1 \qquad \ldots\ldots(33)$$

giving $$\overline{I^2(f)} \propto \frac{\tan^{-1}\omega\tau_1 - \tan^{-1}\omega\tau_2}{\omega \log_e \tau_2/\tau_1} \qquad \ldots\ldots(34)$$

which corresponds very nearly to an f^{-1} variation in the range $1/\tau_2 < \omega < 1/\tau_1$. The spectrum is insensitive to small variations in g(E). Direct evidence for such a distribution of traps on semi-conductor surfaces has been found by Kingston and McWhorter[16]. An analysis deriving a flicker noise spectrum for linear and planar arrays of modulating traps such as might occur at dislocations and surface barriers is given by Morrison[17]. No analysis of the process on the lines of that by Burgess for a single relaxation time has yet been given.

3.6. *Surface and Leakage Noise*

In transistors and diodes it is convenient to distinguish between surface and leakage components of flicker noise, as described by Fonger[18]. Surface noise is attributed to fluctuations in barrier height at the surface of a semi-conductor, which in turn give rise to fluctations in the process of generation and recombination of pairs of holes and electrons and may be described by the equation

$$\overline{G^2(f)} = (\Delta p)^2 \Psi(f) \qquad \ldots\ldots(35)$$

where G = rate of pair generation/unit frequency/unit area
Δp = mean excess minority carrier density
$\Psi(f)$ = parameter characteristic of the surface.

This is the analogue of the equation describing the average rate of generation of pairs

$$r = (\Delta p)s \qquad \ldots\ldots(36)$$

where r = mean rate of pair generation
s = parameter characteristic of the surface (surface recombination velocity).

Leakage noise is attributed to fluctuations in barrier height at a junction, which are directly responsible for fluctuations in leakage current and may be described by a noise current generator in parallel with the junction.

Fluctuations of any sort will be increased by any multiplication (avalanche) process at the junction. McKay[19] describes the noise pulses at a silicon junction diode near breakdown: the noise consists of current pulses of constant amplitude and variable lengths which give a rather uniform spectrum.

Spontaneous fluctuations in temperature, predictable thermodynamically, are a further source of noise where the characteristics of a device are controlled by the behaviour of a small volume of material, as in point contact diodes, for example. Noise also arises in faulty devices owing to intermittent contact: it is then usually erratic and large, especially at low frequencies. In the absence of external power sources, of course, neither these nor any other noise sources give noise greater than the thermal noise.

4. Noise in Transistor Amplifiers—Definitions and Circuit Properties

4.1. *Noise Figure*

The noise figure or noise factor of an amplifier may be defined by

$$F=\frac{(S/N)_i}{(S/N)_o} \qquad \ldots\ldots(37)$$

where S/N is the ratio of the signal to noise power available at the terminals and i, o refer to the input and output. The quantity given by this equation at a spot frequency defines the spot noise figure at that frequency.

The noise figure at the input is determined ideally by the thermal noise in the generator resistance (see Fig. 1).

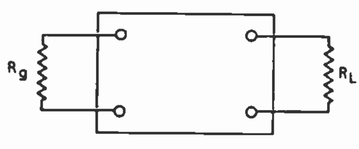


Fig. 1.

If R_g is the generator resistance
R_L is the load,

$\overline{E^2_o}(f)$ is the mean square noise voltage at output in a bandwidth of one cycle per second,

$A(f)$ is the voltage gain of transistor,

$G(f)$ is the transducer power gain of transistor,

$P(f)=\overline{E^2_o}/R_L$ is the noise power dissipated in the load,

F is the noise figure at the frequency f,

we have, from (37),

$$F=\frac{1}{A^2}\,\frac{\overline{E_o^2}}{4kTR_g}\equiv\frac{1}{G}\cdot\frac{P}{kT} \qquad \ldots\ldots(38)$$

Thus the noise figure may be equivalently defined as the ratio of the noise power at the output to that portion of it engendered by thermal noise in the generator resistance. The corresponding quantity defined for a two-terminal device is usually known as the noise ratio. It should be noted that, as the transistor output impedance is the same for signal and noise, the noise figure is independent of the load. It is, however, a function of the generator resistance.

In practice amplifiers of finite bandwidth must be used. Suppose the transistor amplifier is cascaded with a measuring amplifier and square-law detector whose response is $B(f)$. An average noise figure can be defined by

$$\overline{F}=\frac{\int_0^\infty \overline{E^2_0}(f)B^2(f)\mathrm{d}f}{4kTR_g\int_0^\infty A^2(f)B^2(f)\mathrm{d}f} \qquad \ldots\ldots(39)$$

i.e. the ratio of the total noise output at the detector to that portion of it engendered by thermal noise in the generator resistance. In terms of a reference frequency f_0, often the frequency of peak response, the denominator may be rewritten

$$4kTR_gA^2(f_0)B^2(f_0)\Delta f \qquad \ldots\ldots(40)$$

where Δf, the noise bandwidth is given by

$$\Delta f=\int_0^\infty\frac{A^2(f)}{A^2(f_0)}\cdot\frac{B^2(f)}{B^2(f_0)}\,\mathrm{d}f \qquad \ldots\ldots(41)$$

It frequently can be arranged that $A(f)=A(f_0)$ where $B(f)/B(f_0)$ is appreciable. For an ideal singly-tuned amplifier the noise bandwidth is $\pi/2$ times the bandwidth defined by the points where the gain is reduced by 3db.

The most commonly quoted audio-frequency noise figure is that at 1 kc/s which we call F_0. Audio-frequency noise in some transistors is dominated by $1/f$ noise and for such transistors, in a frequency band $f_1<f<f_2$ where the voltage gain is constant, the output noise voltage is

given by

$$\overline{E^2}=\int_{f_1}^{f_2} 4kTR_g A^2(f)F(f)\mathrm{d}f=4kTR_g A^2 F_0 \log_e f_2/f_1 \quad \ldots\ldots(42)$$

a useful formula in designing wide-band audio amplifiers.

4.2. *Relation of Noise Figure to the Equivalent Noise Generators*

An alternative to the use of the noise figure as a parameter is the direct specification of the noise sources. The noise sources may be formally represented by two voltage or current generators at the input and output of the transistor as discussed above. From these the noise figure may be derived. For example we shall consider a transistor whose signal parameters are specified by its real impedance matrix so that

$$\begin{aligned} V_1&=I_1R_{11}+I_2R_{12} \\ V_2&=I_1R_{21}+I_2R_{22} \end{aligned} \quad \ldots\ldots(43)$$

where V_1, I_1 are the input voltage and current, V_2, I_2 are the output voltage and current.

Suppose that the noise voltage generators at input and output are $E_i(f)$, $E_o(f)$ and the generator resistance is R_g. We arbitrarily let the load be infinite and determine the open circut output voltage V (see Fig. 2).

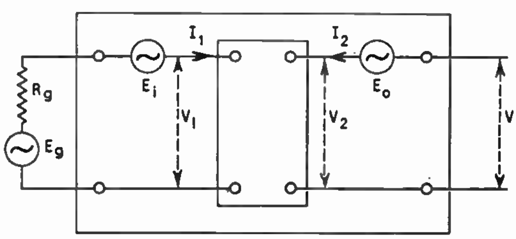


Fig. 2. Transistor with noise voltage generators.

With all the noise generators in circuit,

$$\begin{aligned} E_i+E_g&=I_1(R_{11}+R_g) \\ V&=E_o+I_1R_{21} \end{aligned} \quad \ldots\ldots(44)$$

while with the thermal noise sources alone, using dashed suffixes to indicate the new voltage and current,

$$\begin{aligned} E_g&=I'_1(R_{11}+R_g) \\ V'&=E_o+I'_1R_{21} \end{aligned} \quad \ldots\ldots(45)$$

Eliminating I_1, I'_1

$$V=E_o+\frac{R_{21}}{R_{11}+R_g}(E_i+E_g)$$

$$V'=\frac{R_{21}}{R_{11}+R_g}E_g$$

Use eqn. (12), bearing in mind that E_g is independent of E_i

$$\overline{V^2}=\left(\frac{R_{21}}{R_{11}+R_g}\right)^2\overline{E^2_g}+\overline{E^2_o}+\left(\frac{R_{21}}{R_{11}+R_g}\right)^2\overline{E_i^2}+ + \frac{R_{21}+R_g}{R_{21}}\rho_{io}(\overline{E^2_i}\,\overline{E^2_o})^{\frac{1}{2}}$$

$$\overline{V'^2}=\left(\frac{R_{21}}{R_{11}+R_g}\right)^2\overline{E^2_g}$$

where ρ_{io} is the normalized correlation between E_i and E_o.

Putting $\overline{E^2_g}=4kTR_g$, $F=\overline{V^2}/\overline{V'^2}$, we obtain

$$F=1+\frac{1}{4kTR_g}\left\{\overline{E^2_i}+\overline{E^2_o}\left(\frac{R_{11}+R_g}{R_{21}}\right)^2 +\rho_{io}(\overline{E^2_i}\,\overline{E^2_o})^{\frac{1}{2}}\frac{R_{11}+R_g}{R_{21}}\right\} \quad \ldots\ldots(46)$$

A large number of similar cases may be worked out using different noise generators—or with the transistor specified by different parameters. The converse relation may, in theory, also be used: from a knowledge of the value of F as the (in general complex) generator impedance is varied, the values of the noise generators may be obtained. An interesting approach to the problem of the variation of F with source impedance is given by Rothe and Dahlke[20]. The noisy transistor four-pole is transformed into a noise-free signal four-pole and a preceding four-pole of two partially correlated noise generators. This last four-pole may be further transformed into an equivalent four-pole consisting of a correlated pair of current and voltage generators described by their equivalent noise conductance, G_n, noise resistance, R_n, and a correlation admittance, Y_n. The source admittance for minimum noise figure is obtained in terms of G_n, R_n and Y_n.

There are, however, great practical difficulties in determining the equivalent noise generators directly, especially with good junction transistors, owing to the low levels involved, and a measurement of noise figure may be all that is practicable.

4.3. *Use of Linear and Peak-reading Detectors*

A square-law detector is not always used at the output of the amplifier when measuring the noise figure. If a linear detector is used, the expressions (37) to (39) may be taken to refer to a quantity F_{mean} when the mean square voltages have been replaced by the square of the mean full-wave rectified voltage.

As the thermal noise is obtained from calculation it must be remembered that

$$\overline{|V|}^2 = \frac{2}{\pi}\overline{V^2} \qquad \text{......(47)}$$

for Gaussian noise. A quantity F_{peak} may be similarly defined if a peak-reading detector is used, replacing the mean square voltage by the square of the peak voltage, $\hat{V}$. Although the precise ratio of $\hat{V}^2$ to $\overline{V^2}$ for thermal noise depends on the frequency response of the amplifier and the time-constant of the peak detector, usually

$$\hat{V} \simeq 3(\overline{V^2})^{\frac{1}{2}} \qquad \text{......(48)}$$

Of course, if transistor noise were Gaussian, like thermal noise, we should expect F_{peak}, F_{mean} and F to be the same. In fact, it is found that F_{peak} tends to be larger than F when F is large.

In practice it is found that F_{peak} for a wide-band amplifier is the best measure of the annoyance to a listener in audio applications. As is well known, white noise gives rise to a hiss, while $1/f$ noise sounds rougher. Noise due to intermittent contacts causes irregular crackle which is extremely objectionable although the power associated with it and hence the value of F may be quite small.

The ratio of peak to r.m.s. noise has been called the crest factor by Englund[21]; this is simply related to the ratio of F_{peak} to F. In early transistors measured by him, he found the crest factor was larger for noisy units, while for a later design the crest factor was independent of the noise. This appeared to be associated with a decrease in leakage noise.

5. Noise in Junction Diodes and Transistors—Theory

5.1. *Noise in Junction Diodes*

At low frequencies, that is frequencies low compared with the time for a carrier to diffuse across a diode, the shot noise formula (18) may be applied at once to junction diodes. The diode current may be considered as the algebraic sum of two diffusion currents flowing in opposite directions; thus if V is the applied voltage, $I_0 \exp \Lambda V$ flows in the forward direction and I_0, the reverse saturation current, in the reverse direction, where $\Lambda = e/kT$. Each of these components is itself in general composed of electron and hole currents in opposite directions, though in practice the flow often consists largely of the carrier which is in the majority in the region of lower resistivity. The net current is

$$I = I_0 \left\{ \exp(\Lambda V) - 1 \right\} \qquad \text{......(49)}$$

Applying the shot noise formula to each independent current and adding

$$\overline{i^2}(f) = 2eI_0(\exp \Lambda V + 1) = 2e(I + 2I_0) \qquad \text{......(50)}$$

In this section we write the spectrum of noise currents and voltage with a small letter and direct currents with a large. Thus $\overline{i^2}(f)$ is the power spectrum of noise current per cycle bandwidth. The conductance derived from equation (49) is

$$g = \Lambda(I + I_0) \qquad \text{......(51)}$$

So the noise ratio

$$F = (I + 2I_0)/2(I + I_0) \qquad \text{......(52)}$$

which is unity without bias, as required thermodynamically, approaches the value 0·5 in the forward direction and approaches zero in the reverse. Allowance must in practice also be made for thermal noise generated in the base resistance of the diodes. Values consistent with eqn. (52) after subtracting flicker noise have been obtained in germanium junctions by Anderson and van der Ziel[22]. At higher frequencies the noise ratio is higher than that expected on this theory, as discussed below.

5.2. *Noise in Junction Transistors*

This theory, the emission theory of shot noise in which no allowance is made for transit time, can be applied to the junction transistor except at high frequencies. If two shot noise sources in the transistor are postulated it is necessary to suppose some correlation between them and, using the theory of partition noise, to determine the noise due to the partition of emitter current between base and collector[23,24]. We shall follow an alternative approach due to Giacoletto[25], who considers the diffusion currents between each pair of transistor terminals separately and obtains three independent sources of current noise.

The calculations below neglect two sources of feedback, the variation in the width of the space charge region with voltage and base to collector leakage. Both of these can be shown to be unimportant here. The derivation of the noise current generators is at first made in terms of the intrinsic transistor, that is the transistor without extrinsic base resistance. The noise figure is later calculated after allowance for extrinsic base resistance.

The direct current through the intrinsic transistor obeys the equations

$$I_e = I_{ee}(\exp\ \Lambda V_{eb'} - 1) - I_{ec}(\exp\ \Lambda V_{cb'} - 1)$$

$$I_c = I_{ec}(\exp\ \Lambda V_{eb'} - 1) - I_{cc}(\exp\ \Lambda V_{cb'} - 1) \qquad \ldots\ldots(53)$$

where I_e, I_c are the currents flowing in at the emitter and out at the collector, and V_{eb}' = potential difference between emitter and intrinsic base, etc.

These equations may be re-written as

$$I_e = I_{es}(\exp\Lambda V_{eb'} - 1) + I_{ec}(\exp\ \Lambda V_{eb'} - \exp\ \Lambda V_{cb'})$$

$$I_c = -I_{cs}(\exp\Lambda V_{cb'} - I) + I_{ec}(\exp\ \Lambda V_{eb'} - \exp\Lambda V_{cb'}) \qquad \ldots\ldots(54)$$

where $I_{es} = I_{ee} - I_{ec}$

$I_{cs} = I_{cc} - I_{ec}$

Here the emitter and collector currents are split up into the three components flowing between emitter and (intrinsic) base, emitter and collector, and collector and base. Each component consists of two opposed portions as in the diode treatment. For example, where injection takes place from the emitter only, that is the emitter has unit injection efficiency, $I_{es}\exp\Lambda V_{eb'}$ is injected by the emitter and diffuses to the base and I_{es} is generated in the base and diffuses to the emitter. If phase shifts due to finite transit times between the terminals are negligible, each portion develops full shot noise between the appropriate terminals

$$\left.\begin{aligned} \overline{i^2_{eb'}} &= 2qI_{es}(\exp\ \Lambda V_{eb'} + 1) \\ \overline{i^2_{ec}} &= 2qI_{ec}(\exp\Lambda V_{eb'} + \exp\Lambda V_{cb'}) \\ \overline{i^2_{cb'}} &= 2qI_{cs}(\exp\ \Lambda V_{cb'} + 1) \end{aligned}\right\} \quad \ldots\ldots(55)$$

where q is the electronic charge

and $\overline{i^2_{eb'}}$ is the mean square noise current per cycle per second generated between emitter and base, etc.

As usually biased $\Lambda V_{cb'} \ll 1$, $\exp V_{cb} \cong 0$. Under this condition the d.c. equations become

$$\left.\begin{aligned} I_e &= I_{es}(\exp\Lambda V_{eb'} - 1) + I_{ec}\exp\ \Lambda V_{eb'} \\ I_c &= I_{cs} \qquad\qquad\qquad\quad + I_{ec}\exp\ \Lambda V_{eb'} \\ I_b &= I_e - I_c = I_{es}(\exp\ \Lambda V_{eb'} - 1) - I_{cs} \end{aligned}\right\} \ldots\ldots(56)$$

and the noise equations become

$$\begin{aligned} \overline{i^2_{eb}}\ /2q &= I_b + 2I_{es} + I_{cs} \\ \overline{i^2_{ec}}\ /2q &= I_c - I_{cs} \\ \overline{i^2_{cb'}}/2q &= I_{cs} \end{aligned} \qquad \ldots\ldots(57)$$

The current noise generators above may be used in combination with any transistor equivalent circuit, with or without the addition of feedback elements, to obtain the noise figure in a given circuit configuration. The calculation is again simplified by remembering that the noise figure is independent of the load.

5.3. *Calculation of the Noise Figure for Grounded Emitter Configuration*

It seems worthwhile to calculate the noise figure in terms of the commonly used transistor parameters. We shall consider the white noise sources in a simplified transistor with an extrinsic base resistance $r_{b'}$. For the grounded emitter the well-known equivalent circuit is shown in Fig. 3.

The usual symbols are used:

r_g = generator resistance.

r_e' = emitter resistance in grounded emitter configuration.

$r_{b'}$ = extrinsic base resistance.

α' = short circuit current gain in this configuration.

e, b', b, c denote emitter, base of the intrinsic transistor, extrinsic base and collector respectively.

$i_{b'e}$, i_{ce}, $i_{b'c}$, are the noise current generators between the terminals given by the subscripts.

e_b, e_g, are the thermal noise generators in $r_{b'}$, r_g.

We have

$$\left.\begin{aligned}\overline{e^2_b}&=4kTr_{b'}\\ \overline{e^2_g}&=4kTr_g\end{aligned}\right\} \qquad \ldots\ldots(58)$$

For convenience we choose zero load, then, as all the noise generators are independent, we calculate the short-circuit output current for each generator in turn. The total noise figure is the sum of the increments in noise figure for each separate noise generator, which are given by the ratio of the mean short-circuit output current due to the noise generator under consideration to that due to the thermal noise generator.

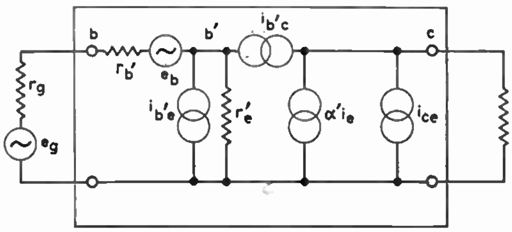


Fig. 3. Eqivalent circuit for grounded emitter.

The short-circuit output current i_o due to e_g, is $\alpha' e_g/(r_b'+r_e'+r_g)$

The short-circuit output current i_1 due to e_b, is $\alpha' e_b/(r_{b'}+r_e'+r_g)$
so the increment in noise figure due to e_b is

$$\Delta F_1=\overline{i^2_1}/\overline{i^2_o}=\overline{e^2_b}/\overline{e^2_g}=r_{b'}/r_g \qquad \ldots\ldots(59)$$

The short-circuit output current due to $i_{b'e}$ is $\alpha' i_{b'e}(r_g+r_{b'})/(r_{b'}+r_e'+r_g)$ so the increment ΔF_2 due to $i_{b'e}$ is given by the expression

$$\Delta F_2=\overline{i^2_{b'e}}(r_g+r_{b'})^2/\overline{e^2_g}=i^2_{b'e}r_g(1+r_{b'}/r_g)^2/4kT \qquad \ldots\ldots(60)$$

Similarly the increment due to i_{ce} is given by

$$\Delta F_3=\frac{\overline{i^2_{ce}}(r_{b'}+r'_e+r_g)^2}{4kTr_g\alpha'^2}$$

$$=\frac{\overline{i^2_{ce}}r_g}{4kT}\left(1+\frac{r_{b'}}{r_g}\right)^2\left(1+\frac{r_e'}{r_b+r_g}\right)^2\frac{1}{\alpha'^2} \qquad \ldots\ldots(61)$$

and that due to $i_{b'c}$, by $\Delta F_4=$

$$\frac{i^2_{b'c}r_g}{4k\,T}\left\{1+\frac{r_{b'}}{r_g}\right\}^2\left\{1+\frac{1}{\alpha'}\left(\frac{r_e}{r_{b'}+r_g}\right)\right\}^2 \qquad \ldots\ldots(62)$$

Putting $(1+r_{b'}/r_g=Z)$, the total noise figure is

$$F_e=1+\sum_1^4\Delta F_i$$

$$=Z+\frac{Z^2r_g}{4kT}\left[\overline{i^2_{b'e}}+\overline{i^2_{ce}}\left\{\frac{1}{\alpha'}\left(1+\frac{r'_e}{Zr_g}\right)\right\}^2+\right.$$

$$\left.+\overline{i^2_{b'c}}\left\{1+\frac{1}{\alpha'}\left(1+\frac{r'_e}{Zr_g}\right)\right\}^2\right] \qquad \ldots\ldots(63)$$

Using (56), if only white noise sources are present,

$F_e=Z+$

$$\frac{\Lambda Z^2r_g}{2}\left[I_b+I_{es}+2I_{es}+(I_c-I_{cs})\left\{\frac{1}{\alpha'}\left(1+\frac{r_e'}{Zr_g}\right)\right\}^2+\right.$$

$$\left.I_{cs}\left\{1+\frac{1}{\alpha'}\left(1+\frac{r_e'}{Zr_g}\right)\right\}^2\right] \qquad \ldots\ldots(64)$$

If I_{es}, I_{cs} are neglected, putting $r'_b=1/\Lambda I_b$,

$$F_e=Z+\frac{\Lambda I_bZ^2r_g}{2}\left[1+\frac{1}{\alpha'}\left\{1+\frac{1}{\Lambda I_bZr_g}\right\}^2\right] \qquad \ldots\ldots(65)$$

With $\Lambda I_br_g=\rho$

$$F_e=Z+\rho\frac{Z^2}{2}\left(1+\frac{1}{\alpha'}\right)+\frac{Z}{\alpha'}+\frac{1}{2\alpha'\rho} \qquad \ldots\ldots(66)$$

F_e is a minimum when

$$\rho=\{Z^2(1+\alpha')\}^{-\frac{1}{2}} \qquad \ldots\ldots(67)$$

At the minimum,

$$F_e=Z\left\{1+\frac{1}{\alpha'}(1+\sqrt{1+\alpha'})\right\}$$

$$\simeq Z(1+1/\sqrt{\alpha'}) \qquad \ldots\ldots(68)$$

For example, where $\alpha'=50$ and $r_{b'}/r_g$ is negligible,

$F_e=1{\cdot}16$ or $0{\cdot}64$ db.

Where $\alpha'=50$ and $r_{b'}/r_g=0{\cdot}1$

$F_e=1{\cdot}28$ or $1{\cdot}60$ db.

The condition (67) gives an optimum value of I_b for a given r_g,

$$I_b=1/\Lambda(r_{b'}+r_g)\sqrt{1+\alpha'} \qquad \ldots\ldots(69)$$

The upper theoretical limit for r_g consistent with low noise figures is imposed by the neglected terms in I_{es} and I_{cs} which give rise to additional terms of the order of ΛZr_gI_{cs}. The lower limit for r_g is imposed by the desirability of keeping Z small. So a smaller transistor is desirable for matching into high resistance sources if other noise sources can be reduced to a sufficiently low level.

5.4. *Calculation of Noise Figure for Grounded Base Configuration*

A similar treatment may be given for the grounded base configuration for which a simplified equivalent circuit is given in Fig. 4.

The symbols are those of the preceding section. In addition

r_e = emitter resistance in grounded base configuration.

α = short circuit current gain in this configuration.

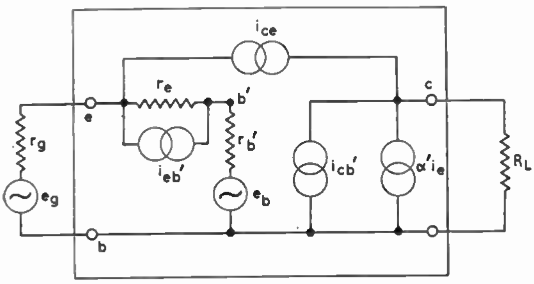


Fig. 4. Equivalent circuit for grounded base.

The short-circuit output currents and increments in noise figure corresponding to the various noise generators are given in the table below.

So the noise figure in the grounded base configuration

$$F_b = Z + \frac{Z^2 r_g}{4kT}\left[\overline{i^2_{b'e}} + \overline{i^2_{ce}}\left\{\frac{1}{\alpha}\left(1-\alpha+\frac{r_e}{Zr_g}\right)\right\} + \overline{i^2_{b'c}}\left\{\frac{1}{\alpha}\left(1+\frac{r_e}{Zr_g}\right)\right\}^2\right] \quad \text{......(70)}$$

Using (55), if only white noise sources are present

$$F_b = Z + \frac{\Lambda Z^2 r_g}{2}\left[I_b + I_{cs} + 2I_{es} + (I_c - I_{cs})\left\{\frac{1}{\alpha}\left(1-\alpha+\frac{r_e}{Zr_g}\right)\right\}^2 + I_{cs}\left\{\frac{1}{\alpha}\left(1+\frac{r_e}{Zr_g}\right)\right\}^2\right] \quad \text{............(71)}$$

If I_{es} and I_{cs} are neglected, putting $r_e = 1/\Lambda I_e$,

$$F_b = Z + \frac{\Lambda I_e r_g}{2}\left[\left(Z+\frac{1}{\Lambda I_e r_g}\right)^2 + \frac{\alpha'}{(\Lambda I_e r_g)^2}\right] \quad \text{............(72)}$$

The minimum value of F_b,

$$F_b = Z\left\{1+\frac{1}{\alpha'}(1+\sqrt{1+\alpha'})\right\} \cong Z(1+1/\sqrt{\alpha'}) \quad \text{............(73)}$$

occurs where

$$I_e = \frac{\sqrt{1+\alpha'}}{\Lambda(r_{b'}+r_g)} \quad \text{............(74)}$$

exactly as for the grounded emitter configuration.

To this approximation the minimum noise figure and the bias conditions required to obtain it are the same in grounded base as in grounded emitter.

Formulae for the noise factor in terms of the intrinsic admittance parameters and extrinsic base resistance are given by Giacoletto[25]. These are valid for white noise for any transistor up to frequencies of the order of the transit time, but are difficult to apply in practice without simplification. At frequencies of the order of the transit time account must be taken of randomness in the generation, recombination and

Noise Generator	Output Current	ΔF
e_g	$\alpha e_g/(r_g+r_e+r_{b'})$	1
e_b	$\alpha e_b/(r_g+r_e+r_{b'})$	r_b/r_g
$i_{b'c}$	$i_{b'c}$	$\dfrac{\overline{i^2_{b'c}}(r_g+r_e+r_{b'})^2}{4kTr_g\alpha^2}$
$i_{b'e}$	$\alpha i_{b'e}'(r_{b'}+r_g)/(r_g+r_e+r_{b'})$	$\dfrac{\overline{i^2_{b'e}}(r_g+r_{b'})^2}{4kTr_g\alpha^2}$
i_{ce}	$i_{ce}\left\{1-\dfrac{\alpha\ (r_{b'}+r_g)}{r_g+r_e+r_{b'}}\right\}$	$\overline{i^2_{ce}}\left\{\dfrac{r_e+(r_g+r_{b'})(1-\alpha)}{4kTr_g\alpha^2}\right\}^2$

diffusion processes in different regions of the diode and transistor. An analysis for a specially simple case has ben given by van der Ziel[26] using a transmission line analogue.

5.5. *Surface and Leakage Noise*

Giacoletto's model has been extended by Fonger[18] to include flicker noise sources. If $G(f,\mathbf{r})$ is the local surface pair generation rate/unit area (compare eqn. (40)) and w_e, w_c are the probabilities of a minority carrier reaching the emitter and collector, the noise currents arising from pair generation in an area dS are given by

$$\left.\begin{aligned} di_{b'e} &= G(\mathbf{r})w_e(\mathbf{r})dS \\ di_{b'c} &= G(\mathbf{r})w_c(\mathbf{r})dS \end{aligned}\right\} \qquad \text{.........(75)}$$

so that the total surface noise generators are

$$\left.\begin{aligned} \overline{i^2_{b'e}} &= q^2\psi(f)\int_E \{\Delta p(\mathbf{r})\}^2 w_e^2(\mathbf{r})dS \\ \overline{i^2_{b'c}} &= q^2\psi(f)\int_C \{\Delta p(\mathbf{r})\}^2 w_c^2(\mathbf{r})dS \end{aligned}\right\} \qquad \text{.........(76)}$$

For the evaluation of these integrals the reader is referred to Fonger's paper. No detailed model of leakage noise has been given; it may be represented by a current generator between the appropriate electrodes, e.g. collector and base.

Fonger has measured the noise in germanium alloy triodes in a number of different ways—as forward and reverse biased diodes, as normal triodes, and as triodes with emitter and collector interchanged. After allowing for modulation of the base resistance by injected carriers, he shows that the above white surface and leakage noise generators are consistent with a wide range of data. Some other plausible models of transistor noise are excluded by the data. Some transistors were found where surface noise was the only source of flicker noise. This surface noise increased with current roughly as $I^{3/2}$. It was almost independent of collector voltage; in fact feedback through the widening of the space charge with collector voltage led to a slight decrease in noise figure with voltage.

6. Methods of Measurement

With the improvements in transistor noise, considerably greater care must be exercised in the design of amplifiers to measure it. Indeed, the direct measurement of the equivalent noise current and voltage generators is often virtually impossible. We shall consider here methods for measuring the noise figure. Typically, we require a battery-driven pre-amplifier, a wide-band amplifier, a selective amplifier defining a narrow band or bands, or a well-defined wide-band, and a detector. Furthermore, some method of determining gain and bandwidth, or at least their product, is required. The amplifiers should be linear for a sine wave input up to three times full scale deflexion of the detector in order to avoid clipping the noise peaks. Non-linearity is particularly undesirable when noise and sinusoids, or Gaussian and non-Gaussian noise, are being compared.

Below are listed four different methods of measuring noise figures, according to the methods used for measuring the gain and bandwidth.

6.1. *Use of Stable Amplifier*

Using a small, known input signal of variable frequency the gain $B(f)$ of the measuring amplifier is determined directly. The gain of the transistor (and where necessary, its variation with frequency over the pass-band of the amplifier) is determined at a conveniently high signal level. The noise bandwidth may then be calculated from eqn. (41). The output noise of the transistor is measured and the noise figure calculated from (39) and (40). Using this method an amplifier of stabilized bandwidth is essential, and stabilized gain is extremely desirable. However, as the lowest noise figures in valve amplifiers are obtained without negative feedback, long-term gain stability in the wide-band pre-amplifier may not be attainable. The method is most suitable for wide-band measurements of noise, e.g. as a production check, when the bandwidth is fairly easily defined and accuracy in assessing very low noise figures may not be needed.

6.2. *Use of Noise Diode to Measure the Amplifier Gain Bandwidth Product*

This method differs from that in 6.1 in that long-term stability of the amplifier is not needed. To determine the gain and bandwidth product, the current noise from a saturated tungsten filament diode (noise diode) is injected into a resistance R in the input of the measuring amplifier. As the noise diode forms a high impedance current source, the voltage V across the input of the amplifier is given from (17) and (18) by

$$\overline{V^2(f)} = 2qIR^2 + 4kTR \qquad \ldots\ldots(77)$$

where q = electron charge.
I = diode current.
$\overline{V^2}$ = mean square voltage at the amplifier input terminals.
R = net resistance across the input.

The additional mean square output voltage is then $2qIR^2A^2\Delta f$ and hence $A^2\Delta f$ is determined. It is also possible to calibrate the amplifier using thermal noise alone; however, the noise diode method allows a wider range of calibrated input noise, though it is open to the objection that flicker noise appears at very low frequencies, even in tungsten filament diodes. The output noise of the transistor and the transistor gain are measured as in the method of 6.1.

6.3. *Calibration by Injection of Diode Noise at Transistor Input*

This method is probably the most accurate and is quite convenient in practice. It does not require the stabilization of amplifier gain or bandwidth. Noise from a noise diode is injected into the input resistance R of the transistor. Then eqn. (77) gives the value of the noise voltage generator in series with the resistance. The mean square output noise voltage in the bandwidth Δf of the amplifier is

$$(2qIR^2A^2 + 4kTRA^2F)\Delta f \qquad \ldots\ldots(78)$$

while without the noise diode it is $4kTRA^2F\Delta f$. The ratio, M, of these quantities with and without the noise diode is

$$M = 1 + qIR/2kT \qquad \ldots\ldots(79)$$

So

$$F = \frac{qIR}{2kT(M-1)}$$

Typically, the transistor output noise is measured in arbitrary units; then the amplification is reduced by 3 db and the diode current increased to give the same reading on the output meter. Here $M=2$,

so $F = 20IR$ at $T = 290°K$.

6.4. *Calibration by Known Small Signal at Transistor Input*

If the noise diode of the preceding method is replaced by a signal generator of known output, the noise figure may be similarly determined provided the amplifier bandwidth is known. The output power is the sum of the output noise and signal power, so if a square law detector is used, the output reading is proportional to $(\overline{V_e^2} + 4kTRF)A^2\Delta f$ with the sine wave input and to $4kTRFA^2\Delta f$ without it, where V_e is the equivalent sine wave input voltage referred to a generator in series with the input resistance. Then, if M is the ratio of the two output readings

$$M = 1 + \frac{\overline{V_e^2}}{4kTRF\Delta f} \qquad \ldots\ldots(80)$$

If a linear rectifier and d.c. meter is used as a detector, graphs of the response of such a detector to the sum of sine waves and Gaussian noise may be used[27].

6.5. *Measurement of Noise Figures near Unity*

When transistors are measured whose noise level is low, particularly if the voltage gain is also low, noise in the measuring amplifier and thermal noise in the load become important. If the output impedance of the transistor is high compared with the load, the combined noise power due to the amplifier and thermal noise in the load may be subtracted from the transistor noise. If the output impedance is comparable with the load it shunts the noise generator associated with thermal noise in the load. In the limit when the output impedance is much less than the load, the thermal noise in the load is shunted by the transistor output impedance and thus only the short-circuit amplifier noise should be subtracted from the measured transistor output noise. We shall consider the accuracy of these approximations using the method of Sect. 6.3.

Referring to Fig. 5,

Z_o = the output impedance of the transistor (supposed purely resistive).
e_Z = the open-circuit output voltage of the transistor in one cycle bandwidth.
R_L = the load (purely resistive).
e_L = the thermal noise voltage in series with the load per cycle.
e_o = the output transistor noise voltage across the load per cycle.
e_A = the amplifier noise voltage referred to the grid per cycle.
e_g = the noise voltage at the grid per cycle.

We suppose that the measuring amplifier has infinite impedance.

Clearly $\qquad e_o = e_Z/(Z_o + R_L)$

Now, using (13), (38) and (78),

$$\overline{e^2_g} = \overline{e^2_L}\left(\frac{Z_o}{R_L+Z_o}\right)^2 + \overline{e^2_o} + \overline{e^2_A}$$

$$= 4kTR_L\left(\frac{Z_o}{R_L+Z_o}\right)^2 + 2qIR_gA^2 + 4kTR_gA^2F + \overline{e^2_A} \quad \ldots\ldots(81)$$

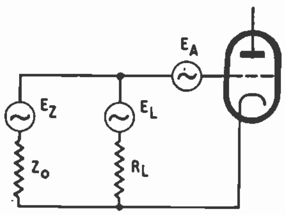


Fig. 5. Equivalent input circuit for noise figure measurement.

If S is the reading on a square law output detector such that

S_{00}—amplifier short circuited,

S_0—amplifier with load R_L, transistor disconnected,

S_1—amplifier with load R_L, transistor connected,

S_2—amplifier with load R_L, transistor connected, and current noise injected from noise diode,

define $\quad M = \dfrac{S_2 - S_0}{S_1 - S_0} \quad$ and $N = \dfrac{S_2 - S_{00}}{S_1 - S_{00}}$

A little manipulation gives

$$F = \frac{qIR_g}{2kT(M-1)} - \frac{(R_L + 2Z_o)R^2_L}{R_g(R_L+Z_o)^2A^2} \quad \ldots\ldots(82)$$

$$F = \frac{qIR_g}{2k\,T\,(N-1)} - \frac{R_LZ_o^2}{R_g(R_L+Z_o)^2A^2} \quad \ldots\ldots(83)$$

The second term is negligible in (82) if $R_L \ll Z_o$ and in (83) if $R_L \gg Z_o$. The first condition can often be obtained with transistors under normal operation and the second with hushed transistors. The noise figure may also be determined without restricting or measuring Z_o by using two noise diodes, one at the input of the transistor and the other at the input of the measuring amplifier. The method is described by van der Ziel[4] (p. 78).

6.6. *Choice of Bandwidth for Measuring Amplifier*

The choice of bandwidth Δf is, of course, bound up with the choice of a detector. The error in a single reading of noise power with an instrument of time-constant τ is $(2\Delta f\tau)^{-\frac{1}{2}}$. The bandwidth should be made as large as possible without distorting the noise frequency spectrum.

If it is assumed that the noise frequency spectrum is of certain types commonly encountered, certain filters distort the spectrum far less than others of the same bandwidth; a useful filter whose bandwidth is approximately the same as the spot frequency at which the noise is required is given by van der Ziel[4]. It is suitable for noise with a f^{-n} frequency dependence, with $0<f<2$. The only advantage of high Q filters is that it is possible to reduce hum.

6.7. *Square Law Detectors*

Most accurate measurements require the use of a square law detector. These may be divided into two classes, thermal detectors and other quadratic detectors. In thermal detectors, such as the thermocouple and thermistor, the heating effect of the input current is measured by some temperature sensitive device. As the heating is proportional to the square of the instantaneous current, the output from a thermal detector is practically always a function only of the mean square current input, though not necessarily a linear function. It is therefore insensitive to changes in waveform. The thermocouple millivoltmeter is a convenient ready-made instrument though its time-constant is short for accurate measurement of spot noise figure at low audio frequencies: it is also liable to burn out if overloaded. The time-constant may be increased by replacing the millivoltmeter by a heavily damped galvanometer. The time-constant of thermocouples is of the order of seconds, as required at low audio frequencies. An indirectly heated vacuum fitted thermistor in a bridge circuit gave an output which was quadratic within 3 per cent. up to 200 millivolts. Thermal drift was of the order of two millivolts without special precautions; it can be overcome by using a constant temperature bath.

A large power is needed to drive most thermal detectors. This disadvantage is not possessed by the other quadratic detectors which depend on obtaining an output signal proportional to the input signal from a non-linear device. The chief forms are the thermionic valve and the crystal detector; in both cases care must be taken that the device is not driven out of the quadratic region on peaks. The thermionic valve is also liable to drift out of the quadratic region if the operating point is not stabilized. The response time of the quadratic element itself is negligible here, so the time-constant is determined by the linear measuring instrument (e.g. rectifier and d.c. milliammeter). It can, however, be increased by capacitance-resistance smoothing.

7. Variation of Noise with Operating Parameters

7.1. *Frequency*

The combination of a more or less constant level of white noise with a quantity of flicker noise which varies greatly from unit to unit gives rise, under normal bias conditions to the variation in frequency sketched in Fig. 6. Curve A, typical of most early transistors, is dominated by flicker noise. Curve B shows the ordinary modern transistor with a noise figure about 10 db at 1 kc/s. Curve C typifies the good, low-noise transistor where the white noise region extends down to the few hundred cycles. The rise in noise figure at high frequencies is associated with loss of gain and transit time effects, and will not be discussed here.

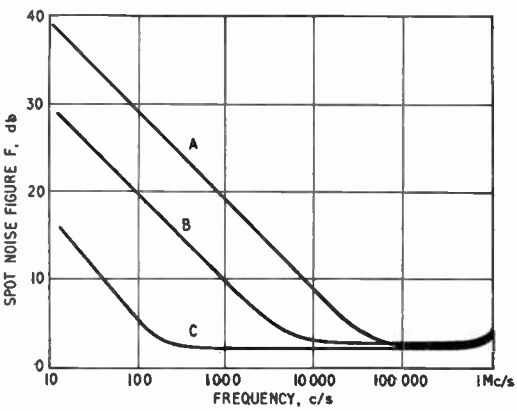
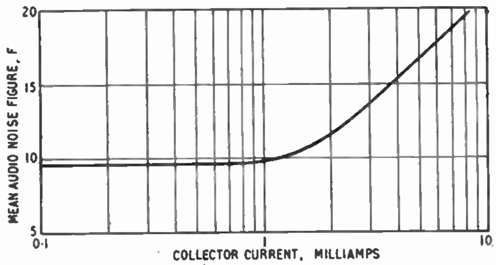
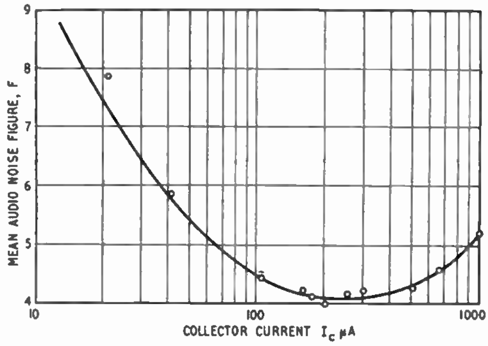


Fig. 6. Variation of noise figure for three grades of transistor.



(*a*)



(*b*)

Fig. 7. Variation of noise figure with collector current.
(*a*) Transistor with leakage and surface noise.
(*b*) Transistor without surface noise.

7.2. *Collector Voltage, V_C*

The variation of noise with collector voltage is quite different for transistors of these different grades. The noisy transistor typified by curve A usually shows a great increase in noise with collector voltage: various sources give this increase as between 5 and 10 db per octave of voltage or again as 2 db/volt, superimposed on a constant background of noise broadly associated with the emitter. This noise is, however, extremely erratic, being non-cyclic under changes in voltage and temperature and not constant with time. The transistor typified by C, being dominated by white noise rather than leakage noise, shows a level of noise nearly independent of collector voltage. Transistors with flicker noise at the level of curve B may show some increase in noise with collector voltage, but more commonly surface noise is dominant and the noise is nearly independent of collector voltage. A slight decrease in the noise figure with collector voltage is sometimes observed; this can be associated with space charge widening.

7.3. *Collector Current, I_C*

The noise behaviour on varying the collector current is exemplified by Fig. 7, which shows the mean noise figure measured by an amplifier of constant response in the range 200 c/s—10 kc/s, and virtually zero response elsewhere. The noise from the transistor of Fig. 7 (*a*) is virtually independent of current up to 1 mA. Such a transistor exhibits both leakage and surface noise, the leakage noise predominating up to 1 mA. In the transistor of Fig. 7 (*b*) leakage noise is low or absent and there is a broad minimum as predicted by (66). The solid line is of the form

$$F = A(I_c + 1/I_c) + B \qquad \ldots\ldots(84)$$

predicted by that equation. The deviations at low bias currents are expected owing to the neglect of saturation currents. The fit does not imply that surface noise is entirely absent however. Transistors which do not show an increase in noise with collector voltage are often quite stable, and it is reasonable to look for, and work at, the minimum in the noise figure vs. collector current curve.

7.4. *Generator Resistance R_g.*

The variation in noise figure with generator resistance for low-noise transistors is also in accord with eqn (66) which predicts a broad minimum of the form

$$F = C(R_g + 1/R_g) + D \qquad \ldots\ldots(85)$$

The minimum is typically at about 500 ohms for lower power audio transistors. In noisy transistors the minimum is less marked, though the increase of noise with generator resistance is finally steeper.

7.5. *Choice of Bias Point in Low-Noise Amplifiers*

The considerations of the previous sections suggest that for ordinary transistors an operating point such as ($I_c = 300\mu A$, $V_c = 0{\cdot}5$V) is generally desirable, but an increased collector voltage may be permitted for transistors with low leakage noise. Other factors affecting the choice of bias will be the signal-handling requirements of the stage and the temperature variations likely to be encountered.

The erratic behaviour of units with leakage noise, and the large peaks occurring in it, make their presence extremely undesirable in low-noise amplifiers. Surface noise must sometimes be regarded as a necessary evil at present,

7.6. *The "Hushed" Transistor*

A technique known as hushing has been introduced by Volkers and Pederson[28] to minimize leakage noise by running the transistor with zero or slightly injecting (e.g. 10 mV) collector bias. In this way they obtained noise figures from ordinary transistors comparable with those attainable from low-noise transistors. Curves given by them show an improvement on normal operation of about 7 db in flicker noise and a white noise component little changed. There is a considerable reduction in gain, but this is partly alleviated by a reduction in output impedance so that interstage transformers need not be used. The stabilization of the bias point is troublesome, as small changes in bias cause enormous changes in gain. Owing to non-linearity the technique cannot be used if the stage is required to handle large signals.

8. Leakage Noise and Leakage Current

The experiments below on an unencapsulated transistor illustrate the irregular behaviour which may be expected under unfavourable conditions from transistors where leakage noise is present. The mean audio noise figure was measured for the band 200 c/s—10 kc/s in conjunction with measurements of the collector current with emitter open circuited, I_{co}.

Row I shows the measurements taken on a freshly etched transistor exposed to air of 38 per cent relative humidity (R.H.). On exposure to air saturated with water vapour I_{co} and $\bar{F}$ rose within a second to the values shown in the second row. The increase in noise was associated with pop and crackle, that is, large irregular peaks. This noise, when attenuated to

	I_{co} (microamps)			$\bar{F}(I_c = 1\text{mA})$ db	
Collector Voltage V_c	3	10	25	3	10
I	5	6	7	9	9
II	v. large	irreg.	~50	25	35
III	10	12	14	16	35
IV	4·5	6	6	9	9

give the same mean output level as the first, was much more objectionable to the listener. The change between stages I and II was at first reversible, but during a large number of humidity cycles $\bar{F}$ and I_{co} rose to the values shown in row III at 38 per cent. R.H. On drying at 100°C. and coating with a silicone resin, $\bar{F}$ and I_{co} fell to the values in row IV, measured in air of 33 per cent. R.H. On exposure to air saturated with water vapour the noise rose slowly and in an irregular manner dependent on the power dissipation in the unit; in a few days $\bar{F}$ and I_{co} had reached high values.

From observations such as these it appears that electrolytic conduction on the surface and the noise associated with it occur at about 40 per cent. R.H. (or considerably less if there is a large number of absorbed ions on the surface) and increase rapidly with humidity. Whenever moisture is present in the can, under suitable temperature cycling sufficient water may be absorbed on the surface at the junction to give electrolytic conduction. Any layer more than two molecules thick is said to be sufficient. The effect of resins is only to slow down the rate at which water reaches the junction. It seems reasonable to ascribe some at least of the irregularities in leakage noise mentioned above to the movement of water vapour and the transport of ionic impurities near the junction. The reduction and elimination of leakage noise in manufacturers' current products must be associated with electrolytic etching and adequate washing to remove ions, and the exclusion of water vapour, both during encapsulation and from the components of the transistor. A typical danger is that water may be temporarily trapped in the plating on a plated can.

In this connection it is interesting to ask how far the observed leakage current is correlated with the noise figure. We define

$$\Delta I'_{co} = I_{co}\ (V_c=10) - I_{co}(V_c=3)$$

$$\Delta I''_{co} = I_{co}\ (V_c=25) - I_{co}(V_c=10)$$

as measures of the leakage current. The histograms of Figs. 9 and 10 show the occurrence of various ranges of $\Delta I'_{co}$ and $\Delta I''_{co}$ for different values of the mean noise figure measured at (I_c=1mA, V_c=10 volts) in a 200 c/s—10 kc/s band. There is a tendency for large values of $\bar{F}$ to be associated with large leakage currents; this is most marked in the histograms of $\Delta I''_{co}$. Exceptions to this trend are numerous so that a method of selection for noise based on measurement of leakage current would be unreliable

e.g. of the 160 diodes for which $\bar{F} \leq 8$, 32 per cent. had $\Delta I''_{co} > 0{\cdot}95\mu A$;

of the 153 diodes for which $\bar{F} > 8$, 44 per cent. had $\Delta I''_{co} < 0{\cdot}95\mu A$.

The correlation of $\bar{F}$ with I_{co} is even less significant. Large values of $\bar{F}$ with low leakage currents may be due to surface noise, while large leakage currents with low noise may be due to surface inversion layers. Possibly a more detailed study of the form of the I_{co} vs. V_c curve would enable a distinction to be made between various sources of leakage current.

9. Comparison of Transistors with Valves

The noise figures obtainable with thermionic valve audio amplifiers are lower than those yet reached using transistors. These low-noise figures are only obtained with generator impedances very much greater than those giving the minimum noise figure with a transistor amplifier. For this reason the claim is sometimes made that transistors are less noisy than valves because in terms of the noise voltage referred to the input terminals, good transistors are less noisy than valves. The comparison should of course be made in terms of power.

Owing to the high input impedance of valves it is convenient to define a parameter R_n, the equivalent noise resistance, such that

$$F = 1 + R_n/R_g$$

R_n is independent of R_g over a wide range. The ultimate limit in valves is also determined by shot noise which contributes about $2{\cdot}5/g_m$ (where g_m is the transconductance) to R_n in a triode. It is typically about 1·5kΩ. This is increased by partition noise in pentodes, the use of which is inadvisable in low-noise stages for that reason. Noise resistances measured in a number of commercial low-noise valves at 1 kc/s were in the range 6—20 kΩ; at 1 kc/s and below, flicker noise is dominant. The corresponding noise figures for a generator resistance of 1 kΩ are 8—14 db, but for high generator resistances the noise figure is very near unity (0·25 db for R_n=6 kΩ, R_g=100 kΩ).

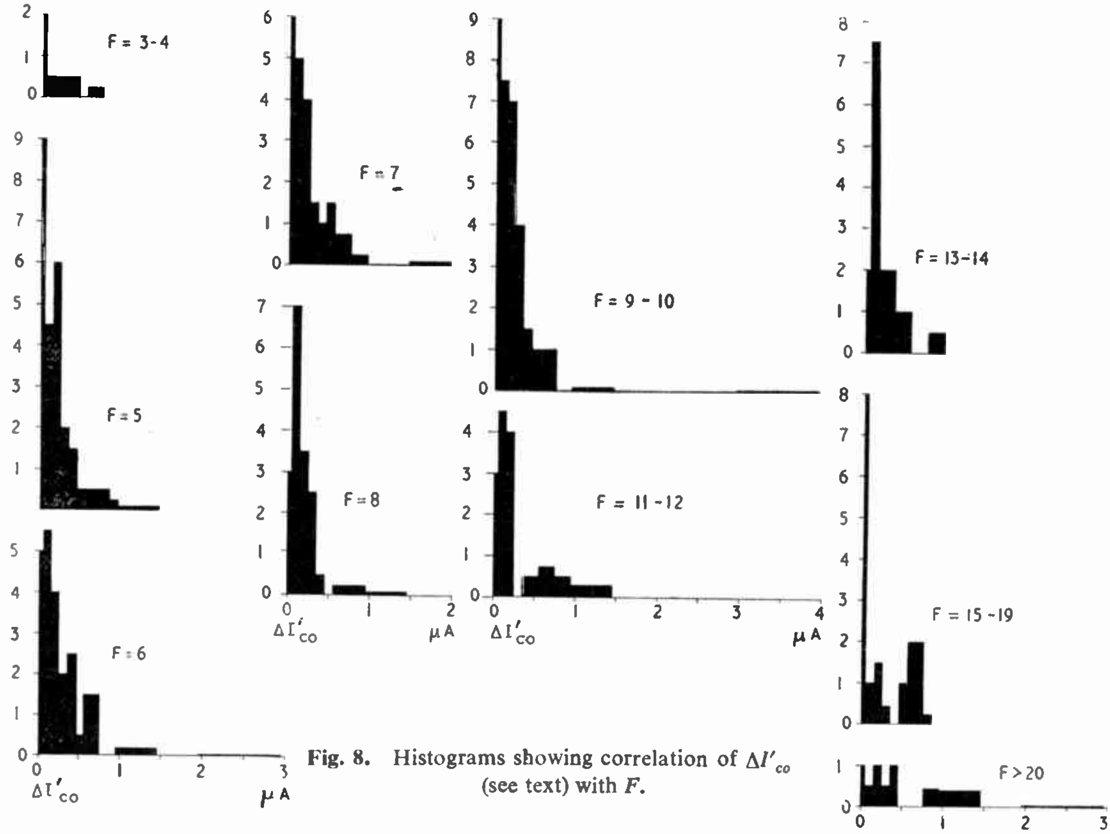


Fig. 8. Histograms showing correlation of $\Delta I'_{co}$ (see text) with F.

By using a transformer these low noise figures may be approached with smaller source resistances. If a transformer has a turns ratio n and the primary and secondary windings have resistances R_p and R_s, the valve sees a generator of resistance n^2R_g in series with a resistance $nR_p{}^2+R_g$ so that the noise figure of the valve and transformer, assuming only thermal noise in the transformer, is

$$F=1+\frac{R_n+R_s+n^2R_p}{n^2R_g}$$

Thermal noise contributes about half a decibel in very good commercial transformers with n between 30 and 300. This means that the noise figures actually obtained with valves at 1 kc/s are comparable with the best theoretically attainable with transistors, even for source resistances around 1 kΩ.

In actual practice, where the last few decibels may be unimportant, the transistor has a number of advantages at low signal levels. Hum pick-up from the heaters and microphonics, both troublesome in valves, are absent in transistors. With low generator resistances, the elimination of the transformer required with valves means a saving in weight and expense, and avoids the possibility of pick-up by the transformer.

10. Acknowledgments

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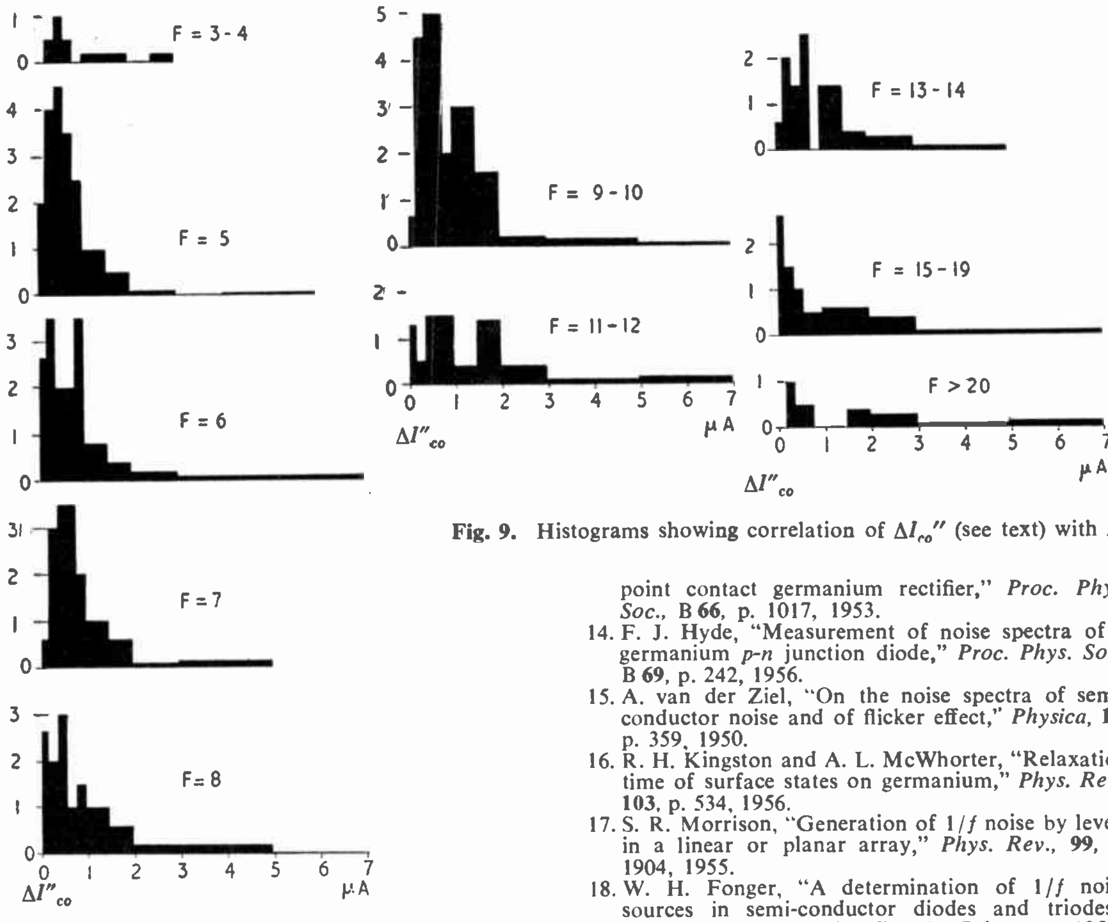

Fig. 9. Histograms showing correlation of $\Delta I_{co}''$ (see text) with $\bar{F}$.

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INSTITUTION NOTICES

Obituary

The Council has learned with regret of the deaths of the following members and has expressed sympathy with their relatives.

Alexander Elliot Macintosh (Associate Member) died on October 4th last at the age of 65 years.

He was Principal of the Glasgow Wireless College, a position which he had held since the end of the 1914-18 War. He was also in charge of similar establishments in Belfast and Aberdeen, concerned with the training of radio officers for the Mercantile Marine and civil aviation.

Mr. Macintosh had been connected with radio communications since the early days of the first World War, when he served with the Wireless Section of the Royal Engineers. He was elected an Associate Member in 1943.

* * *

Henry George Silver, M.B.E. (Associate Member) died on the 4th January at the age of 58. Mr. Silver was an Assistant Director Engineer in the Directorate of Aeronautical Inspection, Ministry of Supply. He first joined the Directorate in 1923, and as an assistant inspector was closely concerned with the expansion of the recruitment and training of staff, and equipment for stations, during the immediate pre-war period. At the beginning of the war he helped to evolve the test organization for the production of radar transmitters, and he subsequently served on the headquarters staff of the Directorate. His work was recognized by his appointment in 1947 as a Member of the Most Excellent Order of the British Empire.

Mr. Silver, who was elected an Associate Member in 1947, served for a number of years on the Institution's Membership Committee.

New Zealand Department of Education

The New Zealand Department of Education has recently stated that Associate Membership of the Institution is accepted as a professional qualification for teachers, provided that it has been obtained by examination since May, 1951.

This recognition now places teachers in New Zealand on the same footing as teachers in this country under the Burnham Committee award of 1951. More recently, the Burnham Committee has accepted Associate Membership of the Institution after 1951, without reservation.*

Professor S. K. Mitra, F.R.S.

Members of the Institution in general, and those in India in particular, will learn with pleasure of the recent election of Professor Sisir Kumar Mitra, D.Sc. (Member) as a Fellow of the Royal Society.

In the announcement of his election at a meeting of the Society on March 20th, the citation states that Professor Mitra is "distinguished for his researches in many branches of upper atmosphere physics."

Professor Mitra occupies the Chair of Physics at Calcutta University, and he is head of its Institute of Radio Physics and Electronics. He has been Chairman of the Calcutta Section of the Institution since its formation in 1952.

A fuller note of his career appeared in the *Journal* for March 1953.

North-Eastern Section

A Dinner is to be held by the North-Eastern Section in Newcastle-on-Tyne on Wednesday, May 21st, at which the President of the Institution will be the Guest of Honour.

Further information and tickets may be obtained from the Local Secretary, Mr. L. G. Brough, 61 Whinneyfield Road, Newcastle.

Waverley Gold Medal Essay Competition

The 1958 Waverley Gold Medal and £100 will be awarded by the Scientific Advisory Board of the journal *Research* for the best essay of about three thousand words based on some recent scientific research or new development. The essays are intended to indicate scientific background, experimental results and potential application in industry, and should be clearly intelligible to a scientist engaged in other fields or a director of an industrial firm.

A second prize of £50 will be awarded and an additional prize, also of £50, for the best entry from a competitor under 30 years of age. The closing date for entries is 31st July 1958; further information may be obtained from the Editor, *Research*, 4/5 Bell Yard, London, W.C.2.

* *J.Brit.I.R.E.*, 17, p. 192, April 1957.

AUTOMATIC METHODS IN RADIO COMPONENT MANUFACTURE*

by

D. Stevenson† and R. B. Shepherd, B.Sc.(Eng.)†

A paper presented at the Convention on "Electronics in Automation" in Cambridge on 29th June, 1957.

In the Chair: Mr. L. H. Bedford, C.B.E. (Past President)

SUMMARY

Part 1 describes a method of controlling a high-speed coil winder by an electronic counter. A specialized machine which winds single-layer coils is described.

Part 2 deals with the development of a tool protection device for use with automatically loaded presses. The device detects the piece-part magnetically as it is ejected and so clears the interlocks to allow the press to continue operating.

Historical Note

Over the past five years the Production Engineering Department of the Company with which the authors are associated has been engaged in the rationalization of production methods in the light of modern developments in the fields of electronic control and servo-mechanisms.

This development has been given fresh impetus by introduction of printed circuit techniques which lend themselves readily to fully mechanized assembly.

It is, however, with two less spectacular developments that this paper is concerned, namely the design and construction of a fully automatic high speed coil winding machine and the solution of an interlocking problem encountered during the mechanizing of a standard single-acting power press.

Part 1—ELECTRONIC CONTROL APPLIED TO COIL WINDING

by R. B. Shepherd.

1. Introduction

This part of the paper describes a method of obtaining accurate control of high-speed machines by electronic counting methods. An application to a specialized machine will be dealt with, which winds single-layer coils of wire of between 20 and 30 S.W.G. for use in the radio-frequency stages of radio and television sets. Reference will also be made to the control of multi-wave-winding and layer winding machines.

Small radio-frequency type coils of from 2 to 40 turns were previously wound on simple manually-operated machines. Turns were registered on a mechanical counter, while the operator turned a handle driving the mandrel. When the correct number of turns had been wound, the coil ends were cut to length by the operator and the coil removed. More wire was then pulled from the reel, attached to the mandrel by wrapping round a projection, and winding recommenced.

The development of a general purpose electronic control unit has enabled a fully automatic machine to be designed to produce the coils required. This machine feeds wire from the reel to a motor-driven mandrel, whose turns

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are controlled by an electronic binary counter. On completion of the exact preset numbers of turns, a cutter severs the wire and the coil is ejected, completing the cycle.

2. Design Problems

Before attempting a fully automatic machine, an intermediate-stage winder was designed, the mandrel being turned by a small electric motor and the turns counted on a telephone type uniselector. Regenerative braking was used to stop the motor on completion of the count, the operator merely having to provide the wire feed, and to cut the coil ends.

The winding speed of the semi-automatic machine was low, owing to the low rate of count obtainable from an electro-mechanical type of counter. The fully automatic machine was therefore designed to fulfil the following conditions:

(1) Several thousand spring type coils are required per day, and if these are to be wound on a single machine this entails well over a million operations of the counter a year.

(2) The winding speed must exceed 1500 rev/min. (or 25 counting impulses a second) to attain the high output required.

(3) The counter must be easily preset, preferably by rotary switches.

(4) The counter must reset in a fraction of a second, between winding cycles.

(5) Absolute accuracy in counting is essential, the maximum tolerance being $\pm\frac{1}{8}$th turn.

(6) Low cost is extremely desirable.

To simplify switching, a series of three stages was chosen to build into a chassis unit with a rotary type switch for presetting the required count from 1 to 8 digits. This means that one unit will give an output pulse after it has received the number of pulses set on its rotary switch, and, thereafter, every eight input pulses. Two units in series will thus count up to 64, switching 1 to 8 on the first unit and 0—8—16 etc., on the second. Scale readings when setting up a number are additive. The circuit is shown in diagrammatic form in Fig. 1.

Mechanical and electro-mechanical types of counter as tried on the semi-automatic machines were found to be incapable of dealing with high winding speeds, and experience on these machines showed that frequent maintenance was necessary even at low speeds.

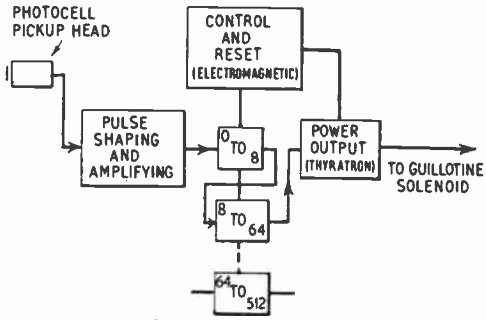

Fig. 1. Block diagram of counter. Counter units comprise 3 binary stages each.

An electronic counter has, therefore, been developed, which, in conjunction with a thyratron output stage, provides a reliable solution to the problems involved. The complete equipment is shown in Fig. 2.

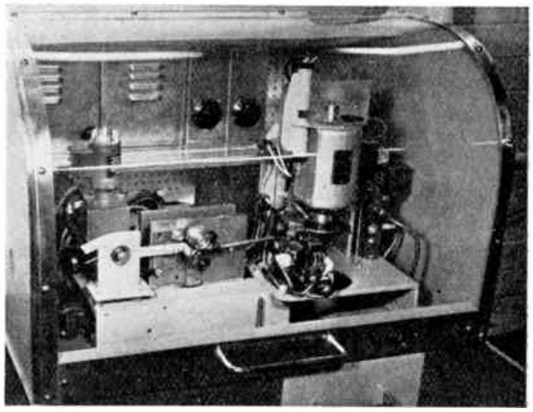

Fig. 2. The coil winder and its automatic control equipment.

3. The Electronic Counter

A binary circuit provides one voltage output pulse for every two input pulses received. A straightforward series of x such stages provides an output pulse for every 2^x input pulses. Although it would be possible to arrange binary units to count in decades by means of internal feedback, this would need more stages per unit.

4. Accuracy of Control

Having constructed an accurate counter to perform satisfactorily at high speed, it remained to make the output pulse perform some operation to control exactly the number of turns wound on the coil. Rapid breaking of the motor was again attempted, but a variable overrun of about half a turn seemed unavoidable, so it was decided to control the number of turns wound by cutting the wire while it was being wound.

The output pulse from the electronic counter is fed to a miniature thyratron. This energizes the solenoid operating a small guillotine, cutting the moving wire with a delay of only a few milliseconds.

The total time which elapses between the output pulse and the operation of the cutter is therefore equivalent to a very small fraction of a turn on the coil being wound, even at high winding speeds.

5. Conclusion (Part 1)

The machine described produces 30,000 coils a week, and is capable of over double this output. About 10 operators used to be employed part of their time in handwinding these coils, but the machine now produces automatically all the coils needed and enables them to spend their time in finishing the assembly.

Each unit in the counter costs £15 approximately, i.e. the total cost of 4 units counting up to 4,000 is £60, excluding the input and output stages. The cost could be reduced considerably by mass production methods.

The counter has been designed as a general-purpose device, with additional plug-on units and a photocell input stage. All that is necessary on any machine that it is desired to control is a shutter and a means of using the output pulse.

Application has been made to multiple wave-winders to increase range and accuracy, and to layer winding machines to enable the machine to be made automatic.

Part 2—AN ELECTRO-MAGNETIC PROTECTIVE DEVICE FOR AUTOMATICALLY-FED POWER PRESSES

by D. Stevenson

6. Introduction

During the course of modernizing the Press Shop, the decision was taken to fit a mechanical hand to one of the standard single acting power presses. This particular press was being used almost entirely for the manufacture of ferrous metal parts.

The mechanical hand consists basically of a power-driven oscillating arm mechanism which is synchronized with the operation of the press by means of a microswitch-operated electro-pneumatic valve. The function of this valve is to control the supply to the air cylinder which operates the clutch-actuating rod. Pick-up of the piece-part is obtained by suction, and ejection by a jet of air. The operation and layout can be seen in Fig. 3.

7. Design Problems

Air ejection unfortunately is never completely positive and has the inherent danger in automatically fed power presses that, if ejection does not take place, a second piece-part is fed and the press tool may suffer severe damage.

The problem, therefore, resolves itself into the provision of a suitable positive, robust detector, the output of which when amplified (if required) can be used to control the operation of the clutch actuating rod by switching the air cylinder control valve.

The first detecting device considered for this purpose was a simple pressure operated micro-switch arrangement, the pressure being derived from the impact of the piece parts after ejection. This was soon abandoned as being even less positive than air ejection.

The next obvious detector was a photo-electric cell but here again the area ratio of the smallest piece part to the ejection aperture was such that either many individually closely spaced parallel beams of light had to be used or alternatively a light curtain had to be obtained by mirror reflections from a single light source. This did not appear at the time to be a very attractive proposition and was not proceeded with any further.

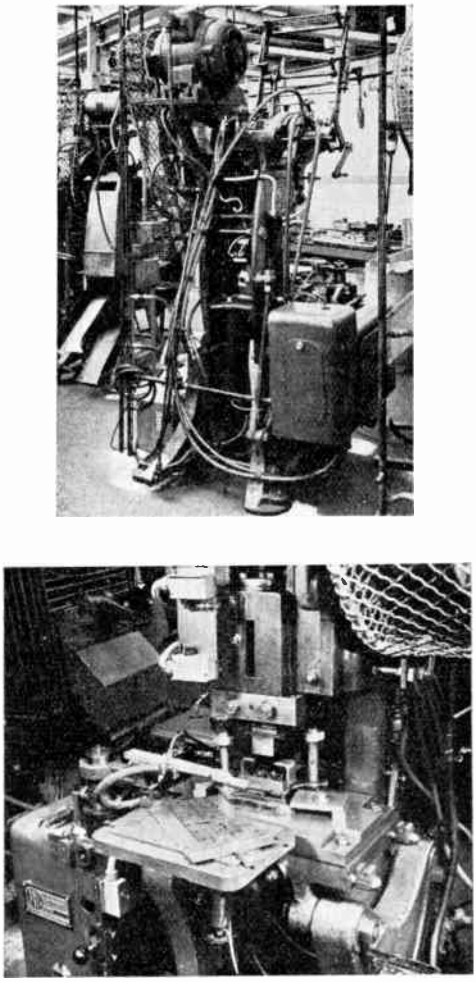

Fig. 3. Press showing mechanical arm. In the lower illustration the coil can be seen at the back and the contacts are on the left. The top photograph shows a rear view with the coil opening just visible.

As the problem was almost wholly concerned with ferrous metal parts it became obvious that the passage of these could be used to produce a change of flux in the magnetic field of a coil and to derive a voltage in accordance with the fundamental law $N(\mathrm{d}t)$, from the second winding on this coil.

Simple experiments utilizing a standard transformer, a coil, an accumulator and an oscilloscope, soon showed this method of detection to be feasible.

8. Design of Equipment

8.1. *Detecting coil former.*

The window size of the coil former was governed by the size of the press throat and had to be such that it did not impose any further restrictions on the passage of the piece-parts. The depth of the coil was governed by considerations of flux distribution. This implied that the coil field should remain unaffected by the movement of the press. Consideration of the theoretical field produced by coils of varying depth enabled a suitable practical depth to be predicted and this was confirmed by fitting an experimental coil to the press and checking for spurious pulses during normal press operation.

8.2. *Windings*

The function of the primary is to produce a magnetic flux throughout the whole window area and should therefore be designed in conjunction with the power supply. From experiments it was found that 400 ampere-turns were sufficient to provide an adequate flux in the particular coil being considered.

Since valve rectification was almost obligatory, the design was based on a suitable valve which was readily available, namely a 5V4G, giving 80mA at approximately 300V. The number of turns required on the primary was then 4250 turns. It was found from wire tables that 38 S.W.G. wire gave a suitable value.

The number of secondary turns was limited by the remaining coil space and minimum diameter of wire which could reasonably be wound. 38 S.W.G. was again chosen and 12000 turns were wound in the remaining space.

8.3. *Feedback*

Unfortunately ripple from the primary supply when induced in the secondary and amplified was sufficient to trigger the switching valve and a feedback circuit had to be used. This was accomplished by the use of a potentiometer in primary circuit which provided an adjustable voltage for injection into the secondary circuit to oppose the voltage induced by the ripple.

It was found, however, that the value of resistance required was too large and a compromise was effected by connecting a capacitor across the primary coil. The use of this additional smoothing capacitor reduced the resistance required to a practical value. Since some ripple was still being passed a sensitivity control was fitted in the amplifier to prevent triggering of the flip-flop by unwanted signals.

8.4. *Amplifier and output stage*
The amplifier and output stage used was one which had been originally designed for use with photo-electric cells. It consisted of a two-stage amplifier using a 6AM6 pentode and a 6AT6 triode feeding a 12AT7 flip-flop circuit.

Modifications were required to prevent oscillation and to reduce the attenuation produced by the input stage required for photo-electric application. The amplifier output was fed via the sensitivity control to the flip-flop output switching stage.

The operation of the circuits can be seen from Fig. 4.

It should be emphasized that the design of the electronic equipment is essentially an improvised one and no claim is made for any special features. The policy has always been

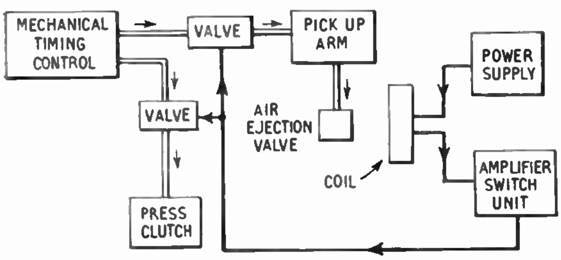

Fig. 4. Diagram showing operation of circuits.

adopted, from the point of view of expense and the limited development effort available, of adapting from existing equipment rather than designing from scratch for optimum performance.

8.5. *Switching and resetting circuits*
To understand the operation of this part of the equipment reference should be made to the switching and resetting schematic (Fig. 5) and the photographs (Fig. 3).

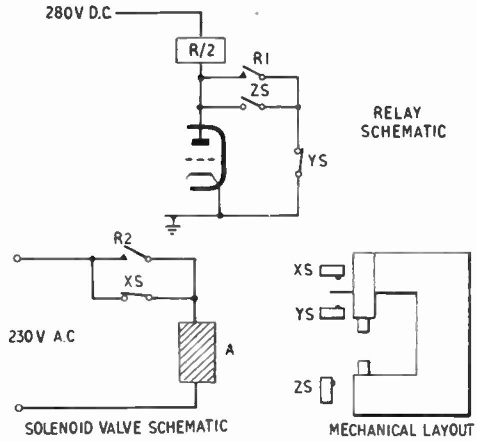

Fig. 5. Switching and resetting circuits.

The operating sequence is as follows:—

(1) Assume press to be at the top of its stroke with relay R de-energized and XS mechanically opened.

(2) To start press, operate resetting switch ZS which energizes relay shutting R/1 which locks in relay and R/2 which energizes air valve.

(3) Press descends and YS is opened at the bottom of the stroke, de-energizing relay R and opening contacts R/1 and R/2. The air valve remains energized, however, as XS is no longer mechanically opened.

(4) If no pulse is given to operate the relay, the press returns to the top of its stroke and remains there since R/2 and XS are both opened.

(5) If a pulse is given, R is again energized and even after the press reaches the top of its stroke, it continues to operate since R/2 is closed although XS is opened.

9. Cost
It is estimated that the complete equipment can be fitted to a standard power press for between £50 and £70.

10. Conclusion (Part 2)
It is intended to continue development of this electromagnetic detection device so that it can be used for interlocking purposes on the hand-feeding of large ferrous metal parts.

APPLICANTS FOR ELECTION AND TRANSFER

As a result of its March meeting the Membership Committee recommended to the Council the following elections and transfers.

In accordance with a resolution of Council and in the absence of any objections, the election or transfer of the candidates to the class indicated will be confirmed fourteen days after the date of circulation of this list. Any objections or communications concerning these elections should be addressed to the General Secretary for submission to the Council.

Transfer from Associate Member to Member

BASSETT, Arthur Geoffrey. *Guildford.*
JAMES, Michael. *Harpenden.*
LAKHANPAL, Dev Datta, B.A. *Bombay.*
WOOLDRIDGE, Geoffrey, B.Sc.(Hons.), M.Sc. *Weybridge.*

Direct Election to Associate Member

BARRASS, Robert. *Cambridge.*
KOLLANYI, Miklos. *St. Albans.*
MAXFIELD, Major Russell Hugh, R. Sigs. *Sheffield.*

Transfer from Associate to Associate Member

CRAWFORD, Robert. *Uddingston.*
DJOKIC, Cedomir, Dipl.Ing., *Wolverhampton.*
FISHMAN, David H. *Haifa.*
KHAN, Major Mohd. Rasheed, B.Sc., Pakistan Sigs. *Rawalpindi.*
REYNOLDS, Frederick Winston. *Shepperton.*
SEED, Thomas James, M.Sc., B.Sc. *Christchurch, New Zealand.*

Transfer from Graduate to Associate Member

BATES, Leslie John. *Bexhill-on-Sea.*
BELL, John Stanley. *Hull.*
BURROWS, Keith, B.Sc.(Hons.), M.Sc. *London, N.W.3.*
BURROWS, William Gordon. *Hatfield.*
CAMPBELL, Iain Robert. *Maidenhead.*
CZAJKOWSKI, Zbigniew, B.Sc.(Eng.). *London, W.9.*
DENBY, Peter. *London, N.W.4.*
FITCH, Stanley William Martin, B.Sc.(Eng.) *Stony Stratford.*
GRAHAME, Jackson. *Kitwe, Northern Rhodesia.*
HOPKIN, Peter Roy. *Hadleigh.*
HUTSON, Geoffrey Henry. *Birchington.*
McCONNELL, Geoffrey James. *London, S.E.18.*
MORRIS, Ian Glynn. *Liverpool.*
PRESTON, Neville Murray. *West Wickham.*
ROBERTS, Stanley John. *Belleville, Ontario.*
SIVASWAMI, Flt. Lt. Rishiyur S., B.A., B.E., I.A.F. *New Delhi.*
VENEEAR, Benjamin James, B.Sc.(Eng.). *Enfield.*

Transfer from Student to Associate Member

MATTHEWS, Capt. Norman Denis, R.E.M.E. *Shrivenham.*
SIMHI, Menashe. *Ramat-Gan, Israel.*
THURGOOD, Hugh. *Chelmsford.*

Direct Election to Companion

COOKE, Frederick George. *London, S.E.9.*
TOOTHILL, John Norman. *Longniddry.*

Direct Election to Associate

ALVES, George Joseph. *Westcliff-on-Sea.*
DALTON, John. *Bolton.**
EVANS, William Charles Allen. *Welwyn Garden City.*
HUSAIN, Ishtiaq, B.Sc. *Reyadh, Saudi Arabia.*
KENWRIGHT, William Arthur. *Warrington.*
MEREDITH, Donald Garmon. *Cardiff.*
PARRIS, Charles Deighton. *Port-of-Spain, Trinidad.*
PIZZEY, Donald. *Leicester.*
REDHEAD, Arthur George. *Montreal.*
SLACK, George William. *Dantbury, Essex.*
WILLIAMSON, Roger John. *Cambridge.*

Transfer from Student to Associate

MEYER, Leighton Francis. *Wellington, New Zealand.*

Direct Election to Graduate

BARLOW, Michael Owen. *Coventry.*
DEWHURST, Geoffrey, B.Sc. *London, W.5.*
GRIFFITH, William Bennett. *Birmingham.*
HILL, Joseph William, B.Sc. *Walton-on-Thames.*
HILTON, Alexander Frederick. *Eynsford.*
JONES, John Clement. *High Wycombe.*
OLORUNSHOLA, Samuel Ademola. *London, S.E.11.*
PRIME, Alec James. *Dartford.*
TALUKDER, Banajit Kumar, M.Sc. *Assam.*
WINSOR, John Michael. *Ilford.*

Transfer from Associate to Graduate

VALE, Gordon Thomas. *Bexley.*

Transfer from Student to Graduate

BALFRE, Paul Victor. *London, S.W.16.*
GALLIVER, Geoffrey Edward Lewis. *Isleworth.*
HELSZAJN, Josef. *London, N.16.*
IZZARD, Malcolm Ian. *Broxbourne.*
KONIECZNY, Gustaw. *London, W.3.*
LUND, Hugh Forsyth. *Natal.*
PREEDY, John. *London, N.17.*
SAPRU, Kanhaiya Lal. *Srinagar, Kashmir.*
SILLS, William Henry Albert. *London, E.11.*
SMIT, Cornelis. *Eindhoven.*
SPENCER, Godfrey Stanley Gibson. *Denmead, Hants.*
WASSENAAR, Dirk Jan. *Schaarsbergen, Holland.*

STUDENTSHIP REGISTRATIONS

ALDERSLADE, David John. *Birmingham.*

BAIRSTOW, Jeffrey Noel. *Bradford.*
BALMFORTH, George Lewis. *Cambridge.*
BOTT, Ronald. *Stafford.*
BOWEN, Joseph Alfred Edward. *Daventry.*
BRYSON-HAYNES, Donald. *Basingstoke.*
BUDD, John William G. *North Harrow.*
BUTLER, Edwin Leslie. *Basingstoke.*

CAPNISTOS, Emman. *Athens.*
CLARK, Tom William. *Montreal.*
COOPER, Thomas Sheamus. *London, W.5.*
CUNNINGHAM, Anthony John. *Leicester.*

DEAKIN, Barry. *Leeds.*

EIZAGUIRRE, Ramon. *Ahmedabad.*

FINUCANE, Mrs. Patricia. *Thorpe Bay.*
FRY, Anthony Jack. *Surbiton.*

HENDERSON, Laurence George. *Shetlands.*
HOWARTH, Edwin. *Torpoint.*

JONES, Basil Anthony David. *Maidenhead.*

KANTAK, Rajaram Mangesh. *Nova Goa.*
KAY, Michael. *Singapore.*
KHAN, Mohammed Akbar. *Lahore.*
KIRKMAN, Charles Neville. *Toronto.**
KOUNTOUEIOTIS, Georges. *Athens.*

LAM CHUN MING. *Southampton.*

MA SEK KIN. *Southampton.*
MASSINGHAM, Richard Peter. *Bury St. Edmunds.*
MUKHERJEE, Parimal Kumar. *Gaya.**

NICHOLAS, Russell John. *Southampton.*
NZE, Nieremibem A. *London, S.E.27.*

OKONGWU, Josiah O. *London, W.2.*
OLOYEDE, Joseph O. *London, S.W.1.*

PENSOM, Croombe Frank. *London, W.8.*
PERYER, Michael Gregory. *Croydon.*

RAMAKRISHNAN, Narayanaswami, B.Sc. *Tanjore.**

SMITH, James Bernard. *Wigan.*
SPEARMAN, Bryan Rosslyn. *Basildon.*
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TRINDADE, Anthony F. *Bangalore.*

VELATE, Anthony Spencer. *London, N.2.*
VENKATACHALAM, Keela Kandani Ramasamy. *Ramanathapuram.*

WALKER, Kenneth Arthur. *Rochester.*
WHITTAKER, George Bryan. *Cottingham.*
WILKINSON, Charles Leslie. *Prescot.*

* Reinstatement.

AN ANALOGUE COMPUTER FOR FOURIER TRANSFORMS*

by

Professor D. G. Tucker, D.Sc. (Member)†

A paper presented at the Convention on "Electronics in Automation" in Cambridge on 29th June 1957

SUMMARY

The paper gives the basic principles of a computer for obtaining a graphical display of the Fourier Transform of a function which can be represented by a finite number of ordinates over a finite range of the input variable. A delay line with linear phase-shift/frequency response is used to give the exponential term in the transform, and a time-variation of carrier frequency in sympathy with the time-base of the display unit gives the required variation of the index of this term.

1. Introduction

Fourier transforms have a wide application in engineering and science; two particular fields in which their application is important are

(*a*) the relationship between waveform and spectrum, and

(*b*) the relationship between the distribution of excitation along the length of an electro-magnetic or electro-acoustic array (or the distribution of sensitivity in the case of reception) and the directional pattern resulting from it.

In the latter field, in particular, several types of analogue computer have already been made.[1-4] A computer of more general application, although using individually-cut profiles for the input function, is that due to Born *et al.*[7] However, it is believed that the computer proposed in the present paper is novel, and has the merits of comparative simplicity and elegance. It has been developed as part of a new electronic beam-scanning process[5] for echo-ranging systems, the details of which will be published at a later date.

The relationships which the equipment can compute are

$$F(x)=A_1\int_{-\infty}^{+\infty} V(r)\,e^{j2\pi rx}\,dr \quad \text{.........(1)}$$

and $$F_m(x)=|F(x)| \quad \text{.........(2)}$$

provided the input function $V(r)$ has (as is usual) a finite value over only a finite range of the variable r, and can be represented as a series of equally-spaced ordinates. The complex transform is obtained in terms of its real and imaginary components, i.e. $F(x)=F_c(x)+jF_s(x)$, where

$$F_c(x)=A_1\int_{-\infty}^{+\infty} V(r)\cos(2\pi rx)\,.\,dr \quad \text{........(1a)}$$

is the "cosine transform" and

$$F_s(x)=A_1\int_{-\infty}^{+\infty} V(r)\sin(2\pi rx)\,.\,dr \quad \text{........(1b)}$$

is the "sine transform."

2. Principle of the Computer

The input is applied, as a signal of carrier frequency, at n input points—corresponding to n equally-spaced ordinates over a finite range in the r scale—which are connected to tappings on a band-pass delay line as shown in Fig. 1.

Let the inputs be $V(r)\,.\,A\cos\omega t$, where $V(r)$ is the function whose transform is required, and which is represented by n values V_1, V_2,... V_r,... V_n.

Let the delay line have a phase-shift φ per section at the frequency ω.

Then the output can be written in vector form as

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† Electrical Engineering Department, University of Birmingham.

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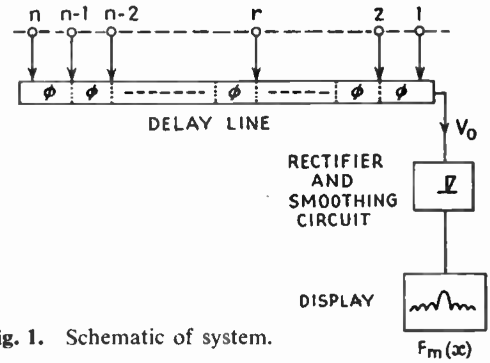

Fig. 1. Schematic of system.

$$V_0 = \mathbf{A} \sum_{r=1}^{n} V_r \, e^{j(r-1)\varphi} \quad \ldots\ldots\ldots(3)$$

where $\mathbf{A} = A e^{j\omega t}$.

Now suppose we vary φ with time. If the variation is linear, $\varphi = at$, say, then

$$V_0(t) = \mathbf{A} \sum_{r=1}^{n} V_r \, e^{j(r-1)at} \quad \ldots\ldots\ldots(4)$$

If n is large and the distribution may be regarded as continuous, then

$$V_0(t) = \mathbf{A}_1 \int_0^1 V(r) \, . \, e^{jra't} \, . \, dr \quad \ldots\ldots\ldots(5)$$

where r is now the continuous variable of $V(r)$ and equal to r/n in the discontinuous case, $a' = na$ and $\mathbf{A}_1 = n\mathbf{A}$. Thus if $V_0(t)$ is displayed on a time-base (by means of a cathode-ray-oscilloscope, for example) with the scale $x = (a'/2\pi)t$ then $V_0(t)$ becomes

$$V_0\left(\frac{2\pi}{a'}x\right) = \mathbf{A}_1 \int_0^1 V(r) \, . \, e^{j2\pi rx} \, . \, dr \quad \ldots..(6)$$

Now $\mathbf{A} = A e^{j\omega t}$ is only a convenient carrier with which to apply the amplitudes $V(r)$. Therefore in order to display the relationship of equation (2), we should rectify and smooth (with a suitable time-constant) before display. The requirements of this process make it clear that ω should be very large compared with the rate of variation of φ. Then, if $a' = 2\pi$,

$$F_m(x) = A_1 \left| \int_0^1 V(r) \, . \, e^{j2\pi rx} \, . \, dr \right| \quad \ldots\ldots\ldots(7)$$

which is one of our required Fourier Transforms.

In practice, we obtain a Fourier Transform of an input in series form, but if n is large, the effect is negligible. In fact, the transform cannot be obtained over a wide enough range of x to introduce a major error, since φ (i.e. the phase-shift per section of delay line) cannot exceed 2π radians—otherwise the pattern is merely repeated.

Before going on to show how the Fourier Transform of equation (1) is obtained, we must discuss how φ is made to vary with time.

3. How is φ made to vary with time?

The simplest way of obtaining the time-variation of φ (and this is the feature which particularly distinguishes the present computer from others), is to vary the frequency ω with time. If the delay-line has a linear variation of phase-shift with frequency, as shown in Fig. 2, then a linear variation of frequency with time will give a linear variation of φ with time.

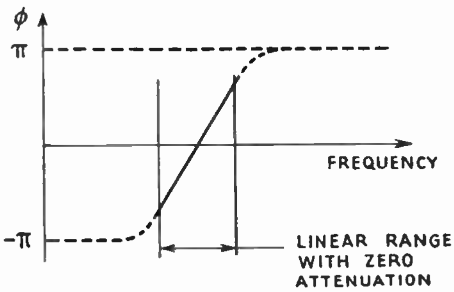

Fig. 2. Suitable phase-shift/frequency characteristic for delay line.

It is not, of course, necessary to have a linear variation of φ with time provided the time-base of the display corresponds with the time-variation of φ. Moreover, a manual variation of frequency will be satisfactory, if low-speed operation is acceptable, on the same condition.

The frequency variation can be most easily accomplished by sweeping the input carrier frequency over a suitable range. A frequency-modulated oscillator with saw-tooth modulation is most suitable. If for any reason the input frequency is fixed, e.g. by previous stages of a large computer of which the present equipment is only a part, then frequency-changers can be inserted in each input channel, the local oscillator being frequency-modulated by the saw-tooth waveform.

4. To obtain the Complex Transform and its Cosine and Sine Components

The method can be extended to give the transforms of equations (1), (1a) and (1b) by the use of so-called "coherent detection," or "phase-sensitive rectification" in place of the simple rectifier of Fig. 1.

To obtain the cosine transform, the rectifier is replaced by a modulator driven by a local supply of $\cos[\omega t + \frac{1}{2}(n-1)\varphi]$. Since φ is proportional to ω, this supply is readily obtained, in spite of the fact that ω is varying, simply by passing the input carrier $\cos\omega t$ through a phase-shifter with a linear phase-shift/frequency response. Observing that we are now concerned, not with vectors as when obtaining the modulus transform, but with instantaneous values of the signals, the process can be written down as follows:

The output of the delay line is

$$V_0 = A\sum_{r=1}^{n} V_r \cos[\omega t + (r-1)\varphi] \quad \text{.........(8)}$$

Therefore the output of the modulator is $V_0 \cos[\omega t + \frac{1}{2}(n-1)\varphi]$, which, after the removal of the high-frequency terms, i.e. terms in 2ω, is

$$\tfrac{1}{2}A\sum_{r=1}^{n} V_r \cos[\tfrac{1}{2}(n-1)-(r-1)]\varphi \quad \text{......(9)}$$

If now the inputs are re-numbered, with 0 at the centre, -1, -2, ... $-m$ on one side, and $+1$, $+2$, ... $+m$ on the other side, the output is

$$\tfrac{1}{2}A\sum_{r=-m}^{+m} V_r \cos r\varphi \quad \text{.........(10)}$$

On putting in the continuous variable, and making φ vary, as described in Section 2, we obtain the cosine transform

$$F_c(x) = \tfrac{1}{2}A_1\int_{-\frac{1}{2}}^{+\frac{1}{2}} V(r)\,.\cos(2\pi rx)\,.\,dr \quad \text{.........(11)}$$

To obtain the sine transform, the same process is used, except that the local supply to the modulator has to be $\sin[\omega t + \frac{1}{2}(n-1)\varphi]$. Since ω varies with time, it is not possible to obtain this directly either from $\cos\omega t$ or from $\cos[\omega t + \frac{1}{2}(n-1)\varphi]$. But, using well-known techniques[6], it is possible to obtain the required effect by inserting, in both the signal input to the modulator and in the local supply, phase-shift networks which have a *difference* in phase-shift of $\pi/2$ radians at all relevant frequencies. After removing the high-frequency (2ω) terms, the output becomes the sine transform

$$F_s(x) = \tfrac{1}{2}A_1\int_{-\frac{1}{2}}^{+\frac{1}{2}} V(r)\,.\sin(2\pi rx)\,.\,dr \quad \text{.........(12)}$$

The complex transform

$$F(x) = \tfrac{1}{2}A_1\int_{-\frac{1}{2}}^{+\frac{1}{2}} V(r)\,.\,e^{j2\pi rx}\,.\,dr \quad \text{.........(13)}$$

is evidently given by $F_c(x) + j\,F_s(x)$.

5. Example

Suppose $V(r)$ is constant over range 0 to 1, and zero elsewhere. Then from (7)

$$F_m(x) = A_1\left|V\int_0^1 e^{j2\pi rx}\,.\,dr\right|$$

$$= \frac{A_1V}{2\pi x}\left|\sin 2\pi x + j(1-\cos 2\pi x)\right|$$

$$= \frac{A_1V}{(\sqrt{2})\pi x}\sqrt{(1-\cos 2\pi x)}$$

$$= A_1V\left|\frac{\sin\pi x}{\pi x}\right| \quad \text{.........(14)}$$

as is well-known.

Also from (13)

$$F(x) = A_1V\int_{-\frac{1}{2}}^{+\frac{1}{2}} e^{j2\pi rx}\,.\,dr$$

$$= A_1V\,\frac{\sin\pi x}{\pi x} \quad \text{.........(15)}$$

which is the same as (14) but without the modulus signs.

6. References

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# of current interest . . .

Television Stations for the South-East of England

The B.B.C. has announced that low power television stations are to be installed at Swingate, near Dover, and in Folkestone. The difficulties that had arisen in finding wavelength channels on which these stations can operate without causing interference elsewhere have now been overcome, and both stations will be put into service as quickly as possible.

In order to provide a service in the Dover area without delay, a temporary station is being installed at the Swingate site, and comes into service on 21st April until replaced by a permanent station. The temporary station will use the same wavelength channel as the permanent station, i.e. Channel 2 (vision 51·75 Mc/s, sound 48·25 Mc/s) and will use vertical polarization.

The coverage of the Dover station when in its permanent form is expected to include Deal and Ramsgate. The effective radiated power will vary between 0·25 kW and 1 kW, with a maximum towards those towns. In its temporary form a simpler aerial will be used, so that the signal strength towards Deal and Ramsgate will be less, but in other directions will be roughly the same as for the permanent station.

The station in Folkestone will use a satellite transmitter of a new type developed by the B.B.C. The equipment is designed to be mounted out of doors. It incorporates a receiver to pick up the signals from another station and rebroadcasts them on a different channel in Band I. In this case the station will work in Channel 4 (vision 61·75 Mc/s, sound 58·25 Mc/s) with horizontal polarization; the effective radiated power will be 10 watts in the direction of the town of Folkestone. The station is expected to be ready for service within a few months.

Computer Exhibition and Symposium 1958

H.R.H. The Duke of Edinburgh has consented to be Patron of the Electronic Computer Exhibition and Business Symposium to be held at Olympia, London, from November 28th to December 4th, 1958.

The Exhibition and Symposium are being organized at the instigation of the National Research Development Corporation by the Electronic Engineering Association and the Office Appliance and Business Equipment Trades Association. More than 40 British manufacturers of electronic computers and ancillary equipment have taken space in the Exhibition.

The Symposium will stress the value of the computer as an aid to management and the papers will describe practical experience of the installation and operation of British computers and data processing systems; the majority of applications will be in general business management.

Immediately preceding the Exhibition and Business Symposium there is to be an associated scientific symposium organized by the National Physical Laboratory and to be held at Teddington, Middlesex, from November 24th to 27th. Admission is by invitation.

Broadcasting in Rhodesia and Nyasaland

On 1st February, 1958, a new broadcasting corporation, the Broadcasting Corporation of the Federation of Rhodesia and Nyasaland, came into being. The new Corporation has its headquarters in Salisbury, the Federal capital.

Broadcasting to this vast country, which has an area of over 485,000 square miles, is provided principally by means of vertical incidence short-wave transmission from two centres, namely Lusaka, the capital of Northern Rhodesia, which is responsible for broadcasting in nine languages to the 6½m. African population, and Salisbury, which provides the main English coverage to the ¼m. Europeans. In addition a primary medium-wave service is given to certain of the larger towns and centres of high population where this can economically be provided.

With the formation of the new Corporation further integration and technical development of the broadcasting system may be expected, and already medium-wave transmitters at Bulawayo in Southern Rhodesia and at Kitwe, the centre of the Copperbelt in Northern Rhodesia, are broadcasting the same English programme, linked by some 800 miles of Post Office trunk lines.

SOME ASPECTS OF THE APPLICATION OF CLOSED LOOP SERVO SYSTEMS TO MACHINE TOOL CONTROL*

by

R. J. F. Howard†

A paper presented at the Convention on Electronics in Automation" in Cambridge on 27th June 1957. In the Chair: Mr. E. E. Webster (Member).

SUMMARY

The paper discusses the combined design of machine tool and control equipment with particular reference to a two-axis profile follower system. The power requirements of the feeds are reviewed and in particular the effects of calling for excessive power on motor size and on motor torque/inertia ratio and hence overall servo system performance. The associated problems of "backlash" and "stiction" and the effects of these factors on the performance of the system as a whole are considered. The effects of the general rigidity of the machine tool itself, questions of vibration and of dragging of the stylus on the model of template are also examined. To illustrate the effects of the various mechanical factors on system performance, typical electronic copying systems are briefly described.

1. Introduction

The application of closed loop servo systems to machine tool control as, for example, in copying applications makes demands upon the design of the machine tool which differ considerably from the requirements encountered when the tool is intended to be manually controlled. The paper which follows outlines some of these problems, the way in which they are approached and the methods which are adopted for their solution, having regard for the fact that in many cases it is not possible completely to re-design the machine tool to suit the servo system and that compromise arrangements have therefore to be accepted.

This paper covers the application of one particular type of servo system used in copying applications, a considerable number of installations utilizing this system having been made during the past five years. The equipment is basically a two-axis profile follower and involves the problem of controlling the position of the tool in relation to the work by causing a stylus, forming part of a sensing head which is rigidly coupled to the tool, to follow the outline of a pattern or template.

2. Tool Positioning System

In the system to be described, an electrical means of positioning the tool making use of electric motors has been chosen primarily belause of its more universal application, particularly where long distances of tool travel are involved in either axis.

The tool feed drive motors are shunt wound d.c. motors and are fed from a split-field generator system, the generators being driven by an a.c. motor. The basic arrangement is shown in Fig. 1. The use of a split-field generator enables the output of the generator to be smoothly controlled throughout its full range of voltage output from the maximum positive voltage through zero to the maximum negative voltage. In this way, the speed of the servo motor can be controlled from full speed in one direction through standstill to full speed in the other direction. A further attribute of the split-field Ward-Leonard system which has been chosen for use in this particular application is the fact that high peak currents can be provided for short periods from the generator without danger of thermal over-loading. In this way, high peak torques

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can be provided at the servo motor shaft and to accelerate the motor. In providing the energy for these high peak torques, use is made of the fact that stored energy is available in the motor generator set. In addition, when the output voltage is reduced to slow down the servo motor, the system inherently regenerates to the new speed, thus ensuring that the speed change at the servo motor shaft is achieved as rapidly as possible.

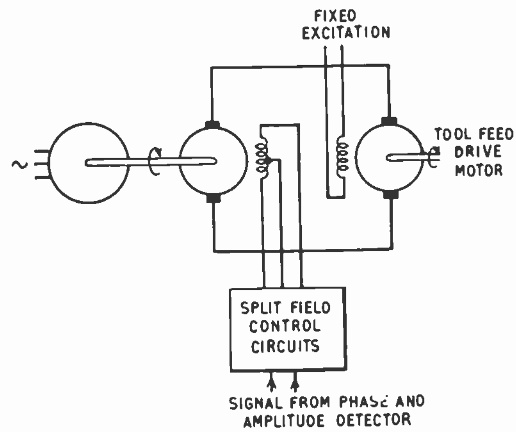

Fig. 1. Basic circuit arrangement of split field Ward-Leonard system to provide continuous reversible control of the positioning servo motor.

It is, of course, important to ensure that the servo motors rapidly follow any call for changes in their speed and the problem of the speed of response of the power section of the servo is particularly important. It is, therefore, vital that the highest possible speed of response is provided in the generator, so that when the field excitation is changed, the output voltage also changes very rapidly. This is achieved by designing the field system of the generator to provide the highest possible factor of merit in this machine, i.e., that it has high gain and fast response, and by designing the circuits which feed the field of the generator so that they force it as much as is permissible.

Whilst it is important to minimize the response time of the generator, the main limitation in the overall servo system lies in the mechanical response time of the servo motor itself. It is therefore vital to provide the highest possible torque/inertia ratio so that a high accelerating torque is being used to minimize the effects of inertia. This, in many cases, means that a special design of motor would enable improvements in performance to be achieved. As in the case of the design of the machine tool, however, production demands are generally too small to warrant the high cost of tooling for a special motor. In consequence, motors already available have to be carefully chosen in order to provide the desired performance. In choosing the motor it is also necessary carefully to consider the rating at which the motor is used, so that it can be short time rated, thus enabling a physically smaller motor to be used for a given rating, with consequent improvement in the torque/inertia ratio. Other expedients, such as forced ventilation and special high temperature rated insulation materials, can be considered, subject to other limitations which are discussed later. It is important to note at this point the extent to which machine tool designers can assist by minimizing the power requirements of the tool feed arrangements and certainly by allowing the servo system designer to provide a motor which is correctly powered and not over-powered. A high price is likely to be paid in relation to performance if considerable safety factors are included when assessing the servo motor power. Apart from other advantages, which will be discussed later, the question of power requirements also emphasizes the need for designing tool feed mechanisms for minimum mechanical losses so that the power requirements are held at the lowest possible level.

When it is considered that many forms of single start lead screws and worms have an efficiency in the region of 20 per cent., the factor by which, in some machine tools, the feed motor power can be reduced will be clear. Certainly, in many cases, motors having a rating many times that necessary are having to be fitted, with consequent increases in the overall control equipment cost.

It will also be appreciated that, as has been indicated earlier, because the output voltage of the generator is controllable through zero, the torque from the servo motor must be as smooth as possible through the zero point. For this reason it is necessary to consider the design of the armature on the motor and arrange for the armature slots to be specially skewed and semi-enclosed. In this way a smooth torque can be provided through the zero speed point without any tendency for "cogging" normally associated with motors having straight slots. The point is emphasized by the requirements which will be discussed in connection with a particular lathe application where the design of the tool feed arrangements are such that 0·014 in. of tool motion is provided by each revolution of the servo motor, which means that 0·001 in. is equivalent to 25 degrees movement on the servo motor shaft. This represents a limit dictated by the operating speeds required, but does illustrate clearly the problem which can arise in such cases.

In considering the short time rating of the motors in order to minimize the size and thus improve the torque/inertia ratio, a further factor must be borne in mind which will be discussed in more detail later. This is the question of "stiction," which is a convenient way of describing static friction or coulomb friction, which results in a high torque being necessary in order to start the motion of the tool carrier from standstill. If the short time rating of the motor is carried too far and the motor made too small it may be found that the saturating torque of the machine is insufficient to break away against the "stiction."

In referring above to the importance of minimizing the power requirements for the tool by the use of anti-friction slides, screws and bearings and similar arrangements, it is appreciated that it is frequently not possible to introduce such expedients without a complete re-design of the machine tool and, again, as in the case of the problem of the associated electrical machine, it becomes necessary for a compromise design to be accepted.

3. The Basic Servo System

Having discussed the method of power control which has been adopted to provide the motion of the tool carrier in two planes, reference can now be made to the particular system used to ensure that the sensing head unit carrying the stylus, rigidly coupled to the tool holder, is forced to follow the surface of the template or model which is being copied. The basic problem here is to arrange a system which will determine the position of the stylus in the sensing head in relation to the template or model and, having determined its position, to make this stylus move along the template or model in intimate contact with its surface and to do this irrespective of the shape of the template.

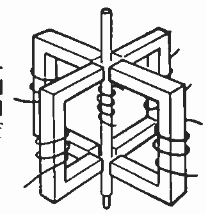

Fig. 2. Diagrammatic representation of sensing head showing positioning of fixed cores and coils and of pickoff coil on stylus rod.

In order to achieve indication of the relationship between the stylus and the template, it is necessary to be able to determine how much the stylus is deflected and in what direction. The design of the sensing head to achieve this can be seen in Fig. 2.* The four coils are energized in phase rotation, i.e., 0°, 90°, 180°, 270°, at a high frequency chosen to provide the requisite information rate, to achieve adequate speed of response and also to avoid interference as a result of unwanted signals from the 50-c/s supply, introduced from the circuits associated with the sensing head. A sensing coil is mounted on the stylus rod itself which, as will be seen from Fig. 2, passes through the centre of the four coils, of which one form is shown in Fig. 3. In the practical configuration of this head unit, the four coils are rigidly mounted in a block of cold-setting resin material, the aperture for the stylus rod being subsequently machined out, so that cylindrical faces are provided on the pole pieces formed by the U-shaped cores. This can be seen from Fig. 4, as can the arrangement of the winding on the stylus rod, which is also provided with two cylindrical pole pieces corresponding to the pole pieces of the fixed coils.

The signal from the coil on the stylus rod will be zero when the stylus rod is in the

* The sensing head, and overall system are covered by British Patent No. 770,645.

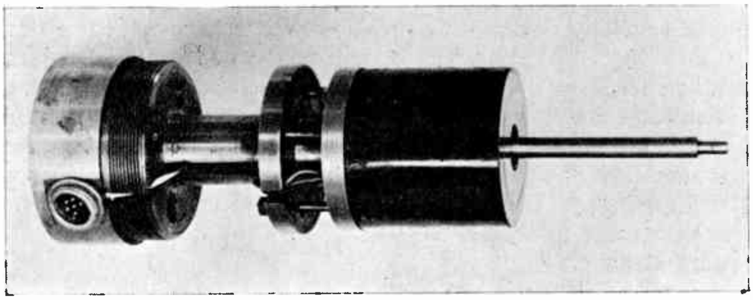

Fig. 3. Interior of a typical sensing head unit. The stylus can be clearly seen, as can the moulding containing the fixed coils. In this version a socket is provided for connection of the incoming leads and the whole head is enclosed in a steel tube which screws on to the thread seen on the lower part of the unit.

electrical zero point of the four fixed coils. Any deflection of the stylus will cause a signal to appear from the sensing coil, the amplitude of this signal depending upon the amount of deflection. Similarly, the phase of the signal will depend upon the direction of deflection.

In the particular form of sensing head arrangement illustrated in Fig. 3 a simple cantilever rod in a special alloy material is used, this material having been chosen after a considerable amount of development work and extensive testing, so as to provide the minimum possible mechanical hysteresis, i.e., if the stylus rod is cyclically deflected in one direction a large number of times it will, after the deflection has been removed, still return closely to the zero position. During development work, many forms of mounting arrangements for the stylus arrangements were tried, including systems incorporating restoring springs, diaphragms, ball pivots and so on. Many of these not only exhibited hysteresis but also ill-defined recentreing which rendered them completely useless. The simple arrangement which can be seen from Figs. 3 and 4 was ultimately found to be the most satisfactory, providing that the material is carefully chosen and the mechanical configuration carefully designed.

Arrangements are provided in the head to enable the stylus rod to be set to the electrical zero point after manufacture. Due to normal manufacturing tolerances the electrical zero does not always exactly correspond with the mechanical centre of the system and provision is, in any case, needed for subsequent recalibration.

Hysteresis in the present design is negligible and is almost certainly masked by effects due to temperature changes. A sensing head of this type has been subjected to one million operations of 0·100 in. deflection and after this very severe test showed a final hysteresis of only a small fraction of 0·001 in. In view of the fact that this particular type of head, as will be seen, is never normally deflected more than 0·025 in., a considerably safety factor in performance has been allowed.

Having discussed the mechanical arrangements of the stylus, the arrangement of the servo system and the method of operation can

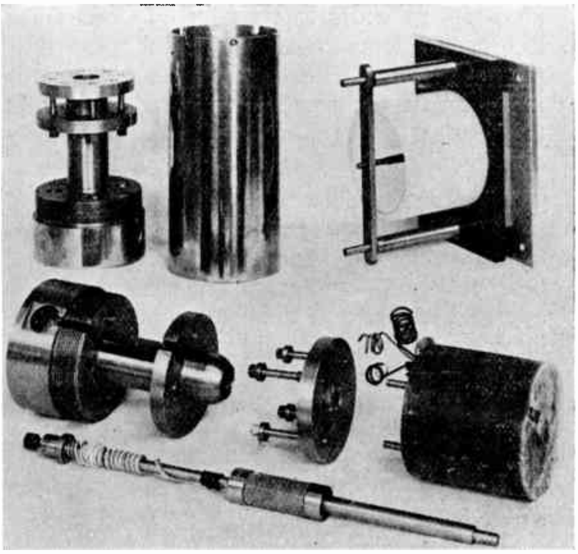

Fig. 4. The component parts of the head unit illustrated in Fig. 3, together with the tool used to produce the moulding. The coil assembly illustrated has not been machined. The spherical bearing arrangement to enable coil centreing can be clearly seen in the lower part of the illustration, as can the design of the stylus rod. The pole pieces on the stylus are clearly shown as is the moulded enclosure for the sensing coil.

be briefly described. Initially, the sensing head is in free air, the stylus not being at that time in contact with the template. Signals are introduced into the servo system which cause the sensing head to be moved towards the template in a predetermined direction. As soon as initial contact is made, the stylus is deflected in a direction normal to the plane of the surface of the template at the point of contact. This initial deflection is immediately arranged to cause the overall tool positioning servo system to move the stylus further towards the template in the same direction. As soon as the stylus has been deflected by some predetermined amount between 0·010 in. and 0·050 in. (a typical deflection being 0·025 in.) the system is so arranged that the phase of the signals fed to the head unit is rotated so that the tool carrier drive systems move the tool carrier, and hence the stylus, in a direction tangential to the surface of the template at the point of contact. The graph in Fig. 5 shows the effect on the speed of the drive motors of the deflection of the sensing head.

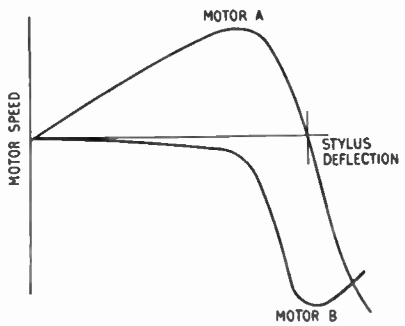

Fig. 5. Illustrating the relationship between the speed of each of the two servo motors (one in each axis) as the deflection of the stylus is increased after first making contact with the template. The operation of the system to feed the tool and sensing head closer to the template and thus further deflect the stylus can be clearly seen, as can the effect of the control system on motor speed as the bias point is reached, thus causing motion tangential to the surface to commence.

It is not within the scope of this paper to discuss the system used in any detail, but the basic arrangement can be seen from the block diagram in Fig. 6. It will be seen that the output of the coil on the stylus is fed to phase and amplitude sensitive rectifier arrangements which in turn provide d.c. signals of magnitude and polarity which call for the required direction of rotation and speed of the servo motor associated with that particular plane of motion. In addition, the output of the coil on the stylus is also fed to an amplitude detector which, when the amplitude reaches a predetermined amount equivalent to the output at the required "bias" deflection, provides a signal to a phase shifting arrangement which then causes a phase rotation of the signals to the four coils on the sensing head unit proportional to the further deflection.

By this means, the signal from the coil on the stylus fed to the phase sensitive rectifiers which initially calls for motion from the servo motors moving the stylus further on to the template, is modified when a selected deflection is reached so as to cause the control circuits to call for motion from the servo motors such that the stylus tip moves tangentially to the surface. In fact, the circuit arrangements are such that the effective rotation of the phase of the signals in the coils of the sensing head unit is achieved by shifting the phase of the reference signal to the phase sensitive rectifiers.

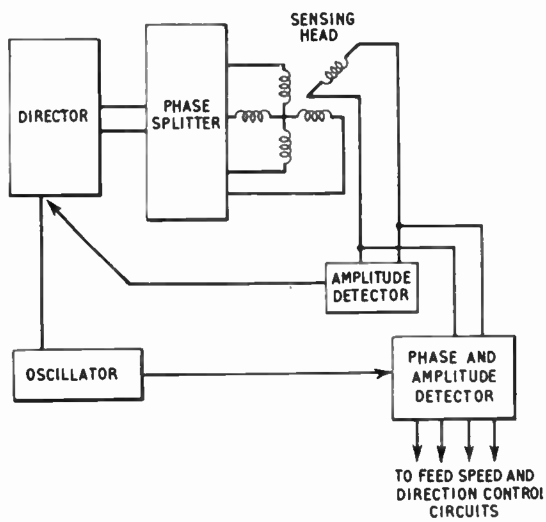

Fig. 6. Diagrammatic illustration of the main elements of the system. The output from the phase and amplitude detectors is fed to the power control circuits which are generally similar to those illustrated in block form in Fig. 1.

For the purposes of description, however, it is easier to consider the phase rotation of the signals within the head unit itself. It will be seen from the graph in Fig. 5 that the full rotation of 90 deg. is achieved very rapidly at deflections beyond the so-called "bias" deflection and that any further deflection beyond this point will, in fact, cause a shift of more than 90 deg. Thus, if the stylus encounters a surface which tends to increase its deflection it will provide the requisite signals to the servo motors to cause the necessary motion to follow that surface. The system is such that the stylus can be caused to follow a circular template and if an annular ring is considered it will, in fact, move in one direction round the outside of the ring and if it is transferred to the inside of this ring it will move in the opposite direction.

Before leaving the description of the stylus head and the servo arrangement, attention should be drawn to the fact that various protective arrangements are incorporated, the most important final protection being an arrangement normally referred to as an "overdrive ring." with which the stylus makes electrical contact, which is arranged to shut down the whole system through a relay arrangement should the stylus for any reason be deflected by an amount somewhat in excess of the normal working "bias" deflection. This protection is incorporated in the sensing head in order to avoid tool damage or stylus head damage in the event of serious over-deflection.

4. Application Problems

Having described the factors associated with the protection of the tool drive arrangements and the general arrangement of the control system which causes the stylus to follow the template, the application problems of these systems to machine tools can now be discussed.

The problems of application can be divided into two broad groups. Firstly, those which might be described as covering the system problems, i.e., which are associated primarily with the design of the servo system itself; and secondly, the application problems, which involve the overall performance of which the machine tool is capable when being controlled by a servo system of the type under discussion.

Dealing first with the system problems, these are primarily associated with the speed of response of the servo system. It will have been seen that speed of response for which the system is designed can be affected by the machine tool characteristics, but nevertheless the problem is primarily one associated with the control system once the parameters of that system have been set by the design of the machine tool itself.

The limitations associated with the system problems are primarily those of "overshoot" of the system when a sudden change is called for. One example is undercutting on inside right-angled corners and overshoot on outside right-angled corners. If, for example, the tool and the stylus are moving along a parallel surface and the template suddenly calls for movement at right angles to this surface, then in theory one servo motor should immediately stop and the other accelerate to full speed in zero time. This is, of course, physically impossible and there is bound to be undercut. The system parameters determine the amount of the undercut. Similar remarks apply to overshoot when the reverse conditions are called for.

There are other characteristics of the system which are important, in particular the question of smooth response through zero speed and the provision of smooth torques at low speeds. The factors involved here have been discussed at some length earlier in this paper. These factors are, however, largely secondary to the problem of speed of response, which is, of course aggravated in the case of systems where very high tool feed rates are called for. There is a further factor to be considered where sudden changes of direction taking the form of right angles are involved, in that the tool itself is generally incapable of producing a right-angled corner and this fact may mask the effects of undercut to some measure and overshoot to a greater extent. The problem arises primarily on lathes and boring mills and is of less importance on other types of machine tool. On lathes, particularly of high production types, where a high tool speed is involved, the difficulties of undercut and overshoot are sometimes overcome by the introduction of fixed tools on a backslide to perform the final operation after the profiling operation has been completed. These fixed tools on a backslide can be selected so as to overcome the difficulties of undercutting and overshoot at points where critical dimensions are involved.

Turning now to problems of application, these are factors directly associated with the machine tool design. The main factors involved are "stiction" which has already been referred to and which necessitates the provision of torques considerably higher than the running torque in order to cause the tool to commence moving; "backlash," in which a discontinuity of tool motion occurs despite rotation of the servo motor; "wind up," in which rotation of the servo motor causes a torsional displacement down screws and shafts without producing tool motion; vibration, which is self-explanatory and can introduce spurious signals into the sensing head unit, and finally a factor which, for want of a better word, can be described as general "slop" in the machine tool. Examples are "crabbing" action of machine tool carriers on slides, insufficient rigidity in template or model holders and so on.

A factor which does not really fall within the scope of a paper dealing with issues associated with the system design, is the question of differential wear, which must, however, be considered if the performance after an appreciable running period is to be taken into account. If the effects of wear adversely affect any of the factors outlined above, the performance will clearly deteriorate from that obtained at the outset.

Rather than endeavouring to examine these factors in an abstract manner, their effects can be perhaps better outlined by describing the results of such effects on typical machine tool applications.

5. Typical Applications

In a lathe which is designed to take work from 1 in. to 10 in. in diameter and which is capable of very heavy cuts at high speeds, the speed "round the work" of the tool is arranged to be adjustable between 1 in. and 20 in. per minute. This descriptive statement of the speed "round the work" is a convenient method of expressing the vector sum of the speed resulting from the operation of the servo motors in the two planes. Normally a control is provided on the servo control systems of the type under discussion which enables this vector speed to be set to the appropriate level for the type of work being machined.

The lathe is shown in Fig. 7 and a close-up of the sensing head in Fig. 8.

On the type of lathe under discussion, the back-lash on the cross-slide can be expected to be of the order of 0·002 to 0·004 in. and on

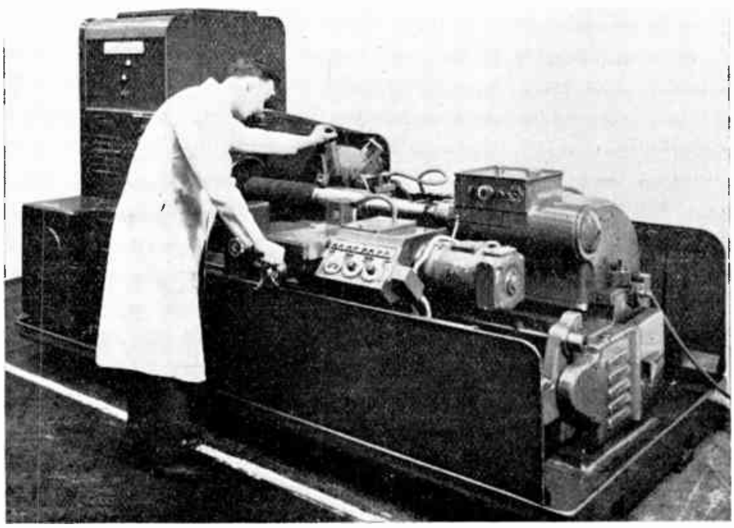

Fig. 7. The Drummond "Maxi-cut" lathe with electronic control cubicle in background. The cross-slide servo motor can be clearly seen alongside the control panel. The two manual controls are for "speed round the work" and the rate of entry of the backslide tool. Push-buttons control the various functions of the lathe which are electrically operated. The sensing head is in the housing which appears behind the operator's right arm.

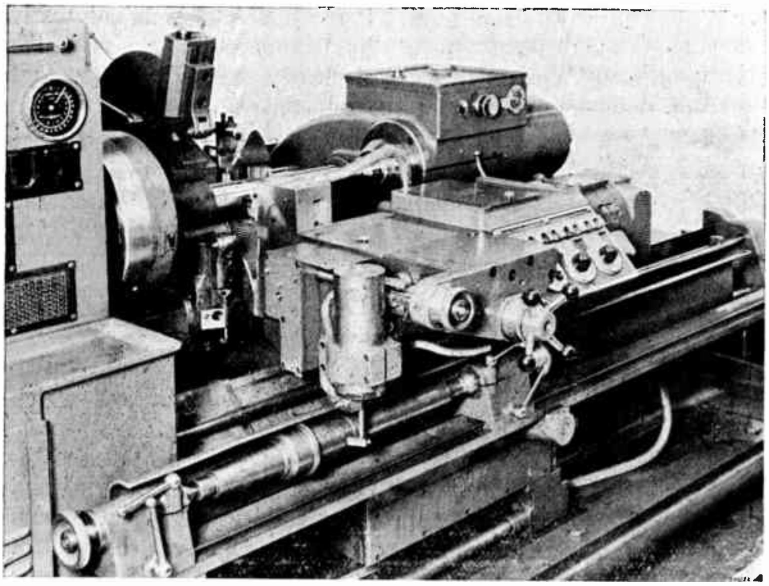

Fig. 8. Showing the sensing head with the cover removed. The work is clearly shown, also the method of mounting the tool tip on the end of the stylus rod.

the saddle some 0·006 in. The stiction, measured in terms of the starting to running torque requirements at the worst point on the cross-slide or saddle, is of the order of 1·5:1 to 2:1. The wind-up on the machine is negligible and, because of the type of work for which it is designed, the machine is of a very rigid type. On a machine tool of this nature, at speeds round the work of up to 6 in./min., copying accuracies of about 0·001 in. on diameter are achieved. Considering an operating speed of 6 in./min. this may obviously be expressed as 6000 thousandths of an inch per minute and under these conditions the undercut is in the region of 2 thousandths of an inch. The ratings of the motors on these machines are 0·5 h.p. for the cross slide and 1 h.p. for the saddle, this being the nominal 100 per cent. duty cycle rating of the machines in question.

As the speed round the work is increased, the problem becomes increasingly difficult, particularly as far as undercut and overshoot are concerned and in addition the problem of maintaining accuracy is also considerably more difficult. At speeds approaching the maximum of 20 in./min. copying accuracies of the order of 0·002 in. are obtained, although vibration, particularly where heavy cuts are involved, becomes a considerable problem. At this higher speed the undercut at right-angled corners is of the order of 0·010 in. and the overshoot, which is not noticeable at the lower speed of 6 in./min., is about 0·003–0·004 in. This machine is fitted with a back slide in order to overcome overshoot and undercut problems at points where accuracy of the work is vital.

Some interesting experiments have been carried out on this particular design of lathe by making alterations to the gear ratios between the servo motors and the tool carrier, thus necessitating a sacrifice in the maximum operating speed. By doubling the number of revolutions per thousandth of an inch of tool travel and taking particular precautions to reduce stiction and also strengthening the template holder and so on, tests have been made possible to assess the ultimate accuracy of this particular machine. With the alterations outlined and tracking the motion of the stylus in relation to the template by means of a clock gauge arrangement, it was found that the actual tracking errors at 4 in./min. were of the order of $\pm 0{\cdot}0001$ in. The machine was used to bore a 9 in. cylinder where the required accuracy was $+0{\cdot}001$ in.–0·000 in. It was found in actual practice that the limitation of this machining operation did not lie in the machine tool but rather in the dimensional changes in the work due to temperature change between the machining operation and removal of the work from the machine.

In undertaking the tests described above, the working bias of the stylus was reduced from the normal figure of 0·025 to 0·010 in., this type of modification only being possible where it can be guaranteed that the backlash in the machine will never approach or certainly exceed the normal working bias. Should this occur, the system will, for obvious reasons, completely lose control because of the discontinuity introduced by the backlash. With the modified sensing head for 0·010 in. bias deflection the stylus tip pressure was of the order of $1\frac{1}{2}$ oz and because of the high overall accuracy capability of the rest of the system limitations were encountered in relation to the drag of the stylus itself on the template. A considerable amount of development work is at present being carried out in relation to stylus and template materials in order to minimize the effects of stylus drag. Even on machines of normal commercial form badly finished templates or stylus tips can introduce appreciable errors due to drag. It is particularly important where ferrous template materials are used to ensure that both the template carrier and the template itself are completely demagnetized, since any magnetization will obviously increase drag problems.

It is appropriate here to refer to the design of the stylus tip, which is usually made of comparable shape to the tool tip itself. Because of the working bias of the system, normally 0·025 in., it is usual to make the template or model oversize by the amount of the bias. It is, in certain cases, possible to arrange the relationship of the tool size to the stylus tip size to correct for the bias deflection, but this is obviously not possible in cases where work on the inside and outside of a particular template is involved, since the correction can only apply in one direction.

In another application, a boring mill, the problem is entirely different to that of the lathe referred to above. In this machine, a much lower speed round the work is involved, 3 in. being the maximum with a 10:1 range, i.e., 0·3 to 3in. In addition, even lower speeds can be selected by suitable modification of the gear ratios between the servo motor and the tool carrier. On this particular design of machine, accuracies of the order of 0·0005 in. are achieved and the performance limitation lies entirely in the machine itself, due to its larger size, the longer drives involved between the servo motors and the tool carrier, which

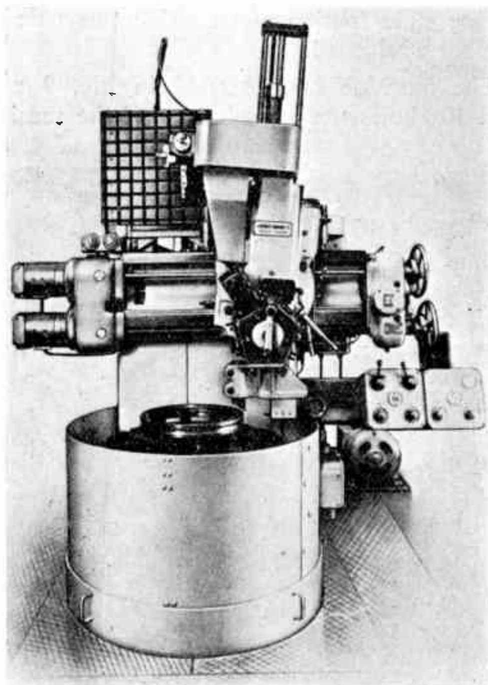

Fig. 9. Webster & Bennett boring mill showing two servo motors to the left and the sensing head mounted on the upper part of the tool carriage.

tend to introduce wind up, and by increased stiction and similar effects due to the large slide size and the mass of the saddle. A considerable amount of work has been undertaken on this particular type of machine tool to minimize all the various factors within the limitations imposed by economic design considerations and the performance referred to above has been achieved on what is, in fact, a standard design of boring mill with only minor modifications.

On this particular machine the performance referred to above has been obtained with backlash figures of 0·008 in. in each axis. Because of the machine size, the stiction ratio varies between wide limits, but can be expected to be of the order of 2:1. The rating of each driving motor is 0·5 h.p. Overshoot and undercut problems are not of major importance because of the relatively low speed round the work. In an endeavour to overcome the stiction problem a considerable amount of work has been undertaken with molybdenum disulphide lubricants as well as by other means of minimizing slide friction and considerable success has been achieved.

The machine is illustrated in Fig. 9 and Fig. 10 shows the arrangement of the sensing head, in this case mounted in the horizontal plane. The template carrier can be retracted hydraulically. Several templates can be seen in position and these may be used with different tools on various parts of the work piece as illustrated by Fig. 11.

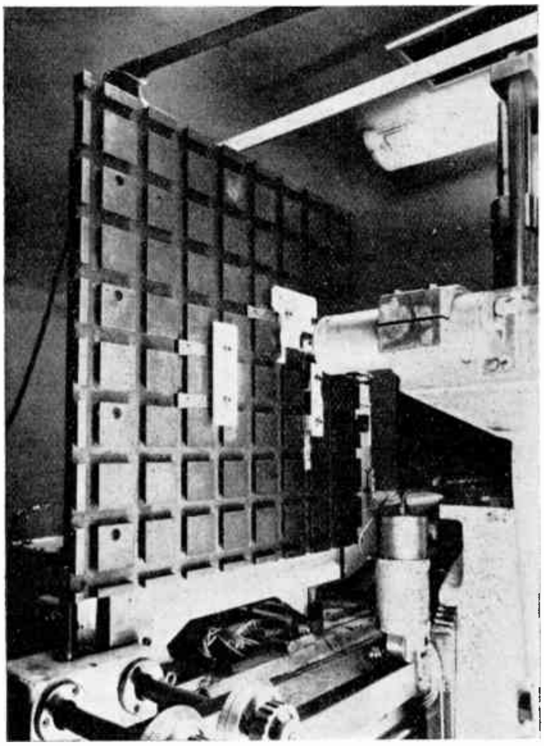

Fig. 10. The method of mounting the sensing head and the template carrier. The latter is hydraulically positioned. Three templates can be seen mounted and these can be brought into use to enable different parts of the work to be machined.

Turning now to a further application, in this case to a horizontal boring mill, which is very much larger in size than the vertical boring mill referred to above, the problem becomes largely one of the magnitude of the backlash which is involved. In the rack drive in the horizontal direction a backlash of some 0·025 in. is typical with 0·010 in. on the vertical screw drive. Because of these backlash figures, a bias deflection of some 0·050 in. has to be used and with the backlash figures and a stiction ratio approaching 3:1 an overall accuracy of about 0·003 in. is obtained at 1 in./min. The speed round the work on this particular machine is 0·5 to 6 in./min.

In a system of this nature there are quite appreciable stabilization difficulties from the servo system point of view, but these have largely been overcome. Some idea of the size of the machine can be appreciated from the fact that the machine will shape plates of metal which are some 25 feet long. In this particular machine the template is mounted alongside the work and the stylus mounted horizontally. This introduces a further problem in that it is necessary to fit a disc on the end of the stylus which is equivalent in diameter to the milling cutter attached to the tool carrier. This means that the stylus tends to drop due to the weight of the model on the milling cutter and the sensing head has to be set to electrical zero for this condition. Naturally, if the weight on the end of the stylus is altered, the sensing head has to be reset to zero condition, but in general an endeavour is made to ensure that all models of the milling cutter mounted on to the stylus within a given range are loaded to the same weight to avoid this necessity. With this system, the question of drag on the template has to be watched very carefully, although the generally lower accuracy of operation due to the machine characteristics tends to obscure difficulties resulting from stylus drag. On the machine referred to above the drive motor for each axis of tool motion is rated at 3 h.p.

An important aspect of the control of any type of machine tool where milling cutters are involved is the question of vibration. It is perhaps a more severe problem on this type of machine because of the "chatter" transmitted from the milling cutter through the machine to the sensing head. This "chatter" is frequently found to occur in the frequency range between 4 c/s and 26 c/s. It can, of course, be minimized by appropriate cutter design and by careful consideration of the cutter design in relation to feed rates. It is possible for the effects of vibration to be minimized in the control system itself, but usually this cannot be achieved without affecting other system performance factors and in particular the speed of response. Unless care is taken to minimize vibration in the machine itself, the ultimate performance of the system may be materially affected.

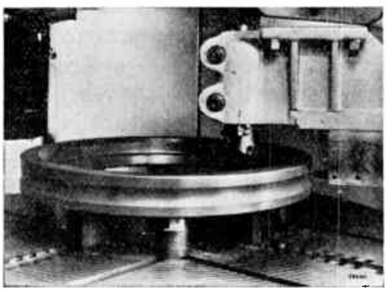

Fig. 11. Illustrating the method by which the tools are used, one for an inside and one for an outside form corresponding to the number of templates mounted on the template carrier.

A further point of interest in relation to these particular machines is the use of a number of pushbuttons radially disposed around a circle on the control panel, these pushbuttons enabling a direction of travel to be selected, prior to commencing the profiling operation, dependent upon the particular button depressed. In this way, the stylus can be readily moved on to the template at any point and from any direction.

6. Conclusion

Summarizing, the problems from the machine tool viewpoint in terms of the system requirements are as follows. Accurately defined power rating of the tool feed motors so as to minimize the overall response time of the servo system and so minimize overshoot and undercut; great care in the design of the tool to limit backlash, stiction, wind up and vibration. Close co-operation between the control system designer and the machine tool designer is essential and, in particular, it is important, when new designs of machine tool are being considered which may ultimately, if not at the outset, be used in conjunction with servo systems, that discussion should take place to ensure that the requirements of the servo system in terms of machine performance are fully appreciated. In control systems of the type which have been described above an important point to be borne in mind is the fact that the control system and the machine can no longer be regarded as separate entities since both form part of the loop of a closed loop servo system.

Although this paper has been concerned with a particular form of two dimensional copying control which has been in operation on a considerable number of various types of machine for several years, many of the points which have been made are, nevertheless, equally applicable to other types of control system and in particular to those in which stored information is used to position the tool instead of a model or template. Work currently in hand in this direction indicates even more need for close co-operation between the machine tool designer and the control system designer where these types of arrangement are involved.

7. Acknowledgments

Acknowledgments are due to the design, development and engineering staff of Lancashire Dynamo Electronic Products Ltd. and in particular to Mr. K. A. Robinson, Chief Development Engineer, Mr. R. H. Whitlock, Senior Development Engineer on machine tool control projects, and to Mr. B. R. Townsend, Chief Designer.

Full acknowledgment should also be given to Messrs. Webster & Bennett Ltd., Drummond Brothers Ltd. and the Asquith Machine Tool Corporation Ltd. who are amongst the companies which have been closely associated with the work described.

DISCUSSION ON

"Machine Tool Control"

Contributions made during Session Two of the Convention on "Electronics in Automation" at Cambridge on 27th June, 1957.
In the Chair : Mr. E. E. Webster (Member).

Dr. D. A. Bell*: Referring to the paper by Messrs. Finden and Horlock[1], is not one of the limitations of the numerical control of jig-borer's the necessity to relate the origin of measurement of the machine to the work-piece, before the machine can select any desired drilling centre ?

P. D. Saw†: Can Mr. Coppin[2] say whether, in view of the comparatively large number of rotations made by the high resolution potentiometer, i.e. approximately 1,000 for a complete traverse of the table, this component can be expected to have a reasonably long life without deteriorating from the original high accuracy claimed?

D. G. Chapman‡: With reference to the machining of actuator gear boxes, as described in the paper on the Inductosyn,[1] do the authors consider the use of the equipment they describe would enable such products to be made without any inspection whatever ?

N. A. F. Williams§: I would be interested to know from Mr. Alexander[3] whether his units are intended for use as generalized analogue computing elements, or for a specific application. They seem similar to Blackburn analyser units and I would like his comments on this point. How does he, for example, perform integration ?

Referring to the paper by Mr. Howard,[4] I should like to know if jittering of stylus could be used to reduce stylus/template drag. Also, has any difficulty been experienced with residual magnetism in the generator of the Ward-Leonard set ?

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† P.A.M. Limited.

‡ Sylvania-Thorn Colour Television Laboratories

§ Brush Electrical Engineering Co. Ltd.

D. F. Nettell*: Referring to the relative merits of punched tape and punched cards as a data input medium, there are many cases where cards are superior to tape. In many processes the numbers of instructions are limited and can be contained in one card. The advantages of one card—one product are obvious. Punched cards can also be used with advantage where the instructions occupy several cards. The main technical objection is that in complicated processes such as profile milling, there is a fear that the cards may get out of order and there is probably a good argument for tape in this case, although serial numbering of the cards would be possible.

An objection often raised against cards is that the equipment is more expensive. This is true, but it must be realized that commercial tape equipment reads one character at a time, and for a ten-digit instruction this would have to be put into a buffer store. A block sensing card reader is, however, a store in itself, and the disparity in cost may, therefore, not be so great as it appears at first sight.

A feature of the use of punched cards is that the same cards can be used for machine control and for accounting purposes. Hence the integration of production processes with control and accounting procedures would be enhanced by their use.

T. A. Waite†: I should like to refer to the papers of Mr. Alexander and Mr. Howard. Concerning Mr. Alexander's paper,[3] I am not clear why the transformers he describes can be made accurate in turns ratio to one part in 10^5. Mr. Alexander stated that this was due to the high permeability cores used and the symmetrical deposition of the windings, but both these conditions can be obtained with conventional

* Powers-Samas Accounting Machines Limited.

† Vickers Group Research Establishment.

windings and cores. Perhaps Mr. Alexander could tell us the fundamental reason for the increased accuracy of his transformers in checked, and the type of apparatus used.

Referring to Mr. Howard's paper,[4] I should like to know why he chose an electric system of power drives for his machines in preference to hydraulic or pneumatic systems.

H. J. Elton*: It may be of general interest to know that the Department with which I am associated is developing an inspection machine that will be automatically programme controlled, using the Ferranti tape system with diffraction gratings. The problem has been to design a measuring head which is universal in its action so that it can follow the programme path normally taken by a cutter. This problem has been overcome, and a recent demonstration was held at Messrs. Ferranti, Edinburgh, the results of which show that this has great promise.[6] Development is proceeding and it seems that it would have many applications in industry.

D. R. Holloway†: My remarks are directed to the fact that all the papers[1—5] deal with the positioning, by template and follower, or other means, of the tool or cutter of the machine tool being controlled. This method of approach does not, of course, take account of the possibility of tool wear, or deflection, which may occur to a varying degree. Have the authors considered this point, and can it be overcome by introducing a factor into the control mechanism? Of course, my fears may be groundless, and the effect of this phenomenon may be negligible.

As a complete stranger to the field of automatic control of machine tools, and one who is therefore obliged to consider the problem on theoretical grounds alone, it would seem that the ideal approach to this control problem, more especially in the case of turning, would be to monitor or measure the work-piece as it leaves the tool, and to control the tool or cutter from the error signal obtained by comparing this measurement with some pre-determined figure.

I realize that this approach would not be easy, and I invite the authors to comment upon these possibilities, and to say whether they have in fact considered them.

* Inspectorate of Electrical & Mechanical Equipment, Ministry of Supply.

† Electroflo Meters Ltd.

AUTHORS' REPLIES

H. J. Finden and B. A. Horlock : In reply to Dr. D. A. Bell, before a work piece can be machined it is usual to provide a jig fixture involving a minimum of two dowel locations. In conventional manually operated jig borers all boring or drilling centres are located in terms of the first centre. It is not a disadvantage to work from a reference which can be specified "x" and "y" co-ordinates each of zero. In conventional techniques the jig borer operator makes his own chart from the blueprint and his additions and subtractions from his first hole centre are possible sources of error. The inductosyn can be regarded as a precision scale and with its associated closed-loop servo all work on a jig borer can be held to dimensional tolerances higher than are likely to be required. Because of this high accuracy, conventional drawing-office practice calling for positional tolerances of groups of holes is no longer necessary.

Replying to Mr. Chapman, the reliability and accuracy of the machining of actuator gear boxes is likely to be far higher than that obtained by manually controlled jig borers as the human element is eliminated. We recommend occasional inspection, using a machine employing the inductosyn-measuring techniques.

We agree with Mr. Nettell's statement that punch cards are preferable to punch tape for work of the nature described in our paper. The standard transit mechanism enables a batch of cards to be read serially, in our case eight hole centres can be specified per card, which means that the number of cards likely to be wanted is always small.

Mr. D. R. Holloway brings up the question of tool wear and deflection. There have been a number of attempts to monitor at the actual work surface, but this is difficult because of the presence of swarf, lubricants, etc. In practice, machine stiffness must be adequate for the tolerance required and modern tools maintain long periods of consistency, in fact, our tool room advise us that if a tool wears as much as one-tenth of a thousandth of an inch it should be sent back for regrinding, and this fact is apparent because of the bad cutting qualities. Where ball-ended mills are used for contouring, finishing cuts may be necessary to obviate deflection of a tool.

K. J. Coppin : The point raised by Mr. P. D. Saw regarding wear of potentiometers is, of course, a vital one. In the equipment described, great care has been taken to design for minimum wear, the resistance values and former sizes being chosen to permit a fairly robust winding, 34 or 36 s.w.g. "Eureka" wire being used. Most alloys used for resistance wire are quite hard and durable, except in the very fine gauges, and the moving contact, in fact, wears much more quickly than the winding. For this reason we use roller contacts of coin silver on silver-graphite bearings, so mounted that they can be readily changed when noticeable wear occurs. In an assembly on which life tests are being carried out, we have found no measurable shift of balance after a million revolutions.

Mr. D. F. Nettell's remarks on the subject of punched-card systems of data storage are interesting. We feel that for inherently discontinuous processes, such as spot positioning of work for drilling or boring operations, the punched card is the most logical medium, especially as intelligent use of cards can obviate the necessity for a buffer store. The advocates of film or tape, which have undoubted advantages where continuous processes are to be controlled, are apt to press for their general adoption for the sake of standardization, but in the author's experience, slavish adherence to "standard practice" often leads to unwieldy compromise arrangements.

The remarks by Mr. D. R. Holloway on the subject of monitoring the work as it leaves the tool are, of course, specially apposite in the case of turning, boring, and shaping operations. Even in the case of a system such as I have described, which operates solely to position the work correctly with respect to the tool, uncertainty of the position and stability of the latter places a limit on the accuracy obtainable in the finished part. In long production runs, it is sometimes possible to correct automatically for "systematic" (as opposed to "random") errors, on a sampling basis, but the problem of continuously monitoring the part being produced bristles with difficulties, and an economically feasible solution in the general case would still seem to be a long way off.

R. J. F. Howard : Stylus "jittering" could undoubtedly be used to reduce stylus drag, as Mr. Williams suggests, but has not been applied in practice, mainly because of the desire to minimize complexity of stylus head and equipment design. If "jittering" were introduced the choice of amplitude and frequency would have to be carefully considered to avoid introducing an even greater reduction in performance than that due to stylus/template drag. In most instances drag has not been found to be the ultimate limiting factor in overall performance and for this reason its effects have been minimized by the more simple expedients such as careful choice and preparation of stylus tip and template materials.

In the system described, no difficulty has been encountered due to residual magnetism in the generators. In fact, the effect of residual magnetism would show itself as a small positional error, but in practice this error is extremely small and does not compare in magnitude with unwanted errors due to other causes.

Replying to Mr. T. A. Waite, electrical systems were chosen because of the ease of application without major change to the design of existing machines and because of the difficulties in obtaining adequate distances of tool travel with the more simple fluid-actuated systems. There is also some advantage in a system based on one method of operation in that the difficulties of measuring or sensing in one way and converting to a second method of power control are avoided. For certain types of machine and application, however, there are undoubtedly merits in a "mixed" system and current development does not neglect the possibilities of such arrangements.

With regard to Mr. D. R. Holloway's inquiry about tool wear, this is not, in practice, a major problem, particularly with modern tool materials. In general, wear of a significant magnitude would in any case affect the cutting characteristics of the tool and necessitate replacement or regrinding. When tool size is altered by regrinding a corresponding adjustment in the stylus tip position or size is made.

It is difficult to envisage practical solutions to the many problems of devising a control system utilizing work measurement, particularly where rapid rates of change of tool direction are involved and where larger amounts of material are being removed from the workpiece. In the latter case, the problem of the surplus material affecting the measurement seems almost insurmountable. Such a system might have some possibilities in the case of precision-finishing operations where little material is being removed and where relatively slow rate of change of tool direction is involved.

D. A. Alexander: I think Mr. Williams will find partial answers to his questions in the published version of my paper. Integration is discussed briefly, though I should add that successful use had been made of the techniques based on different methods for the integration and differentiation of unknown functions.

Toroidal transformers can be used in conjunction with ordinary analogue computing elements such as resolvers and potentiometers. However, they will work most efficiently only at certain voltages and frequencies.

Additional useful potentialities of the circuits for storing information are not developed in the paper. Briefly, it is possible to place up to some 50 secondaries on a core. The secondaries may, in fact, consist simply of wires threaded through the core: Wires threaded through combinations of cores energized from different sources will then represent different pieces of information.

The circuits employing toroidal transformers do have a general application. Complete computers based on these circuits are valuable, particularly for complex interpolation and real time working, and where electrical or mechanical outputs are required.

Theoretically, there is no reason why toroidal transformers should not have accuracies up to a few parts in 10^8. The cores, made from spiral tape, are perfectly symmetrical and of constant cross-section; end-effects are eliminated. With high-permeability cores, symmetrical windings of only a single or a few layers and very low-load currents, the leakage flux is very small. The low currents, measured in milliamperes, and thick wire cut down copper losses. For the more accurate transformers the resistance of each section is precisely adjusted. By using frequencies of the order of cycles per second and spiral strip only one to five thousandths of an inch thick, iron losses are kept very low. Furthermore, the auto-transformers frequently employed effectively have coincident primary and secondary windings and eliminate errors due to *IR* drops in the primary winding.

It has been necessary to develop special test gear for measurement. Three units are used—a potentiometer to provide in-phase voltages, another to provide small voltages in quadrature, and a tuned amplifier acting as a null-voltage indicator. The potentiometers are based on toroidal transformers and manual decade switching is employed. The in-phase potentiometers generally used have an accuracy of 2 in 10^6. The checking of the most significant digits of the potentiometers used as standards (and of very accurate transformers) involves a number of steps. One test measures the variations in the voltages between successive taps on a transformer by comparing them in turn with an isolated voltage of about the same value. Errors that are symmetrical about the mid-point are determined by reversing the polarity of the input to the potentiometer and re-measuring arbitrary voltages. Resistances that could cause errors are measured and adjusted, and consistency between potentiometers is checked.

I agree with Mr. Nettell that punched cards do provide the most satisfactory input medium in some cases. In one application made for large milling machines, each card is given a serial number and also the number of the next card, whose number is checked before its information is accepted. It is very simple to change any part of the programme: the appropriate cards are withdrawn and replaced. Information is fed to the machine at a high rate.

However, the extra cost of punched-card readers over punched-tape readers is considerable in relation to the cost of a control equipment for, say, a 30 in. × 10 in. milling machine or a jig borer. The punching equipment is also more expensive and less convenient. For milling machines, information stored on tape occupies much less space. It may be possible to put the control information for a simple drilling job on a card already containing accounting data, but the case hardly arises with the many cards required with milling machines.

For the milling machine controls described in my paper the use of a reader as a buffer store is not satisfactory, despite the fundamental attraction. Two readers are required so that information is continuously available. In addition the reader contacts are not suitable for the analogue circuits or for a number of other purposes.

Dr. Bell may be interested to know that zero shift switches can conveniently be fitted to an analogue system of control. Workpieces then need only be correctly orientated and a predetermined point picked up by rotating the switches.

With Mr. Holloway, I would very much like to take a direct measure of the work being cut. However, I do not believe we can solve the problem of direct measurement at the present time, though an expensive solution utilising X-rays or radio-active tracers may be feasible one day. Difficult servo problems might be created.

Positioning only the centre of the tool gives rise to many errors. It is possible to devise automatic circuits to overcome some troubles such as backlash and lead-screw errors. Automatic cutter-wear compensation has been provided on a number of machines; however, the variation of the mean effective cutting diameter from the nominal tool diameter must be set manually. This variation, incidentally, is usually calibrated in steps of one or two thousandths of an inch, changes that would not normally even warrant re-grinding. The effect of table and lead-screw distortions can be reduced by the proper positioning of suitable measuring devices.

Automatic compensation for other errors such such as that due to distortion of the frame of the machine tool under load would be difficult. Good programming, however, will minimize their effect. In general, the electrical accuracy obtainable in machine-tool controls is several times better than the mechanical accuracy of the tool.

C. H. Braybrook, C. R. Borley and L. Coates: In answer to Mr. Holloway, we agree in principle that the work should be monitored while machining is in progress. The difficulty is the practical one of designing a measuring head which can follow closely enough behind the tool to provide information early enough to allow corrective action to be taken. It must work in the presence of swarf, coolant and workpiece vibration and deflection. The advantages gained by solving this problem over accepting open-loop control of workpiece dimensions do not, in our opinion, warrant the effort.

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DEVELOPMENTS IN COMPONENT DESIGN

A further selection of summaries of papers read at the International Symposium on Electronic Components held in Malvern in September, 1957.

The Choice of Batteries for Use in Electronic Equipment

F. M. Booth*

Mr. Booth classified the batteries to be described as follows:

(1) Secondary batteries which are capable of giving quite high ratings for short periods; (2) Water-activated batteries which are also capable of dealing with moderately high rates of discharge for limited periods; (3) Primary batteries.

In secondary batteries the highest output available was from the silver oxide-zinc cell but the storage properties and the low temperature properties at high discharge rates were not so good. Nickel cadmium batteries are now using modern materials and he showed some batteries made with sintered plates (made from nickel powder sintered into a porous plaque and then impregnated with the appropriate salt—either nickel salt for the positive or cadmium salt for the negative). After four or five impregnations to get the requisite amount of active material deposited in the pores of the sinter it is then processed electro-chemically (formed) for the final plate. Plates can be made down to about 0·015 in. thick which, although expensive, gave this type a definite superiority over the lead acid battery at high rates of discharge.

Small pocket type batteries potted in resin and fully sealed, and others in polythene cases were also discussed. In lead-acid batteries he stressed that modern materials were very much to the forefront. Examples of these used high impact polystyrene cases and plastic grids.

The latest types of Leclanche type batteries owed much to improvements in the materials which were now available. Curves of voltages obtained on load of the electrolytic manganese dioxide showed an improvement over natural manganese dioxide. Modern examples were the mercury battery, silver/zinc batteries, lead peroxide/cadmium batteries and magnesium batteries. Brief mention was made of thermal, solar and nuclear batteries. The latter are only capable of delivering currents of the order of micromicro-amperes.

Finally, Mr. Booth stressed that comparative figures of discharge performance, etc., may be misleading and that battery requirements should be discussed with competent battery designers so that the best battery can be selected for the particular application.

* Signals Research and Development Establishment, Ministry of Supply, Christchurch.

U.D.C. No. 621.352/5

Transformer Miniaturization using Fluorochemical Liquids and Conduction Techniques

L. F. Kilham, Jnr.*

The need for miniaturization of electronic equipment has led to the development of new techniques for the design of reliable transformers of reduced size and weight. An obvious approach is to increase the maximum permissible temperature of the component, but this requires the use of materials capable of withstanding high temperatures while having good heat transfer properties and, in the case of insulating materials, good dielectric properties. The design considerations discusssed in the paper were related to a maximum temperature of 200°C. Metallic conductors inserted between windings, to carry away the heat, have aided miniaturization; and more recently the use of liquid and gaseous fluorochemicals has brought further improvement.

These fluorochemical liquids and gases, which include the perfluoro compounds $(C_4F_9)_3N$ and $C_8F_{16}O$, have been thoroughly tested for compatibility with all materials normally used in transformers operating in the region of 200°C. They are non-toxic, stable and inert. They have lower boiling points and higher specific gravities than liquids commonly used in transformers and in several important respects exhibit superior electrical properties. The most suitable coolant for

* The Raytheon Manufacturing Co., U.S.A.

U.D.C. No. 621.314.213.3

a particular case may be a mixture of fluorochemical liquids of different volatilities, or a liquid combination in conjunction with a gas combination may be chosen to give optimum heat transfer and adequate dielectric protection within desired pressure limits.

The "evaporative cooling" technique in a closed system was described. Here the transformer is enclosed in a sealed container, in which the free space is only partially filled with fluorochemical liquid. The liquid is conveyed to the windings by means of an inorganic wick. When a hot spot is reached the liquid vaporizes, evaporative cooling takes place, and the vapour rises until it comes in contact with the container wall, where it condenses and gives up its acquired heat. The author compared the latent heat of vaporization of the fluorochemicals (about 16·5 to 20·9 cal/gm) with the specific heat of conventional transformer oils (about 0·5 cal/gm), and drew the conclusion that the rates of heat transference of the two methods of cooling are in similar proportion. An example was given of a transformer cooled by direct metallic conduction to a metal heat sink, in conjunction with the vapour phase heat exchange process just described.

Figures were given to demonstrate the superiority of fluorochemical liquids over the other liquid coolants, in the case of a transformer totally immersed in liquid. An example was worked out for a transformer design which makes use of evaporative cooling, and a high voltage transformer totally immersed in fluorochemical liquid was described. The author also referred to various methods of cooling a group of components (e.g., a power supply unit), including heavy vapour cooling with forced circulation. He concluded by saying that heat transfer techniques should not be generalized, and that the method to be adopted in a particular case should be chosen carefully.

The Development of High Temperature Transformers

A. G. Gilmore*

In his introduction, Mr. Gilmore outlined the general problem of producing lighter and smaller components for ever increasing levels of ambient temperature.

He went on to describe how this problem of miniaturization had been tackled when applied to the production of a range of lightweight transformers for operation in ambient temperatures of from −60°C to +120°C, and a maximum operating altitude of 60,000 ft. Maximum coil temperature was assumed to be 250°C. For mechanical reasons, a choice of construction fell on the open type with solid impregnant. As regards the materials used, reliability and availability were over-riding factors. With this in mind, materials were selected and tested, resulting in the following choices.

Silicone resin impregnated, braided glass cloth was used for the wire covering. This material had been found to operate satisfactorily at 280°C for periods greatly in excess of 1000 hours.

For inter-coil insulation, silicone impregnated glass cloth was again found to be the most suitable material. However, in some parts of the coil, mica was used owing to its good mechanical properties.

The coils were wound on split aluminium-alloy formers to enhance the dissipation of heat, and then impregnated with a silicone varnish selected for simple processing.

The author then went on to describe in detail the methods used to reduce the temperature gradient in the transformer. With the core operating at inductions near saturation it was found that, at full thermal loading, the losses in the coil were approximately twice the dissipation in the core. It was, therefore, obvious that particular attention would have to be given to the thermal aspect of coil design. As the core is operating at a lower temperature than the coil it will tend to absorb heat from the windings, and to enable it to do this more efficiently a split metal-coil former with extended tongues making good thermal contact with the core assembly was used. The heat removed in this way has to flow through the entire depth of the winding and to reduce the resulting winding gradient a second heat conducting strip is inserted midway between the coil former and the surface.

Having extracted heat from the coil, it now remained to dissipate it to the surroundings. If a component is designed to give up its heat by convection and radiation only, then its rating will be influenced by the air pressure. At a pressure corresponding to an altitude of 60,000 ft., the heat transfer coefficient is only approximately 35% of the value at sea level, and the rating must be reduced to 75% of its output at normal pressure if the safe winding temperature is not to be exceeded.

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U.D.C. No. 621.314.213

Fortunately the cooling of the component can be considerably increased by conducting heat away from it, thereby increasing the proportion of cooling which is independent of air pressure. The provision of conductive cooling also increases the thermal rating of the component at sea level and with the component mounted on an "infinite" heat sink this increase can be as high as 70%. In practice good thermal conduction is achieved by a combined core clamp and base of light alloy, machined on the bottom to give good thermal contact to the supporting structure. The inside of the base is shaped to follow the outside surface of the coil and a similarly shaped shroud is used to provide conduction from the top of the coil.

After giving details of a range of single and three phase units, to show the saving in weight achieved by the application of the above principles, and the factors influencing the choice of a frame size for a given loading, Mr. Gilmore concluded by referring to future developments in this type of component. He mentioned the specification of ambient temperatures in the range 200-300°C and pointed out the difficulties of maintaining good regulation when confronted with an increase of winding resistance of approximately 30% at coil temperatures of 400°C. It was felt that, although some advantage could be derived from the use of silver conductors and special magnetic materials with higher saturation inductions, the most significant contribution could be made by the equipment designer in reducing his very strict regulation requirements, which all too often have limited the scope of high temperature operation.

Design, Development and Standardization of R.F. Cables

W. T. Blackband*

The first true radio-frequency cables were made about 1932 for radio relay systems and these used a rubber wax dielectric. The rubber waxes were later outclassed by polyethylene polystyrene and poly-isobutylene. These substances have excellent electrical properties, their dielectric constant is relatively low, approximately 2·3, their insulation resistivity is too high to measure easily and they are free from dielectric loss. An important step in development of modern r.f. cables came with the introduction of P.V.C. as a jacketing material. The development of r.f. cables from 1939 onwards followed different lines in the Allied and Axis countries, largely because in Britain I.C.I. had developed the first polythene and in Germany there was no polythene but a well developed production of polystyrene. The overwhelming superiority of solid polythene cables has now resulted in a truly international standard set of cables illustrated in the table of comparable types given below.

These solid polythene cables are supplemented by semi-airspaced flexible cables having fins and tube, or thread and tube of polythene for applications where low capacitance is very important. The latest form of this cable has a single piece polythene helix (British) or a helix built up of a series of overlaid polystyrene tapes (German). The losses in these cables are of the same order as those in a waveguide WG10 although their

N.A.T.O. Type	I.E.C. Type	British (Uniradio Series)	U.S. and Canadian	French	Swedish	Russian
NWR 1/S	50-7-1	67	RG-8/U	KX50MDI	HK50-7	PK-47
NWR 2	50-3-3	43	RG-58/U	—	HK50A-3	—
NWR 2/S	50-3-1	76	RG-58A/U	—	HK50B-3	—
NWR 3	75-7-3	57	—	KX75MMI	—	—
NWR 3/S	75-7-1	—	RG-11/U	KX75MDI	—	PK-20
NWR 4	—	64	RG-63/U	—	—	—
NWR 5	50-17-1	74	RG-17/U	—	HK50-17	—
NWR 6	75-17-1	77	RG-164/U	—	—	—
NWR 7	—	78	RG-133/U	KX100MMI	—	—
NWR 7/S	—	—	—	KX100MDI	—	—
NWR 8	—	87	RG-65/U	—	—	—

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cross-sections and weights are considerably less, as shown in the table below:

	Attenuation (db/100 ft. at 3000 Mc/s)	Weight (lb./ft.)	Cross-section (sq. in.)
WG 10 brass ...	1.10	2.75	4.50
WG 10 copper ...	0.56	2.60	4.50
Uniradio No. 63	3.4	0.24	0.43
HM.7	2.3	0.57	1.54

High temperature application cables using a dielectric of P.T.F.E. for temperatures up to 200°C and mineral dielectric cables for temperatures between 200° and 400°C have been developed. Neither of these cables is at present entirely satisfactory.

Present day cables also include small quantities made for specialized applications, for example:

(*a*) Low characteristic impedance
(*b*) Low capacitance per unit length
(*c*) High delay per unit length
(*d*) High attenuation per unit length
(*e*) High uniformity for use at s.h.f.
(*f*) High long term stability
(*g*) High stability of capacitance on flexing.

Mr. Blackband outlined the characteristics of these types of cable. Finally he discussed future developments of r.f. cables, in which the emphasis is likely to be upon:

(*a*) The mastery of the technique of extruding expanded polythene.
(*b*) The improvement of the uniformity of s.h.f. cables.
(*c*) Development of high temperature cables of better quality.
(*d*) Introduction of miniature r.f. cables.
(*e*) Use of copperclad steel inner conductors.

Components suitable for Automatic Assembly

J. W. Buffington*

The electronics industry afforded an excellent example of the need for automatic techniques. Traditional techniques for electronic assemblies defied the use of automatic handling methods. The printed wiring board with its orderly layout of conductors was the foundation on which the automatic fabrication of electronic assemblies has been based. Mr. Buffington said that by using multiple dip soldering the printed wiring board could reduce the labour required for soldering by 98%.

Although automatic machines could save about two-thirds of the cost of assembling small and medium products by hand, the investment for automatic assembly equipment is large and will remain high over the next five to ten years because of developmental expenses. The author instanced that a 30-component "in-line" conveyor system would cost about $175,000, a punched card programmed automatic assembly system with automatic soldering, automatic testing, would cost approximately $500,000, a semi-automatic bench machine would cost between $25,000 and $30,000.

He then discussed the many configurations and form factors of components suitable for use with these machines. He stressed the need for a set of design guide rules for component designers governing the shape, size, orienting and indexing features, leads and dimensional tolerances of components. He suggested that dimensions and spacing be based on a grid system of 0·050 in., 0·150 in. and 0·10 in. He doubted whether a 0·025 in. grid would be practical because of manufacturing tolerances and variations in the dimensions of many components.

Components could be divided into two categories—those which are to be handled by their leads and those which are to be handled by their bodies—and he described automatic handling equipment designed for these two types. He stressed the need for orienting features on components and particularly standardization of form factor, including orienting and indexing factors, as well as standardization of component sizes.

He then discussed the problem of soldering leads and, for lead held components, recommended flexible, solid, round, copper wire, 0·025 to 0·045 in. in diameter, tinned or solder coated. For body held components, round, solid, stiff wire or tubular leads which are tinned or solder coated should be used.

He concluded by saying that in order to promote the rapid achievement of the advantages of automation, components suitable for automatic assembly must be developed. We must not, however, over-emphasize automation or standardization to the extent that we sacrifice research towards technological advancement. The component designers must keep well informed as to the requirements of automatic handling.

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U.D.C. No. 621.396.69

Semi-Conductor Rectifier Developments for Power Supplies to Electronic Equipment

A. H. B. Walker & K. G. King*

The authors presented a general review of monocrystalline (germanium and silicon) and polycrystalline (selenium and cuprous oxide) semi-conductor rectifiers currently available.

Monocrystalline rectifiers. — The distinctive characteristics of germanium and silicon junction diodes as compared with the polycrystalline types were discussed. A method of construction was given and parameters to be taken into account when rating these diodes were listed.

Germanium junction diodes are characterized by low forward voltage drop, moderate reverse voltage ratings and low permissible junction temperatures. Typical figures for these characteristics were given and compared with silicon junction diodes which have higher permissible junction temperatures, larger forward voltage drops, and higher reverse voltage ratings with lower reverse currents.

The poor overload characteristics of monocrystalline rectifiers were discussed and suitable precautions against failure described.

Polycrystalline rectifiers. — Characteristics and factors to be taken into account when rating were described. The advantages and disadvantages over the monocrystalline types were given. The good overload characteristics of polycrystalline rectifiers were particularly stressed.

Finally, recent developments in selenium rectifiers, e.g. "edge-cooled" types, were described and the lecturer finished with practical comparison of the four available types of semiconductor rectifiers, taking into account efficiency, regulation, size and weight, temperature limitations, range of sizes required, life, reliability and cost.

* The Westinghouse Brake & Signal Company Ltd., London.

U.D.C. No. 621.382.2:621.314.632.4.

Automation in Component Testing

John A. Sargrove† (*Member*)

Before components are accepted for use in modern electronic equipment they are normally subjected to a rigorous series of tests, some of which are destructive, to confirm that they are of a satisfactory standard of reliability. In addition to these tests, a particular component for use in an equipment must satisfactorily pass a limited series of non-destructive tests to ensure it is up to standard before insertion in the equipment.

With the large number of components involved and the necessary testing effort required, especially with the growing complexity of modern equipments, efforts have been made over the years to introduce automatic testing to minimize the labour required and to speed up the work. Some early examples of this automation were concerned with the testing for pin holes in enamelled wire and the crystal cleavage testing for tungsten filaments. An equipment evolved by the Ministry of Supply during the last war tested simultaneously up to 1,000 vibrators under different climatic conditions. The recording of results, however, was entirely manual.

A modern example of automatic component testing equipment had been developed for the Royal Radar Establishment. With its associated pan-climatic chamber it is capable of testing up to 1000 resistors of one value from 10 ohms to 1 megohm under any required sequence of climatic cycles, with or without electrical loading on the resistors. The equipment is controlled by a 5-hole Creed tape from which 32 separate commands are encoded. The encoded tape moves forward 1/10th in. at intervals of two minutes, the drive being governed by a synchronous motor.

A test schedule lasting 12 weeks can be accommodated. Ten of the commands are used to operate 10 pre-set temperature range panels which cover the temperature from $-75°C$ to $+100°C$. Five more are used to operate a further 5 pre-set panels to provide up to 95% relative humidity from $+5°C$ to $+100°C$. Other commands are used for resistance measurement and for power loading of the resistors.

Resistance measurement can only be carried through when standard conditions ($+20°C$, 75% R.H.) have been reached in the chamber. The measurements are recorded on a chart which is divided into 100 sections, each section being allocated to one particular resistor. Synchronization between the chart and the switches selecting

† Automation Consultants & Associates Limited, London.

U.D.C. No. 621.317.73.

the resistors is effected by a photo-electric cell operating from a light source on the chart paper. Measurements are made by means of a self-balancing Wheatstone's Bridge standardized at the nominal value of the resistors under test. The resistance is recorded as a percentage deviation from standard. Eighteen seconds are allowed for the measurement of each resistor. Up to 40 measurements can be recorded during a test schedule. The accuracy is within ¼%.

Electrical loading.—The resistors are divided into four separate sections of 250, and each section is electrically loaded from a separate power unit so that resistors of four different power ratings can be tested. The applied voltage is recorded on a 4-point recorder.

Climate control.—Thermistors are used as the wet and dry bulb sensing heads which, together with the variable resistors in the various temperature humidity range panels, form one arm of a bridge circuit feeding into an amplifier arranged to operate relays which select either heat, cooling or humidity as required. The overall accuracy of temperature control is within $\pm 2°$C with a R.H. accuracy of $\pm 2\%$.

During the discussion the results of a test at R.R.E. were given. The results showed that four out of the 1000 resistors increased beyond 10% from their initial value, i.e. 0·4% failure rate. 86% of the resistors remained within 2% tolerance, the remainder spread over a gradual increase to 10% tolerance.

Components for Severe Environmental Conditions

Colonel J. S. Lambert*

The development of high-speed aircraft and missiles has created environmental conditions far more severe than those encountered in previous applications. Electronic equipments will be required to operate in high ambient temperatures which result from aerodynamic heating effects, and in intense vibration and acoustic noise fields associated with highly powered aircraft. The safe limits of conventional components have been reached and it is necessary to develop new constructional techniques and components. Although little is known as yet, tests have shown that radiation (e.g. in nuclear-powered aircraft) does adversely affect materials and components in current usage.

Three approaches to the problem of operation in severe environmental conditions are:

(*a*) The use of conventional components but providing adequate protection, such as shielding, liquid cooling, insulation, pressurizing and shock mounting. Such techniques will have limited application and would impose a weight and bulk penalty on the aircraft.

(*b*) Develop conventional components to their limits of operation; this would involve an extensive programme of investigation.

(*c*) Develop entirely new materials, devices and design concepts. The most difficult problem to solve is that of temperature, but it is thought once this parameter is met the others will also be solved (e.g. ceramic materials which meet temperature requirements are tolerant to nuclear radiation). This approach, however, involves the development of components for operation at temperatures up to $+500°$ C. The activities described in this field include ceramic valves, high-temperature transistors and transformers, wire-wound and metal-film ceramic resistors, the investigation of dielectric materials such as magnesium oxide, alumina and boron nitride, and ceramic-coated cables.

It is doubted whether high-temperature circuitry would be satisfactory for operation at room temperature due to temperature co-efficient and voltage stabilization problems. Circuit pre-heating before use would be necessary.

In summarizing it was considered that cooling, shielding and protecting electronic components from severe environmental conditions are only stop-gap methods. In several years' time, high-temperature components would become available giving more flexibility in equipment design, allowing equipments to be operated under extreme conditions.

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U.D.C. No. 621.396.69

Note.—Summaries of the remaining four papers will be published in the May issue of the *Journal.*

. . . Radio Engineering Overseas

A selection of abstracts from European and Commonwealth journals received in the Library of the Institution. Members may borrow these journals under the usual conditions. All papers are in the language of the country of origin of the journal unless otherwise stated. The Institution regrets that translations cannot be supplied.

621.317.32.015.34:621.369.67

Correlation measurements in the short-wave range.—J. GROBKOPF, J. SCHOLZ and K. VOGT. *Nachrichtentechnische Zeitschrift*, **11**, pp. 91-99, February 1958.

The results from measurements of the correlation conditions for the voltages received by two aerials are reported. The two aerials have been installed at right angles to, as well as in line with the direction of propagation and have been crossed as well.

621.372.51.029.6

Matching elements for a wave guide directional coupler with periodic structure.—I. LUCAS. *Archiv der Elektrischen Ubertragung*, **12**, pp. 91-96, February 1958.

For directional couplers with periodically arranged coupling elements the matching conditions are stated and rules given for proportioning the coupling matching elements at either end of the coupling section. With one matching element at either end the reflection can be cancelled for a single frequency. When two matching elements are employed, the derivative of the reflection coefficient with respect to frequency disappears also. The width of the band over which a matching condition exists is calculated for one and two such matching elements at either end. Measurements on a directional coupler with an even distribution of the power to both outlets suggest that a single element at either end will be adequate. This matching method can be quite generally adopted on waveguides with periodically distributed discontinuities.

621.375.018.756

A summary of the theoretical basis for broadband distributed amplifiers suitable for very short pulses.—D. DOSSE. *Nachrichtentechnische Zeitschrift*, **11**, pp. 61-68, February 1958.

An explanation of the general requirements, such as pulse shape and pulse spectrum as well as smallest possible pulse distortion which must be met by distributed amplifiers as low-pass systems, is followed by a description of the operation and delay lines in the amplifier. The transfer characteristic of a distributed amplifier is derived under the assumption of lossless valves. Statements concerning the type and magnitude of losses in valves precede a derivation of a transfer characteristic with valve losses taken into consideration. The possibility of normalization of this characteristic is shown and the characteristic is evaluated for the purpose of obtaining design formulae.

621.375.4

Transistors for medium frequency—a detector circuit for radio-communication apparatus. G. ROSIER. *Tijdschrift van het Nederlands Radiogenootschap*, **23**, pp. 9-16, January 1958.

Application of transistors in professional equipment, which generally has to meet stringent requirements, often leads to a development of circuitry which differs from the customary tube technique. To illustrate this a m.f. amplifier is described, of which the selectivity is determined by a separate filter at the input of the amplifier, while the amplifier stages proper are R-C coupled. A detector circuit is also indicated, in which the transistor is used as a diode detector and as an amplifier at the same time.

621.382.3

Manufacturing principles of transistors. J. LANTIERI. *L'Onde Electrique*, **38**, pp. 18-28, January 1958.

The problem is to define the physical conditions which must be realised in a transistor to obtain the desired electrical characteristics. The author describes the principles of the functioning of a transistor, and defines the physical parameters which must be verified as they determine the following electrical characteristics in amplification, frequency range of use, power rating as well as theoretical limitations.

In the second part some conventional manufacturing techniques are described in their different stages: drawing, alloying, diffusion.

621.382.3

High frequency germanium and silicon transistors. J. DEZOTEUX. *L'Onde Electrique*, **38**, pp. 29-35, January 1958.

After a general description of the development of transistors, the author discusses the parameters which are important at high frequencies, taking the equivalent circuit as the basis, and further indicates the direction of development to improve these parameters (structure of transistor with respect to internal field). The second part is devoted to the consideration of transistor circuits. The definitions of cut-off frequency and merit factor are re-stated. The parameters which determine the performance of transistors in each of their three major applications as amplifier elements, that is to say, wide-band amplifiers, tuned amplifiers and oscillators are indicated. Some values of these high frequency parameters are given in tabular form for various industrial and laboratory transistors. The technique of manufacture by diffusion applied to silicon should lead to the solution of the problems of power, frequency and temperature.

621.382.3

Stabilization of the operating point of transistors. E. E. P. POELMAN. *Tijdschrift van het Nederlands Radiogenootschap*, **23**, pp. 1-7, January 1958.

In professional applications it is necessary to design a transistor circuit carefully with respect to changes in ambient temperature. Stabilization of the operating point by feedback is preferable, the latter being maximum for direct-current and having the appropriate value in the frequency-range involved. One correct and two incorrect methods of feedback are given. An amplifier is shown, in which an amplified direct-current-feedback is applied. A compact oscillator based on the outlined principle, is discussed.

621.386.82:621.397.62

Dosimetry of the very weak X-radiation generated in television receivers and X-ray diffraction apparatus.—W. J. Oosterkamp, J. Proper and J. J. F. de Wijk. *Philips Technical Review*, **19**, pp. 264-267, March 1958.

Television picture tubes emit very soft, extremely weak X-radiation, which can be detected at the outside surface of a home television receiver. To preclude all danger for the user, the dose rate according to international recommendations should not exceed 2 milliroentgens per hour (in the future the permissible limit may well be set still lower). The dose rate can be checked with thin-windowed Geiger-Muller counters, whose windows are sufficiently transparent to the soft radiation. The counter tubes cannot be directly calibrated with the standard ionization chambers normally used for therapeutic dosimetry, since the latter are not sufficiently sensitive. For this reason a substandard chamber with a large measuring volume has been designed, which acts as an intermediary in the calibration For protection measurements on X-ray diffraction apparatus, with which dose rates outside the useful X-ray beam of 6 mr/h are permissible, ionization chambers are more suitable, although the counter tube here too is useful for giving a picture of the exact dose-rate distribution around the apparatus.

621.396.11:621.397

The influence of multipath propagation on the spectrum of a received television signal.—K. Bernath and H. Brand. *E.B.U. Review*, Part A, No. 47, January 1958.

When the difference in the length of the paths is significant compared with the wavelength, the result is fading which affects the various frequencies of the vision band in different ways. Distortion caused in this way can obviously not be corrected at the receiver. The paper deals with the subject from the theoretical point of view by comparing the theory with measured results.

621.396.969:551.508.8

A radar sonde system for upper air measurements.—N. E. Goddard and H. A. Dell. *Philips Technical Review*, **19**, pp. 258-263, March 1958.

Short description of a radar sonde system, for measurements of wind speed, wind direction, temperature, pressure and humidity up to high altitudes. A telemetering thermometer, barometer and hygrometer, together with a transmitter, receiver and encoder, are carried by a free, hydrogen-filled balloon. The receiver is periodically interrogated by pulses transmitted from the ground station and the airborne transmitter responds with a signal which is used to determine the balloon's position. Its instantaneous speed and direction of movement (i.e. the wind vector) are determined and recorded by an automatic wind computer. The same signal also carries the temperature, pressure and humidity information coded as pulse delays; these are decoded and the readings continuously recorded at the ground station.

621.397.331.24:621.385.832

Television picture tubes for cathode modulation with increased effective perveance.—W. Niklas, C. Szegho and J. Wimpffen. *Archiv der Elektrischen Ubertragung*, **12**, pp. 54-60, February 1958.

Analytical expressions for the grid modulation characteristic and the perveance of picture tubes are derived for the paraxial space after explaining the difference between cathode and grid modulation. "Effective perveance", related to the perveance as expressed by Child, Langmuir and Schottky, represents a practical design parameter for television picture tubes. Expressions for the beam current characteristic and perveance are derived analytically in the paraxial space for cathode modulation and the relations extended empirically to the marginal space. The influence of the high voltage penetration on the perveance is also investigated. The requirements for a new picture tube designed for optimal utilization of cathode modulation are listed. The electron gun of the new tube is discussed in detail. A special feature is that the second grid of this gun possesses a smaller aperture than the first grid. In addition, the new tube is approximately 2 in. shorter than the conventional tube. Results obtained with the improved tube are listed. It is shown that the prevailing smaller cut-off voltage range can be realized in mass production without difficulty. Spot diameter, current defocusing, and focusing voltage range of the new tube are entirely comparable with corresponding characteristics of a conventional tube while transconductance and maximum beam current are substantially higher. An increase of the effective perveance to 5·1 microamps/$V^{1\cdot5}$ has been achieved without deterioration in resolution.

621.397.743:621.397.813

Non-linear distortions in television links. J. Muller. *Archiv der Elektrischen Ubertragung*, **11**, pp. 485-494, December 1957.

The definition of non-linear distortion with television programme transmission systems and the limit of the perceptibility of gamma distortion in the television picture, and the non-linear distortions of different types of television programme line are reviewed and their influence on the picture quality is discussed. The discussion relates in particular to the kind of frequency-dependent non-linear distortions that may appear on frequency-modulated radio links. Emphasis is here placed less on a clarification of their mechanism than on their effect on the video signal and the television picture by reference to oscilloscope records and diagrams. Results are given of measurements of non-linearities of television programme lines and the new 4-kMc/s radio link and 1-Mc/s cable system.

621.397.813

The cause and the measurement of level sensitive phase and amplitude fluctuations in the transmission of colour-television chrominance carriers.—J. Piening. *Nachrichtentechnische Zeitschrift*, **11**, pp. 70-77, February 1958.

The most important causes of differential gain and differential phase in colour-television transmissions are indicated and investigated quantitatively. A brief description of equipment for measuring these distortions is given and practical results of measurements are reported.

681.142:621-52

The theory and design of a sampled data control systems.—S. Bellert. *Rozprawy Elektrotechniczne*, **3**, pp. 472-530, No. 4, 1957.

The fundamentals of the theory of feed-back control systems working on sampled data are given. The theory is a continuation of the method given by J. Z. Tsupkin, but is not based on the "discrete," but on the integral Laplace transformation. A number of concepts well known from continuous working systems is utilized, i.e., the concept of the transfer function, frequency response, time response, etc.