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Laboratory Automation

ELECTRONICS has been affectionately referred to by a Past President of this Institution† as the greatest intellectual nosey-parker of all time—it is a technology which finds some application in virtually any scientific or engineering activity that one cares to call to mind. It is perhaps rather surprising that this enormous widening of the application of electronic devices did not really extend much beyond communications until after the Second World War; however, during the 'forties and 'fifties electronic techniques were adapted to a seemingly endless stream of uses, and the invention of the transistor and the subsequent development of semiconductor devices of great versatility have added to the breadth of that stream.

For nearly all these developments, instrumentation and control are the basic disciplines and it is possible to distinguish between their applications to the production of materials or manufactured goods and to the production of what may loosely be termed 'information'. A clearly defined area of this latter field is the laboratory or observatory where physical and chemical properties are determined. The Institution's Instrumentation and Control Group Committee identified two particular aspects of what it decided could be termed 'Laboratory Automation'—the title given to the next I.E.R.E. Conference—namely Automatic Analysis and Computer-controlled Experiments.

The work of typical automatic analysis equipment is to carry out a large number of repetitive tests with consistent accuracy and usually at high speed, thus enabling technical staff to concentrate on tasks which cannot be made automatic. Computer-controlled experiments represent a way of carrying out investigations which often simply cannot be done without some techniques for handling and processing large quantities of data in a realistic time. There is obviously a merging of these two approaches in many of the applications which are being categorized as 'laboratory automation' but it will clearly be seen that they all have the same general aim as, and many similarities to, the automation techniques familiar in the process and manufacturing industries.

The Conference which is being held in London on 10th, 11th and 12th November is divided into three sessions:

1. Automation in Chemical, Bio-chemical and Nuclear Laboratories;
2. Automatic Methods in Spectrometry;
3. Automation in Mechanical, Electronic, Electrical and other Laboratories and Observatories.

A list of the 31 papers is given elsewhere in the issue and it will be seen that their origins are spread over government organizations, university departments and industry, both users and manufacturers being represented. There are also three papers from outside Great Britain. Various techniques in spectrometry, automation as applied in electronics laboratories, and techniques relating to mechanical engineering laboratories, give some idea of the wide coverage of the programme.

Because of the wide ranging nature of the applications it is certain that there will be 'cross fertilization' of ideas between the different disciplines. This should be the greatest attraction of such a conference—techniques applicable in, say, chemical laboratories can well provide an inspiration for solving a thorny problem in mechanical engineering. While it is certainly a 'nosey-parker' in the sense that Professor Emrys Williams had in mind, electronics provides a common and relatively new forum which few other technologies can provide in promoting exchanges between scientists and engineers of all disciplines.

G. S. EVANS

† Williams, Emrys, 'Presidential address,' *The Radio and Electronic Engineer*, 33, No. 1, pp. 1–8, January 1967.

INSTITUTION NOTICES

C.E.I. Graham Clark Lecture 1970

The 16th Graham Clark Lecture will be given on Tuesday, 17th November 1970 by Sir Henry Jones, K.B.E., Chairman of the Gas Council.

The title of Sir Henry's Lecture will be 'Engineers in a changing world', and he will discuss the ways in which engineers should be ready to take responsibilities outside the field of engineering itself if the need arises—they will increasingly occupy posts in management. Recognition must also be given to the importance of the non-professional engineer and craftsman upon whom the success of the professional is dependent. Our technological society would make faster progress if more members of the community possessed some scientific or technical knowledge and this will entail the development of a more literate, numerate and mechanically apt population for which an improvement in the quality of education is essential. Finally, emphasis is given to the important part that the engineer has to play in adaptation to technological change, and the way in which engineers employed in the gas industry are adapting themselves to change is described.

The Lecture will be given at the Institution of Civil Engineers, Great George Street, London, S.W.1, and will start at 6 p.m. (tea from 5.30).

I.E.R.E. Conference on Mechanical Handling—Preliminary Announcement

This Institution, with the association of several other Institutions and organizations, is holding a Conference on 'Electronic Control of Mechanical Handling' at the University of Nottingham from 6th to 8th July 1971. Further details, including the 'call for papers', will be published shortly.

Conference on Aerospace Antennas

Organized by the Electronics Division of the Institution of Electrical Engineers in association with the Institution of Electronic and Radio Engineers, this Conference, to be held in London from 8th to 10th June 1971, is intended to be of interest to those responsible for the design, installation and use of radio and radar antenna systems for aircraft and other aerospace vehicles.

Both military and civil applications will be considered and the following aspects will be covered:

Radomes; Environmental problems; Electronic scanning; Aerials with built-in electronics; Multi-purpose antennas; Special problems of helicopters and light aircraft; Radiation patterns of electromagnetic compatibility; Design of antennas for aeronautical satellite systems; Design of h.f. antennas; Fixed and scanning radar systems; Measurement of antenna performance; Antenna system components; Problems of commissioning and maintenance.

The Organizing Committee invite offers of contributions not exceeding 3000 words for consideration for inclusion in the Conference programme. Those intending to offer a contribution should submit a synopsis (approximately 250 words) to the Conference Department, I.E.E., Savoy Place, London, WC2R 0BL, without delay. Complete manuscripts of papers will be required for assessment not later than 15th February 1971.

Registration forms and further details will be available a few months before the event and may be requested from the above address.

Norman Hayes Memorial Award

This Institution assists the Institution of Radio and Electronic Engineers of Australia by making recommendations in alternate years (with the Institute of Electrical and Electronic Engineers of America) for the Norman Hayes Memorial Award that is presented annually to the author of the most outstanding paper published in the Proceedings of the Australian Institution.

The 1970 Award for papers published during 1969 is to be made to Mr. A. N. Thiele for his paper entitled 'Horizontal Aperture Equalization'.

The I.E.R.E. and the I.R.E.E. Australia have an agreement for mutual reprinting of outstanding papers from each other's publications and Mr. Thiele's paper therefore appears in this issue of *The Radio and Electronic Engineer.*

The Quebec Section

Members may like to note the following visits which are being arranged by the Committee of the Quebec Section:

4th November—Television Studios at the C.B.C. International Broadcasting Centre, Cité du Havre.

3rd February—Radio, electronics and flight simulator facilities of Air Canada at Dorval Airport.

2nd April—Facilities, including the sound reproduction equipment at the National Film Board Studio, 315 Côte de Liesse, Montreal.

These are all evening functions starting at 7.30 p.m. Further information may be obtained from the Meeting Secretary of the Quebec Section, Mr. K. E. Hancock, c/o Canadian Marconi Company, 2442 Trenton Avenue, Montreal 301, P.Q.

U.D.C. 621.396.969.34

The Estimation of Loss of Echoing Area with Very High Resolution Radars

By

D. C. COOPER,
Ph.D., C.Eng., M.I.E.E.†

Presented at an Aerospace, Maritime and Military Systems Group Symposium on 'High Resolution Radar Systems' held in London on 8th April 1970.

The paper considers the use of recently published experimental data in estimating the manner in which the echoing area of an aircraft is reduced when the range resolution cell of the observing radar is smaller than the aircraft dimensions. A simple mathematical model of the echoing process is postulated and the experimental results are used in selecting the model parameters. The model is then used to predict the behaviour of the echoing area, and in addition, the echoing properties for a resolution cell which embraces the whole aircraft are shown to be in excellent agreement with the widely accepted 'one dominant' model proposed by Swerling.

List of Principal Symbols

A amplitude of dominant reflexion component

x_1, y_1 quadrature reflexion components for short pulse length

x_3, y_3 quadrature reflexion components for arbitrary pulse length

R_1 peak video signal for short pulse length

R_2 peak video signal for arbitrary pulse length

$z = R_1/R_2$ peak amplitude quotient

σ_1 root mean square value of x_1 and y_1

σ_3 root mean square value of x_3 and $y_3 = \sqrt{\sigma_1^2+\sigma_2^2}$

$$k_1 = \frac{\sigma_1}{\sigma_2}$$

$$k_2 = \frac{A}{\sigma_2}$$

$$u = \frac{R_2}{\sigma_1 \sigma_2 \sqrt{2}} \{(z^2+1)\sigma_2^2+\sigma_1^2\}^{\frac{1}{2}}$$

$$p = \frac{z\sigma_2^2}{(z^2+1)\sigma_2^2+\sigma_1^2}$$

$$q = \frac{A\sigma_1}{\sigma_2\sqrt{2}\{(z^2+1)\sigma_2^2+\sigma_1^2\}^{\frac{1}{2}}}$$

$I_0(a)$ Bessel function with imaginary argument

$$= \sum_{n=0}^{\infty} \left(\frac{a}{2}\right)^{2n} \frac{1}{n!\,n!}$$

${}_1F_1(\alpha, \beta, \gamma)$-confluent hypergeometric function

$$= 1 + \frac{\alpha}{\beta}\gamma + \frac{\alpha(\alpha+1)}{\beta(\beta+1)}\frac{\gamma^2}{2!} + \frac{\alpha(\alpha+1)(\alpha+2)}{\beta(\beta+1)(\beta+2)}\frac{\gamma^3}{3!} + \ldots$$

† Department of Electronic and Electrical Engineering, The University of Birmingham, P.O. Box 363, Birmingham 15.

1. Introduction

Recently Davies and Bromley[1] have reported the results of some experiments which were used to provide some indication of how the echoing area of an aircraft varies when the range resolution cell of the observing radar is improved beyond the point at which it is comparable with the overall dimension of the aircraft. The experiments were conducted with a pulse compression radar system equipped with two receiving channels of different bandwidths which therefore gave different values of range resolution, or effective pulse length, at their outputs. In this way the target echoing properties for two sizes of resolution cell could be measured at the same time, thus avoiding difficulties which would have been encountered with separate measurements taken on the fluctuating echo.

As a matter of convenience measurements were always conducted with one receiver channel operating at full bandwidth, or minimum resolution cell, and the other at some reduced bandwidth giving poorer esolution.

In assessing the results of observations it was noted that the ratio of the echo signals in the channels was a parameter that would not be affected by changes in the received signal energy, perhaps caused by the aircraft not being tracked perfectly and certainly to be expected as the aircraft range changes. Since the returns in each channel were quite detailed in the range region embracing the aircraft, a fairly arbitrary decision was made to take the peak levels as the quantities whose quotient was formed. It was found that, for the head-on, or tail-on, aircraft aspects used in the majority of the observations, the peak levels were generally in a region which corresponded to dominant echoing parts of the aircraft, such as the engine air intakes and wing root.

Using a large number of independent observations on any one type of aircraft Davies and Bromley were able to find a mean or average value for the ratio of the peak echoes corresponding to two different values of range resolution or pulse length and this peak echo quotient was used to give a direct indication of the ratio of the actual echoing areas for the two pulse lengths.

The use of the peak echo quotient in the above way is not strictly permissible because the peak echoes used to form the quotient cannot be considered to be statistically independent. The actual echoing areas are the mean values of the peak echoes and the mean value of the quotient of two random variables is not equal to the quotient of their mean values unless the variables are statistically independent. This paper outlines an alternative and improved method for using the results given by Davies and Bromley to provide an indication of the way in which the actual echoing area of an aircraft is reduced when the range resolution cell, or the pulse length, of the observing radar becomes smaller than the aircraft dimension. A simple mathematical model is used to represent the echoing properties of an aircraft and the statistical properties of the peak echo quotient which corresponds to the model are compared with the practical results. In this way the best values for the parameters used in the model can be selected and the actual echoing properties may then be evaluated.

2. The Simple Mathematical Representation of Aircraft Echoes

The proposed model can be visualized as comprising a reflecting object of fairly large echoing area but small physical dimensions, for example, a corner reflector, surrounded by a large number of items of small echoing area distributed over a region corresponding to the dimensions of the actual aircraft. Such a model is somewhat idealized but it is well known that aircraft do give echoes which have a dominant part and a fair number of small components from various parts of the structure.

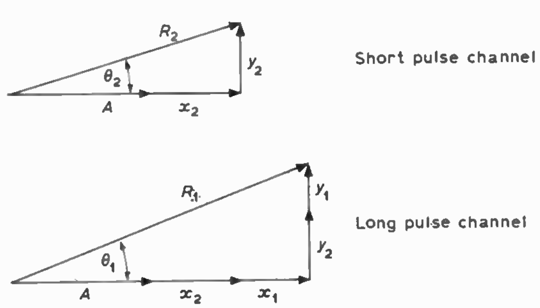

Fig. 2. Phasor diagram for the two pulse lengths.

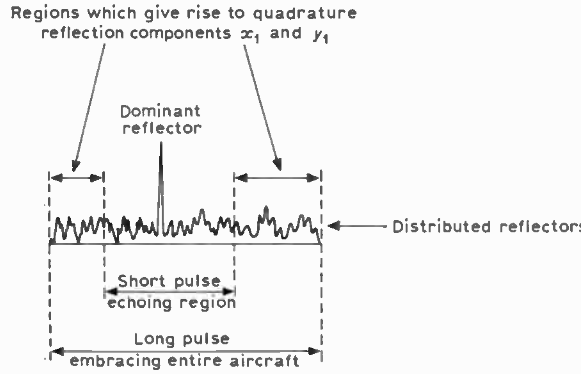

Fig. 1. An illustration of the mathematical model.

In dealing with the echoes produced by observing the aircraft with a radar using two separate receivers giving different resolution it is convenient for us to consider the situation depicted in Fig. 1. In this figure we have indicated the two regions of the aircraft which we shall consider to contribute to the peak signal in the two receiver channels. It will be observed that in each case the dominant part of the aircraft is assumed to be within the region and although this situation could conceivably not result in a peak level such an event will occur very infrequently when the dominant echo is fairly large. It will be shown that the model does fit the experimental results only if a large dominant part is selected.

The total or peak returns corresponding to the two resolution cells can now be represented by the use of a phasor diagram. This is shown in Fig. 2 in which it will be noted that each channel of the radar has three common components, a component due to the dominant reflector and two quadrature components arising from the distributed reflecting items within the common small resolution cell. The total return for the channel with the larger resolution cell is just that for the small cell plus two extra quadrature components which arise from the reflecting items in the region between the smaller resolution cell and the larger one.

It is assumed that the reflecting items are randomly distributed with constant average density within dimensions corresponding to those of the aircraft, and for a large number of items the movements of the aircraft causing changes in the total echoes will produce random changes in the quadrature components of the phasor representations. Movements such as yaw and pitch will change the relative phases of all the components produced by the distributed small reflecting objects and such changes will tend to give quadrature components that have Gaussian probability density functions (p.d.f.s.). The dominant component is assumed to be unaffected by movement.

At this stage the phasor diagrams for the two channels contain only a fixed phasor due to the dominant reflexion and random quadrature components with Gaussian p.d.f.s. It is now suggested that the variances of the quadrature components will be proportional to the number of contributions producing them and hence to the size of the regions, in the reflecting object space, with which the components are associated.

The model is now complete and conveniently expressed in the form of the phasor representations of Fig. 2. One cannot fix a value for the magnitude of the phasor representing the dominant reflexion but it is postulated that the remaining phasors have Gaussian p.d.f.s and their variances are directly proportional to the size of the reflecting regions to which they correspond.

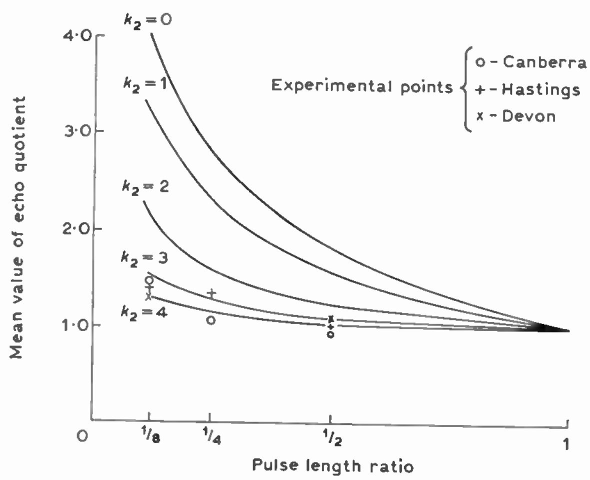

Fig. 3. Mean echo quotient vs. pulse length ratio.

3. Selection of the Model Parameters

A precise analysis as given in Appendix 1 shows that the joint p.d.f. for the peak video echoes R_1 and R_2 produced by the simultaneous use of a long and a short radar resolution cell or pulse, can be obtained in closed form (Appendix 1, equation (21)).

The quotient of R_1/R_2 corresponds to the ratio used by Davies and Bromley, and the p.d.f. for this quotient z, evaluated for the mathematical model, is obtained in Appendix 2. Equations (26) and (27) indicate that,

$$p(z) = \frac{2p^2k_1^2}{z^2}\exp\left\{q^2 - \frac{k_2^2}{2}\right\} \times$$

$$\times \sum_{n=0}^{\infty} \frac{(2n+1)!\,p^{2n}}{n!\,n!}\;{}_1F_1(-2n-1, 1, -q^2) \quad \ldots(1)$$

where

$$p = \frac{z}{z^2+1+k_1^2}$$

$$q = \frac{k_1 k_2}{\sqrt{2}(z^2+1+k_1^2)^{\frac{1}{2}}}$$

$$k_1 = \frac{\sigma_1}{\sigma_2}$$

and $$k_2 = \frac{A}{\sigma_2}$$

The parameter k_1 can be related to the radar resolution cell sizes, or the effective pulse lengths, since it has previously been postulated that σ_1 and σ_2 will be determined by the size of the regions in the array of scattering objects, whose echoes are represented by the quadrature components.

Hence

$$\frac{\sigma_1^2+\sigma_2^2}{\sigma_2^2} = \frac{\text{size of large cell}}{\text{size of small cell}}$$

$$= \frac{\text{bandwidth for small cell}}{\text{bandwidth for large cell}}$$

and

$$k_1^2 = \frac{\text{larger bandwidth}}{\text{smaller bandwidth}} - 1 \quad \ldots\ldots(2)$$

Using equation (2) to determine the parameter k_1 and taking values for k_2 of 0, 1, 2, 3, and 4, the p.d.f.s for the peak echo quotient z have been evaluated with the help of a digital computer. In addition to producing the p.d.f.s the computer was programmed to evaluate the mean or average value of z for each case since the mean value of the peak echo quotient was the quantity used by Davies and Bromley in estimating the loss of echo resulting from a reduction in radar pulse length.

The results of the mean value calculations are presented in Fig. 3 in which the ratio of channel bandwidths, or pulse lengths, is used as the abscissa. Also shown in Fig. 3 are the experimental points obtained by Davies and Bromley and it will be noted that these points indicate that a value of k_2 in the neighbourhood of 4 is an appropriate choice for the mathematical representation of all the aircraft used in the experiments.

4. Reduction of Echoing Area when Pulse Length is Short

The probability density function for the peak video signal, which is the magnitude R_1 shown in Fig. 4, can be obtained in the following well-known form:

$$p(R_1) = \frac{R_1}{\sigma_3^2}\exp\left[-\frac{R_1^2+A^2}{2\sigma_3^2}\right] I_0\left[\frac{AR_1}{\sigma_3^2}\right] \quad \ldots(3)$$

The mean square value of this peak signal is found to be

$$\overline{R_1^2} = A^2+2\sigma_3^2 \quad \ldots\ldots(4)$$

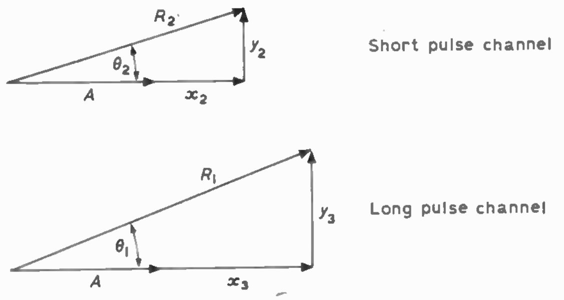

Fig. 4. Simplified phasor diagrams for the two pulse lengths.

and the maximum value obviously occurs when σ_3 takes its maximum value of $\sigma_{3(\max)}$, the pulse length then being equal to the aircraft dimension.

We now obtain the ratio of the mean square peak signal to its maximum value, and this will give us the ratio of the actual echoing area to its maximum value. Thus we have,

$$\frac{\overline{R_1^2}}{\overline{R_{\max}^2}} = \frac{A^2+2\sigma_3^2}{A+2\sigma_{3(\max)}^2} = \frac{\left[1+\frac{2\sigma_{3(\max)}^2}{A^2}\left\{\frac{\sigma_3}{\sigma_{3(\max)}}\right\}^2\right]}{\left[1+\frac{2\sigma_{3(\max)}^2}{A^2}\right]} \quad \ldots(5)$$

Since the maximum pulse length that has been considered is 8 times the shortest value and the quadrature components used in the model have variances which are assumed to be proportional to the pulse lengths, we have, $\sigma_{3(\max)}^2 = 8\sigma_2^2$. In addition it has been observed that a reasonable value for $k_2 = A/\sigma_2$ is close to 4 and using this value we have $A^2/\sigma_{3(\max)}^2 = 2$

Using the above values,

$$\frac{\overline{R_1^2}}{\overline{R_{\max}^2}} = \frac{1}{2}\left[1+\frac{\sigma_3^2}{\sigma_{3(\max)}^2}\right]$$

$$= \frac{1}{2}\left[1+\frac{\text{effective pulse length}}{\text{max. effective pulse length}}\right] \quad \ldots(6)$$

and we note that the echoing area for the model falls off linearly with the shortening of the effective pulse length. If the effective pulse length used is one half of the maximum the loss of echoing area will be approximately 1·25 dB and we see that this result differs somewhat from the figure of 0·5 dB obtained for this case by Davies and Bromley. Obviously further experimental work would have to be done to determine the precise figure and aircraft types would have to be considered separately.

It is interesting to note that for very short pulse lengths the loss of echoing area indicated by equation (6) will be approximately 3 dB and this is close to the values given by Davies and Bromley. The echoes will be virtually uncorrelated in this case and the above agreement is not too surprising. Equation (6) gives a very simple relationship for the fall-off in echoing area when the effective pulse length is smaller than the aircraft dimensions, and in the absence of more detailed experimental evidence it is probably the most accurate expression available.

5. Fluctuation of the Overall Aircraft Echo

The probability density function for the peak video level in the case of a pulse which embraces the entire aircraft is, for the model, given by

$$P(R_{\max}) = \frac{R_{\max}}{\sigma_{3(\max)}^2}\exp\left[-\tfrac{1}{2}\left(\frac{R_{\max}^2+A^2}{\sigma_{3(\max)}^2}\right)\right] I_0\left[\frac{AR_{\max}}{\sigma_{3(\max)}^2}\right] \quad \ldots\ldots(7)$$

where an acceptable value for A is given by $A^2 = 2\sigma_{3(\max)}^2$

In this case it can be shown that the mean value of $R_{\max}$ is

$$\overline{R}_{\max} = \sigma_{3(\max)}\sqrt{\frac{\pi}{2}}\,{}_1F_1(-\tfrac{1}{2}, 1, -1) \simeq 1{\cdot}812\sigma_{3(\max)} \quad \ldots\ldots(8)$$

and the mean square value of $R_{\max}$ is,

$$\overline{R_{\max}^2} = 4\sigma_{3(\max)}^2 \quad \ldots\ldots(9)$$

Considering now the 'one dominant' probability density function proposed by Swerling[5] for the representation of target echo fluctuations the peak video level R is postulated to have p.d.f.

$$P(R) = \frac{R^3}{2\sigma^4}\exp\left\{-\tfrac{1}{2}\frac{R^2}{\sigma^2}\right\} \quad \ldots\ldots(10)$$

and it is not difficult to show that,

$$\overline{R} = \frac{3}{2}\sqrt{\frac{\pi}{2}}\,\sigma = 1{\cdot}88\sigma \quad \ldots\ldots(11)$$

and

$$\overline{R^2} = 4\sigma^2 \quad \ldots\ldots(12)$$

Comparing equations (8) and (11) we note that for the same average received power, that is, for $\sigma_{3(\max)}^2 = \sigma^2$, the mean values of the video waveforms are only about 4% different. Hence the mathematical model gives results which are consistent with Swerling's 'one dominant' model and this latter model has been widely utilized in radar system studies for describing the fluctuation of echoes obtained from all parts of an aircraft.

6. Conclusions

The simple mathematical model that has been used to represent the echoing properties of aircraft has been shown to be of value in situations where the range resolution cell of the radar is smaller than the aircraft dimensions. This situation is particularly important because modern radars can have very good range resolution and the radar designer needs to know whether or not the detection of an aircraft in a background of clutter will be improved by the use of better range resolution. The background clutter power is in general directly proportional to the size of the radar resolution cell so that halving the radar pulse duration reduces the clutter level by 3 dB.

The experimental results given by Davies and Bromley have led to the setting-up of a simple mathematical model and this has been used to obtain a prediction of the way in which the aircraft echoing area falls when the radar range resolution cell is reduced. This loss of echoing area is shown to follow the extremely simple law given by equation (6). It is found that a loss of about 1·25 dB results if only one half of the aircraft is covered by the pulse at any given time. Hence we may conclude that a reduction of pulse length should lead to an improved echo-to-clutter ratio.

The mathematical model has been shown to be consistent with the well-known Swerling 'one dominant' description for the overall echoing properties of an aircraft and this is considered as a further justification for the use of the model.

7. Acknowledgment

This paper is contributed by permission of the Director of R.R.E. and the Copyright Controller H.M. Stationery Office.

The author is also indebted to Mr. B. E. Callan, who devised the computer programmes used in the work.

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9. Appendix 1: The Statistical Properties of the Peak Echoes

It is convenient to modify slightly the phasor diagrams shown in Fig. 2 to obtain the form shown in Fig. 4. All that has been done is to introduce a new pair of random quadrature phasors x_3 and y_3 which are defined by the equations

$$x_3 = x_1 + x_2 \qquad y_3 = y_1 + y_2 \qquad \ldots\ldots(13)$$

Since x_1, x_2, y_1 and y_2 are independent Gaussian variables it is obvious that x_3 and y_3 will be independent with Gaussian p.d.f.s but x_3 will be correlated with x_2, and y_3 will be correlated with y_2.

Thus the joint p.d.f. for the random variables x_2, x_3, y_2, y_3 can be expressed in the form

$$p(x_2, x_3, y_2, y_3) = p(x_2, x_3)\ p(y_2 y_3) \qquad (14)$$

The joint probability density function for x_2 and x_3 will be of bivariate normal form[2] so that

$$p(x_2, x_3) = \frac{1}{2\pi B} \exp\left[-\frac{1}{2B^2}\{x_2^2\sigma_3^2 - 2x_2 x_3 \overline{x_2 x_3} + x_3^2\sigma_2^2\}\right] \qquad \ldots\ldots(15)$$

where $B^2 = \sigma_2^2\sigma_3^2 - (\overline{x_2 x_3})^2$

σ_3 = r.m.s. value of x_3

σ_2 = r.m.s. value of x_2

and $\overline{x_2 x_3}$ = mean value of product $x_2 x_3$.

It is easily shown that $\overline{x_2 x_3} = \overline{(x_1+x_2)x_2} = \sigma_2^2$ so that

$$p(x_2, x_3) = \frac{1}{2\pi\sigma_1\sigma_2} \exp\left[-\frac{1}{2}\left\{x_2^2\left(\frac{1}{\sigma_1^2}+\frac{1}{\sigma_2^2}\right) - \frac{2x_2 x_3}{\sigma_1^2} + \frac{x_3^2}{\sigma_1^2}\right\}\right] \qquad \ldots\ldots(16)$$

Since y_1 and y_2 have the same r.m.s. values as x_1 and x_2 it follows that

$$p(y_2, y_3) = \frac{1}{2\pi\sigma_1\sigma_2} \exp\left[-\frac{1}{2}\left\{y_2^2\left(\frac{1}{\sigma_1^2}+\frac{1}{\sigma_2^2}\right) - \frac{2y_2 y_3}{\sigma_1^2} + \frac{y_3^2}{\sigma_1^2}\right\}\right] \qquad \ldots\ldots(17)$$

Combining equations (16) and (17)

$$p(x_2, x_3, y_2, y_3) = \frac{1}{2\pi^2_2\sigma_1^2\sigma_2^2} \exp\left[-\frac{1}{2}\left\{(x_2^2+y_2^2)\left(\frac{1}{\sigma_1^2}+\frac{1}{\sigma_2^2}\right) - \frac{2}{\sigma_1^2}(x_2 x_3 + y_2 y_3) + \frac{x_3^2+y_3^2}{\sigma_1^2}\right\}\right] \qquad (18)$$

The ratio of the envelope values of the echoes in the model is given by the ratio of R_1 to R_2. Therefore it is necessary to change to the use of polar co-ordinates R_1, θ_1 and R_2, θ_2 in the phasor representations.

This transformation of variables is defined by the equations

$$\begin{aligned} x_2 &= R_2\cos\theta_2 - A, \qquad & y_2 &= R_2\sin\theta_2 \\ x_3 &= R_1\cos\theta_1 - A, \qquad & y_3 &= R_1\sin\theta_1 \end{aligned} \qquad \ldots(19)$$

where A is the magnitude of the dominant part of the echo.

The joint p.d.f. for R_1, θ_1, R_2 and θ_2 can be found by the appropriate use of the Jacobian[3] for the transformation and the value of the Jacobian can be shown to be $R_1 R_2$ in this case.

Thus

$$p(R_1, R_2, \theta_1, \theta_2) = \frac{R_1 R_2}{4\pi^2\sigma_1^2\sigma_2^2}\exp\left[-\frac{1}{2}\left\{R_2^2\left(\frac{1}{\sigma_1^2}+\frac{1}{\sigma_2^2}\right)+\frac{A^2}{\sigma_2^2}+\right.\right.$$
$$\left.\left.+\frac{R_1^2}{\sigma_1^2}-\frac{2AR_2\cos\theta_2}{\sigma_2^2}-\frac{2R_1R_2\cos(\theta_1-\theta_2)}{\sigma_1^2}\right\}\right] \quad \text{......(20)}$$

The phase angles θ_1 and θ_2 are of no interest in this analysis and it is convenient to integrate them out of equation (20) to obtain an expression for the joint p.d.f. of R_1 and R_2 alone.

Hence

$$p(R_1 R_2) = \int_0^{2\pi}\int_0^{2\pi} p(R_1, R_2, \theta_1, \theta_2)\, \mathrm{d}\theta_1\, \mathrm{d}\theta_2$$

and it is not difficult to show that for $p(R_1, R_2, \theta_1, \theta_2)$ defined by equation (20) we have,

$$p(R_1 R_2) = \frac{R_1 R_2}{\sigma_1^2\sigma_2^2}\times$$
$$\times\exp-\left[\frac{1}{2}\left\{R_2^2\left(\frac{1}{\sigma_1^2}+\frac{1}{\sigma_2^2}\right)+\frac{A^2}{\sigma_2^2}+\frac{R_1^2}{\sigma_1^2}\right\}\right]\times$$
$$\times I_0\left(\frac{R_1R_2}{\sigma_1^2}\right)I_0\left(\frac{AR_2}{\sigma_2^2}\right) \quad (21)$$

10. Appendix 2: The p.d.f. for the Quotient R_1/R_2

Given $z = R_1/R_2$ the p.d.f. for this quotient can be shown to be given by

$$p(z) = \int_0^\infty R_2\, p(R_1, R_2)_{R_1 = zR_2}\, \mathrm{d}R_2 \quad \text{......(22)}$$

and by using equation (21)

$$p(z) = \frac{z\exp[-A_2^2/2\sigma_2^2]}{\sigma_1^2\sigma_2^2}\int_0^\infty R_2^3\times$$
$$\times\exp-\left[\frac{R_2^2}{2}\left(\frac{z^2+1}{\sigma_1^2}+\frac{1}{\sigma_2^2}\right)\right]\times$$
$$\times I_0\left(\frac{zR_2^2}{\sigma_1^2}\right)I_0\left(\frac{AR_2}{\sigma_2^2}\right)\mathrm{d}R_2 \quad (23)$$

Making the substitutions,

$$u = \frac{R_2}{\sigma_1\sigma_2\sqrt{2}}\{(z^2+1)\sigma_2^2+\sigma_1^2\}^{\frac{1}{2}}$$

$$p = \frac{z\sigma_2^2}{(z^2+1)\sigma_2^2+\sigma_1^2}$$

and

$$q = \frac{A\sigma_1}{\sigma_2\sqrt{2}\{(z^2+1)\sigma_2^2+\sigma_1^2\}^{\frac{1}{2}}} \quad \text{......(24)}$$

equation (23) can be written in the form

$$p(z) = \frac{4p^2\sigma_1^2}{z\sigma_2^2}\exp\left[-\frac{A^2}{2\sigma_2^2}\right]\times$$
$$\times\int_0^\infty u^3\exp[-u^2]\,I_0(2pu^2)I_0(2qu)\,\mathrm{d}u \quad (25)$$

In order to evaluate the integral of equation (25) the term $I_0(2pu^2)$ is expanded as a power series in u and the resulting series of integrals are then of a standard form.[4] Using this approach, equation (25) ultimately becomes,

$$p(z) = \frac{2p^2\sigma_1^2}{z\sigma_2^2}\exp\left[q^2-\frac{A^2}{2\sigma_2^2}\right]\times$$
$$\times\sum_{n=0}^{\infty}\frac{(2n+1)!\,p^{2n}}{n!\,n!}\,{}_1F_1(-2n-1, 1, -q^2) \quad (26)$$

For the purposes of computation using equation (26) it is convenient to define two new parameters k_1 and k_2 by setting $k_1 = \sigma_1/\sigma_2$ and $k_2 = A/\sigma_2$. In terms of k_1 and k_2 the parameters p and q become,

$$p = \frac{z}{z^2+1+k_1^2} \quad \text{and} \quad q = -\frac{k_1k_2}{\sqrt{2}(z^2+1+k_1^2)^{\frac{1}{2}}} \quad (27)$$

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Polarity Coincidence Techniques for Correlation Function Measurement and System Response Evaluation

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The correlation functions of many classes of signals can be derived quite simply from the correlation function computed from the waveform polarity (one-bit quantization) and in situations where one-bit quantization does not yield meaningful correlation functions, the addition of auxiliary signals can lead to a proper evaluation. The design of a flexible one-bit correlator is described using Micronor II (9 ns) circuitry which operates at a basic clock rate from 0–10 MHz and yields information on signals up to frequencies of 25 MHz. Performance details are given for the evaluation of auto-correlation functions and cross-correlation functions of various signals, and, in particular, the application to system response evaluation using Gaussian noise and binary pseudo-noise sequences.

1. Introduction

The powerful tool of signal-correlation has applications in almost every field of science where measurements may be reduced to an electrical signal. Correlation may be required where observations are contaminated by noise and other extraneous stimuli; where the spectral content of the signal may be required, or when a system impulse function is to be obtained by low level random noise stimulation.

Analogue correlation techniques are in general likely to remain the most efficient in terms of operating speed and possibly simplicity, but these are usually designed for special applications. Their versatility as a general correlation instrument is usually limited, and their dynamic range and stability cannot compete with digital correlation techniques.

Despite the advantages of digital correlators, they have not been prominent laboratory instruments, mainly because of the high cost of the digital delay store, multiplier units and accumulators. The hardware involved amounts to the equivalent of a small computer and indeed a small modern computer equipped with two fast digitizers can often be used as quite a versatile correlator. The computer cannot usually compete in speed with a specially designed correlator, however, because it has only one single processor; a full correlator will contain a multiplier unit for each channel, presenting simultaneously the correlation function at many values of the delay.

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Many of the complexities of correlator design are very significantly reduced if a relaxation is permitted on the quantization accuracy of the input signals. In the limit, the coarsest level of quantization corresponds to a polarity detector. The signal is observed to be either positive or negative, and may be assigned a logical value 1 or 0. This technique reduces the input digitizer to a comparator; the store containing delayed signal information reduces to a simple one-bit shift register, the multipliers reduce to modulo-2 adders and the accumulators are replaced by pulse counters. The information lost by rejecting this amplitude information is surprisingly small, and can usually be retrieved for random stationary processes by a slightly longer integration time. The correlation function computed from the one-bit correlator may be easily converted to the true correlation function by utilizing the appropriate relationship (Sect. 2). Alternatively, the mixing of appropriate signals with the input signals before quantization can lead to a direct output of the true correlation function from a one-bit correlator. With binary input data the true correlation function is obtained directly.

The advantages of polarity coincidence techniques (one-bit correlation) were clearly demonstrated in radio astronomy by Weinreb[1] where a search for the deuterium line was carried out using an integration time of 21 weeks. More recently a 256-channel one-bit correlator has been commissioned at Jodrell Bank for spectral observations.[2]

2. Relationships in Correlation

A full account of the general relationships in correlation may be obtained in textbooks[3,4] for example. Some pertinent relationships are summarized here and compared to the equivalent relationships resulting from a one-bit correlation technique.

2.1. *The Auto-correlation Function*

The auto-correlation function $R_{11}(\tau)$ of a waveform $f_1(t)$ may be expressed as

$$R_{11}(\tau) = \frac{1}{2T}\int_{-T}^{+T} f_1(t)f_1(t+\tau)\,\mathrm{d}t \qquad \ldots\ldots(1)$$

It may be normalized to

$$\rho_{11}(\tau) = \frac{1}{\sigma_1^2 2T}\int_{-T}^{+T} f_1(t)f_1(t+\tau)\,\mathrm{d}t \qquad \ldots\ldots(2)$$

For a periodic waveform the integration period $2T$ equals the period of the wave, and the correlation function is also repetitive. For a random stationary process, $T\rightarrow\infty$ and the function is not repetitive. In both cases the function is symmetrical with respect to $\tau = 0$ and $R(\mathrm{o}) \geqslant R(\tau)$ for all values of τ.

An important property of the auto-correlation function in the study of random processes is that the power frequency spectrum of the noise is equal to the Fourier transform of the auto-correlation function. A further property is that the auto-correlation function of a signal comprising the sum of a periodic signal and a random process yields the sum of the individual auto-correlation functions of the signal and of the noise.

The auto-correlation relationship when the waveform $f_1(t)$ is Gaussian noise and is quantized to one-bit accuracy $\rho'_{11}(\tau)$ has been related to the true auto-correlation function by Van Vleck:[5]

$$\rho_{11}(\tau) = \sin\left(\frac{\Pi}{2}\,\rho'_{11}(\tau)\right) \qquad \ldots\ldots(3)$$

This correction also applies to a pure sine wave, and McFadden[6] has shown that this holds approximately in the case of a weak sine wave plus Gaussian noise. McFadden's correction to equation (3) is of the second order in the signal/noise power level and may be neglected for small values of the signal to noise level.

2.2 *The Cross-correlation Function*

The normalized cross-correlation function of two waveforms $f_1(t)$ and $f_2(t)$ may be expressed as:

$$\rho_{12}(\tau) = \frac{1}{\sigma_1\sigma_2 2T}\int_{-T}^{+T} f_1(t)f_2(t+\tau)\,\mathrm{d}t \qquad \ldots\ldots(4)$$

Unlike the auto-correlation function this is not in general symmetrical with respect to $\tau = 0$, and the function may not have a maximum at $\tau = 0$. If $f_1(t)$ and $f_2(t)$ are random processes with a joint probability distribution, it has been shown that a one-bit cross-correlation may be performed[7,8,9] and that a correction similar to the Van Vleck correction for auto-correlation may be applied:

$$\rho_{12}(\tau) = \sin\left(\frac{\Pi}{2}\,\rho'_{12}(\tau)\right) \qquad \ldots\ldots(5)$$

If the functions $f_1(t)$ and $f_2(t)$ are random processes bounded by levels $\pm A_1$ and $\pm A_2$, Weinreb[1] following earlier publications[8,9] has shown that the addition of auxiliary signals of uniform probability distribution $\pm 1/2A_1$ and $\pm 1/2A_2$ respectively to the signals $f_1(t)$ and $f_2(t)$ leads to a one-bit correlation result identical in form to the true correlation function and modified only by a scaling factor:

$$R'_{12}(\tau) = R_{12}(\tau) = A_1A_2\rho'_{12}(\tau) \qquad \ldots\ldots(6)$$

and we have therefore

$$\rho'_{12}(\tau) = \frac{\sigma_1\sigma_2}{A_1A_2}\,\rho_{12}(\tau) \qquad \ldots\ldots(7)$$

The restriction placed on the signals that they be bounded by $\pm A$ may usually be approximated for Gaussian inputs by choosing a value of A several times the standard deviation, but it must be noted that the statistical uncertainty of the estimated correlation function increases with increasing values of the bounds. Alternatively random auxiliary signals of Gaussian probability distribution may sometimes be added to signals prior to one-bit correlation if one-bit quantization does not lead to a meaningful result, but in this case equation (5) would be used to derive the approximate correlation function.

2.3 *Correlation Functions in a Linear System*

If $f_i(t)$ and $f_o(t)$ represent respectively a random input and the output of a linear system, which may be characterized by a unit impulse response $h(v)$, it may be demonstrated[3] that the input–output correlation function $R_{io}(\tau)$ equals the convolution of the input auto-correlation function and the system impulse response:

$$R_{io}(\tau) = \int_{-\infty}^{+\infty} h(v)R_{ii}(\tau-v)\,\mathrm{d}v \qquad \ldots\ldots(8)$$

If at any point in the linear system a further disturbance $f_D(t)$ is present, the input–output correlation function is modified:

$$R_{io}(\tau) = \int_{-\infty}^{+\infty} h(v)R_{ii}(\tau-v)\,\mathrm{d}v + \overline{f_i(t)}\cdot\overline{f_D(t)}\cdot\int_{-\infty}^{+\infty} g(v)\,\mathrm{d}v \qquad \ldots\ldots(9)$$

where $g(v)$ represents the impulse response to the disturbance $f_D(t)$. If $f_D(t)$ is introduced at the same point as $f_i(t)$ then $h(v) = g(v)$. The last term of equation (9) is usually constant, but also, if either $\overline{f_i(t)}$ or $\overline{f_D(t)} = 0$, the term is zero, and the disturbance will not affect the input–output correlation relationship except where it may increase the statistical fluctuation in the estimate over a limited observation time.

The input–output correlation function computed from a one-bit quantization may be related to the true correlation function by equation (5) or (7) and the unit impulse response derived from equation (9). This procedure may often be very much simplified in practice. For example, if $f_i(t)$ is a random Gaussian function with a spectral content extending to frequencies much higher than the passband of the system, the input–output one-bit correlation function procedure will then yield the impulse response by a simple correction:

$$h(v) = \sin\left[\frac{\Pi}{2}\,\rho'_{io}(\tau)\right] \qquad ......(10)$$

If appropriate signals are added to the input and output signals before quantization, then the impulse response results directly from the one-bit correlator with recourse only to a constant scaling factor given by equation (7).

The output auto-correlation function may also be of interest since its Fourier transform is the output power frequency spectrum. The output auto-correlation $\rho_{oo}(\tau)$ function is shown[3] to be the convolution of the input auto-correlation function and the auto-correlation function of the system impulse response. The one-bit correlation estimate of this function will again relate to the true correlation function by equation (3).

3. Design Aspects for a Polarity Coincidence Correlator

3.1 *General Features*

The basic features of a digital one-bit correlator are shown in Fig. 1. Since cross-correlation functions are not necessarily symmetrical with respect to $\tau = 0$, the delay will also have to be introduced to both signals in order to evaluate the complete function. This may be achieved either by reversing the inputs $f_1(t)$ and $f_2(t)$, or by using two shift registers (Fig. 1(b)). This latter system was not incorporated in the system constructed, but has much to recommend it, particularly for visual display outputs where the simultaneous display for $\pm\tau$ is desirable.

The maximum frequency of components of the input data which are sampled at a frequency f_s by the correlator must be limited to $\leqslant f_s/2$ if the values of the auto-correlation function are to be interpolated to yield the true correlation function. The finite sampling time t_s restricts the correlator frequency resolution at frequencies $\gtrsim 1/2t_s$, but in general the sampling time is shorter than the sampling period and thus this restriction does not obviate the need for filtering prior to sampling. The restriction on the input frequency spectrum may be lifted, however, if it is acceptable to sample the input data at periodic intervals less than that required for the complete recovery of all the information content. There are two techniques convenient for this operation which extend the performance of a basic correlator:

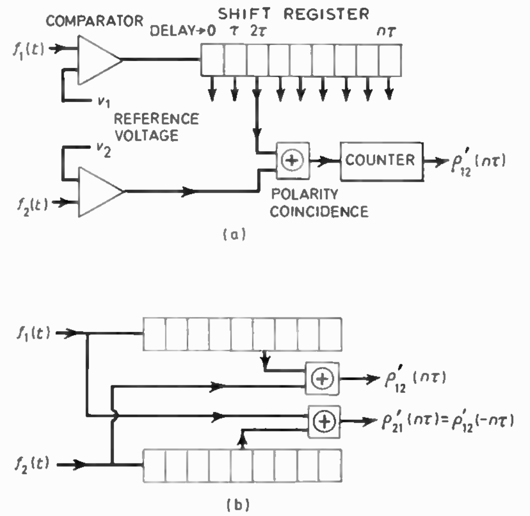


Fig. 1. (a) The basic features of a one-bit digital correlator. In a simple system, a single counter may be used to display the correlation function $\rho_{21}(n\tau)$ at a particular delay $n\tau$. Negative arguments of the function may be obtained by interchanging the inputs $f_1(t)$ and $f_2(t)$ as in (b) where a delay is incorporated in each channel.

(i) Variation of the sampling frequency

Figure 2 shows an arbitrary correlation function for a signal containing frequency components $>f_s/2$. It may be seen that the results of the correlation represent points on the correlation function, but the points are not sufficient to evaluate the true correlation function. By decreasing the sampling frequency the whole range of delay $>1/f_s$ may be scanned but no information is gained in the range $\tau = 0$ to $1/f_s$. Figure 2 shows the effect of a 20% decrease in sampling frequency.

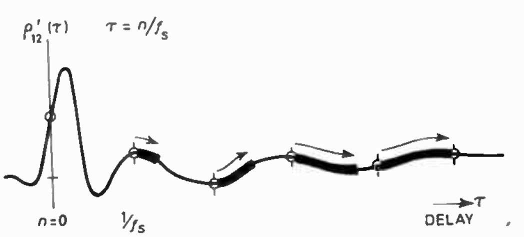


Fig. 2. An arbitrary correlation function and typical correlation results obtained when the waveform is 'under sampled'. A variation of the sampling frequency f_s enables the points to be spread out to cover the range of delay $\geqslant 1/f_s$. The effect of a 20% reduction in sampling frequency is illustrated.

(ii) Sampling phase variation

If the inputs are sampled at the same instant of time, $\rho'_{12}(n)\tau$ will be obtained for $n = 0, 1, 2$, (Fig. 3). If a delay δt is now introduced between the two instants at which the data are sampled, the correlation function $\rho'_{12}(n\tau+\delta)$ will be obtained. By varying δ over the range $\pm 1/f_s$ the entire correlation function may be explored. This is illustrated in Fig. 3. The maximum frequency limit of the input data is now determined only by the sampling width of $\sim$20 ns, corresponding to a potential performance on input frequencies of up to 25 MHz, but sampling diodes could well achieve a performance of up to two orders of magnitude greater than this frequency.

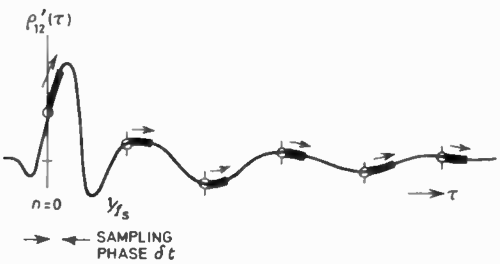


Fig. 3. The effect of a sampling phase variation δt on the measured correlation function is illustrated. The entire correlation function may now be explored by varying δt from 0 to $1/f_s$.

3.2 *System Construction*

The input comparators are Fairchild μA 710 units which drive a TTL 9009 power gate. The comparator reference setting is obtained without undue regard for high stability from Zener reference voltages and a linear potential divider, which enables any comparison voltages within the range $\pm$4 V to be selected, with a stability of 10 mV and a threshold sensitivity of 2 mV. Digital inputs may be presented directly to the sampling pulses. Delays are generated by coaxial delay lines so that sampling pulse widths and sampling phases may be obtained accurately over operating frequencies from 0 to 10 MHz. Ferranti Micronor II DTL Series E circuitry is used in all applications except in the comparator. A 'speeding up' of waveforms throughout the system is achieved by optimum resistor loading. The schematic layout of the system is shown in Fig. 4.

4. Polarity Coincidence Correlator Performance

4.1 *Probability Distribution*

If one channel of the correlator is set to a constant logical level (say, 1) and the comparator of the other channel set at x volts, then a count will be registered

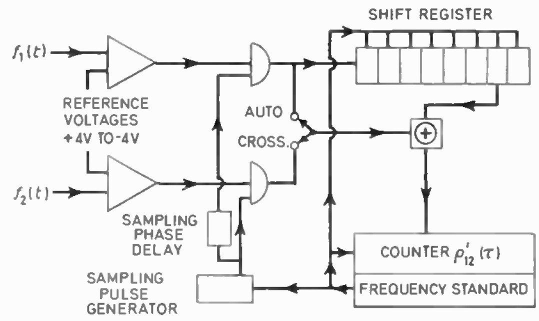


Fig. 4. Schematic layout of the constructed one-bit 0 to 10 MHz correlator.

when a sampling pulse detects the input to be greater than x volts. Providing that sampling intervals are uncorrelated with the input waveform, the cumulative distribution may be obtained from the output of any channel of the correlator. This restriction is fully met for random inputs and is not usually a serious limitation for periodic inputs providing the sampling frequency is not phase-locked to the input waveform and that the integration time is long compared to the period of the waveform. Experimental results of the cumulative noise distribution measured by this technique are shown in Fig. 5 for the output of a 5 MHz noise generator. Over an integration period of one second, the cumulative probability function $q(x)$ in the region of $q(x) \sim 0{\cdot}5$ may be measured to an accuracy of 0·04% at a sampling frequency of 10 MHz. The differential probability density distribution $p(x)$ may now be simply derived from the slope of this curve, and from this distribution the characteristic function and the moments of the distribution may be

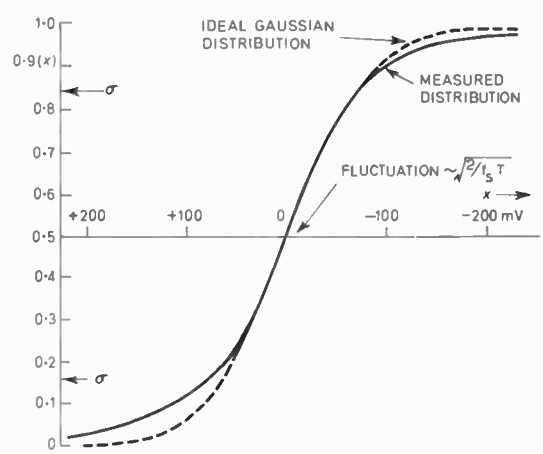


Fig. 5. Measured values of the probability distribution of a 5 MHz noise generator and the ideal Gaussian distribution. The mean and variance may be derived from this measurement. Statistical fluctuations near $q(x) \simeq 0{\cdot}5$ amount to less than 0·04% for an integration time of 1 s.

calculated. Of the various moments, the mean and variance are more frequently required. If the cumulative noise distribution is symmetrical about the value $q(x) = 0{\cdot}5$ then the mean may be read directly from the value of the ordinate at this point on the graph; if the noise is approximately Gaussian then the variance may also be calculated from the ordinate values at $q(x) = 0{\cdot}159$ and $0{\cdot}841$. The probability distribution provides useful information on noise sources, and the simplicity and accuracy with which measurements may be made by this technique suggests that it could compete very well with more conventional techniques of noise power measurement.

4.2 *Auto-correlation Function*

Measurements were made of the one-bit auto-correlation function of Gaussian noise from a 5 MHz Gaussian noise generator and of the same noise to which a sine wave of power $0{\cdot}07\sigma_N^2$ was added. The values were corrected using the Van Vleck formula and are displayed in Fig. 6. The difference in the abscissae of the corrected auto-correlation functions of the signal plus noise and of the signal yields the auto-correlation function of the signal, but reduced by the factor $[1+(P_N/P_S)]$ where P_N and P_S represent respectively the average noise and signal powers. For small departures of the one-bit correlation function from zero, the one-bit correction formula

$$\rho_{12}(\tau) = \sin\left[\tfrac{1}{2}\pi\rho'_{12}(\tau)\right]$$

may be approximated to

$$\rho_{12}(\tau) = \tfrac{1}{2}\pi\rho'_{12}(\tau).$$

In the case of a very weak periodic signal and noise, then, the amplitude of the measured one-bit correlation function for large values of τ approximates to

$$\rho'_{11}(\tau) = \tfrac{1}{2}\pi(P_S/P_N)\rho_{ss}(\tau).$$

The one-bit auto-correlation estimate thus yields a scaled version of the true signal auto-correlation function $\rho_{ss}(\tau)$ for values of τ greater than the noise correlation time.

Auto-correlation functions of a three-bit pseudo-noise sequence and a sine wave are shown respectively in Figs. 7 and 8. The results from the binary input yield the true correlation function without recourse to correction. The effect of spurious noise in a sine wave signal near the comparator threshold sensitivity is shown in Fig. 8.

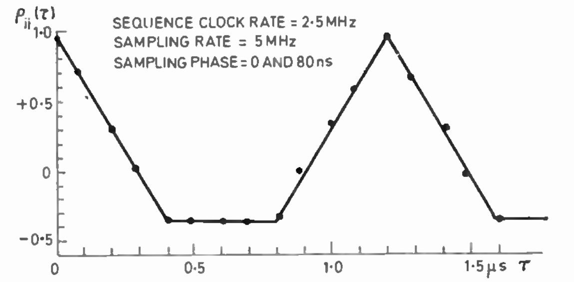
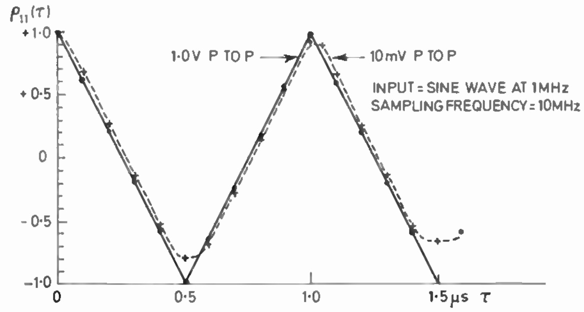
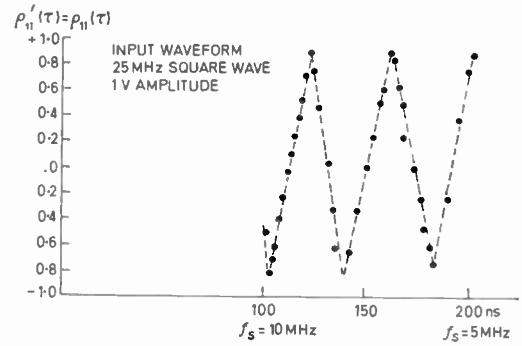


Fig. 7. Measured values of the auto-correlation function of a three-bit pseudo-noise sequence (110).



(a) Measured value of the auto-correlation function of a 1 MHz sine wave sampled at 10 MHz, for signal amplitudes 10 mV and 1 V. The effect of a noisy environment on the measured auto-correlation function of the smaller signal is illustrated.



(b) Partial auto-correlation function of 1 V square wave at 25 MHz obtained by sampling frequency variation.

Fig. 8.

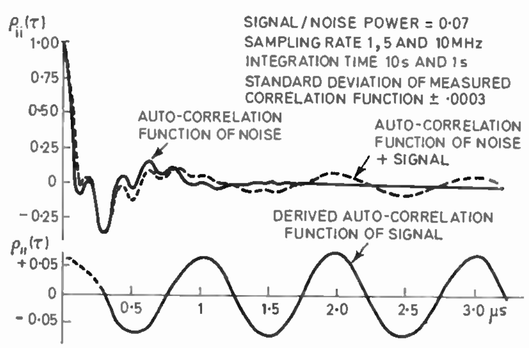


Fig. 6. Corrected values of the measured one-bit auto-correlation functions of a signal and a signal plus Gaussian noise. The derived signal auto-correlation function is also shown.

4.3 *System Impulse Response by One-bit Correlation*

4.3.1. Random noise stimulation

Gaussian noise from a 0–5 MHz noise generator was used to stimulate a simple RC test circuit. The auto-correlation functions of the input $\rho'_{ii}(\tau)$, the output $\rho'_{oo}(\tau)$ and the cross-correlation function $\rho'_{io}(\tau)$ are shown in Fig. 9. The form is that expected by the convolution relationship in equation (8). The severe disadvantages of this form of input stimulation are highlighted by the comparison of this result with the ideal result which would be obtained by white-noise stimulation. The extended nature of the noise auto-correlation function poses an awkward deconvolution problem. An ideal Gaussian roll-off of the power frequency spectrum of the noise would lead to a more convenient auto-correlation function, but the use of very accurately filtered noise sources would seem to be obviated by the use of more conveniently generated pseudo-noise binary sequences (Sect. 4.3.2).

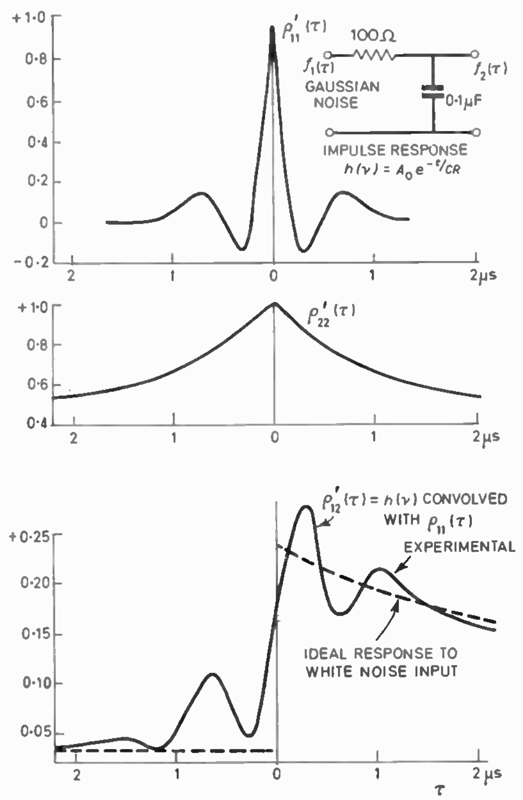


Fig. 9. One-bit correlation functions measured for an RC circuit under Gaussian noise stimulation, showing the effects of the extended auto-correlation function of the noise upon the measured input-output correlation.

4.3.2. Pseudo random noise stimulation

The output of a low-pass filter stimulated by a short pseudo-noise sequence of high repetition frequency will show a fundamental frequency equal to the sequence repetition frequency. Although higher frequency components will be present at the output, the one-bit quantization of the signal will largely mask the presence of these, and since this process is deterministic, the phase of these components will never vary relative to the repetition frequency. One-bit cross-correlation of the input–output waveforms cannot be expected therefore to yield a meaningful result in this situation. The result of a one-bit input–output correlation of an RC integration circuit stimulated by a relatively short (127-bit) sequence is shown in Fig. 10(a).

Following the variable clipping level technique suggested in Section 2.2, the clipping level at the output of the RC filter was varied by adding an auxiliary triangular wave of period equal to the integration period. The triangular waveform satisfies the requirement that the probability density be uniform. If the

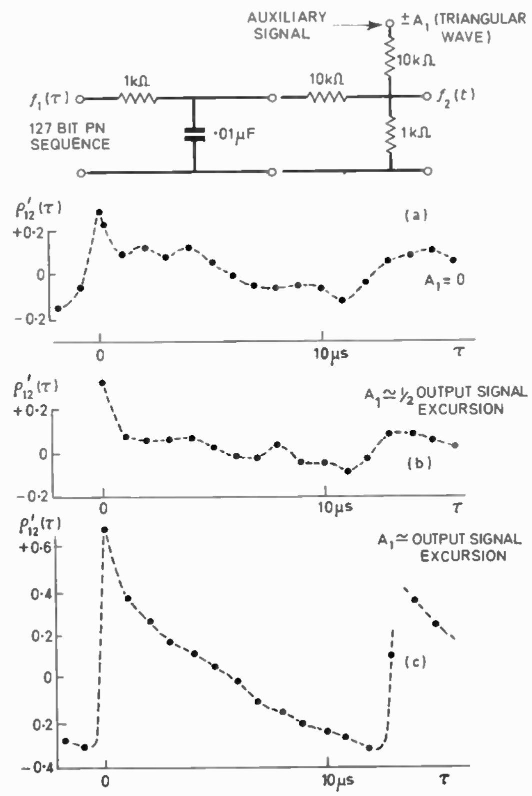


Fig. 10. One-bit correlator functions measured for an RC circuit stimulated by a 127-bit pseudo-noise sequence. The effect of auxiliary signals on the output of the RC circuit leads to an approximate representation of the response to a train of impulses (c).

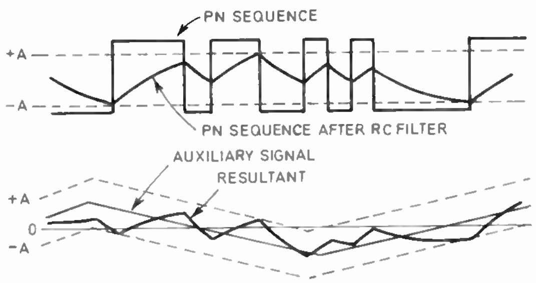


Fig. 11. A typical output of an RC filter stimulated by a binary sequence and the effect of the addition of an auxiliary signal of probability density $\pm 1/2A$.

system output is limited to $\pm A_1$, an auxiliary signal of amplitude A_1 ensures that all amplitudes of the waveform are 'swept past' the clipping level. The effect on the waveform is shown in Fig. 11, and the results of the subsequent one-bit correlation are shown in Figs. 10(b) and (c). It may be noted that a most useful feature of the use of (binary) random signals for the input stimulation of this technique is that the input waveform need not be disturbed by an auxiliary signal to yield the true correlation function, since this procedure does not lead to the presentation of extra information to the correlator input; complete information on the signal is already obtained by a one-bit quantization. The auxiliary signal should have a period equal to the integration period or an exact multiple of this, in order to ensure a mean value of zero. A fast auxiliary signal (dither) may also be used[10] or even a random dither satisfying the requirement of a uniform probability distribution. Stassen[11] has shown that the application of dither to both inputs of a polarity coincidence correlator yields the true correlation function for random inputs of arbitrary probability density.

The amplitude of the measured correlation function is inversely proportional to the auxiliary signal amplitude A when only one clipping level is varied. Although it has been demonstrated that the results obtained with an auxiliary signal very much improve the one-bit correlation technique, the use of short sequences leads only to an approximation of the true impulse function to the extent that they may be considered approximations to random functions. The system response correlation functions measured with stimulation from pseudo-noise sequence of length 8 388 607 bits (generated from a 23-stage shift register) is shown for an RC circuit in Fig. 12 and for an LCR circuit in Fig. 13, both without the application of an auxiliary signal. The need for an auxiliary signal is obviated with a long sequence, but its use might be invoked to avoid the use of correction factors to the results of a one-bit correlation.

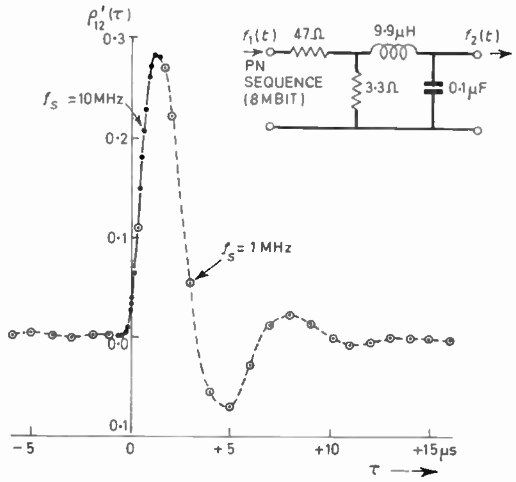


Fig. 13. One-bit correlation functions measured for an LCR circuit stimulated by a pseudo-noise sequence of length 8 388 607 bits (without auxiliary signals).

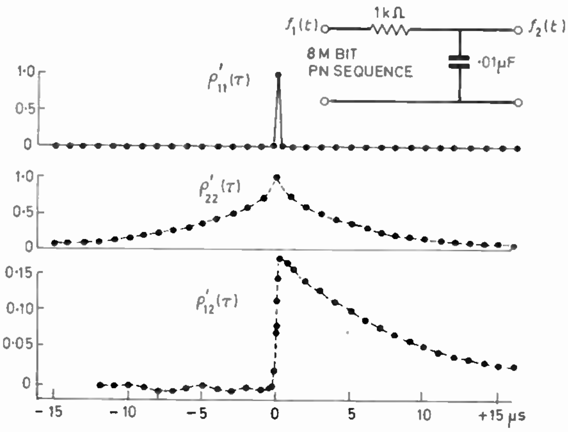


Fig. 12. One-bit correlation functions measured for an RC circuit stimulated by a pseudo-noise sequence of length 8 388 607 bits (without auxiliary signal).

The use of digitally generated noise also offers advantages in system response measurements.[10] The frequency spectrum may be controlled over a very wide range without sophisticated filtering techniques, and it is also fully reproducible. The well-behaved form of the auto-correlation function is perhaps the most important property of these sequences, however, for system response evaluation, where the effects on the cross correlation function are readily predictable.

The impulse response measurements shown in Figs. 12 and 13 using long sequences may be measured to a great precision—typically 0·1% with an integration period of 1 s. Similar results would be expected by low level stimulation of an operational system over a longer integration time.

5. Conclusion

(i) A polarity coincidence correlator with a dynamic range of 0–10 MHz has been constructed which operates with full capability on signals up to 5 MHz. A modification of the sampling system which leads to an extension of its useful range to ~25 MHz has been demonstrated.

(ii) The correlator has been used to measure white noise probability distribution to a very high accuracy and also correlation functions of various combinations of signals and noise.

(iii) Polarity coincidence techniques for system response evaluation have been investigated by computing the system input–output correlation under stimulation by Gaussian noise and pseudo-noise sequences of two lengths. The use of band-limited Gaussian noise is shown to lead to difficulty in the interpretation of the results; using short pseudo-noise stimulation, the technique has been shown to lead to approximate correlation function measurement and hence the impulse response function, provided that the clipping level of the output waveform may be varied throughout the integration period. With long sequences very accurate determinations of the correlation function have been obtained without varying the clipping level. This feature might usefully be included in one-bit correlation measurements, however, to avoid the use of correction factors.

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Quality Control in Capacitor Production and Testing

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The properties of materials and the appropriate dimensions must be controlled so that the requirements of the electrical specification are met from the beginning of the life cycle of the product. The influence of design on control of manufacture is discussed and the factors which contribute to variability in production are examined. The main quality control function is to establish a 'right first time' philosophy. While the concept of yield is inevitable in component manufacture, low yield due to lack of control is unacceptable. In-process inspection with feedback for corrective action not only maintains control but also points the way for continuous improvement of design, methods, quality and reliability. In some cases the involvement of the production operator can be obtained by allowing him to carry out this checking.

Acceptance and quality assurance testing are of primary importance in demonstrating to the customer that the requirements of his specification have been met. Failures in this testing are warnings of lack of control in design, materials, or manufacture. The BS 9000 scheme is welcomed as establishing a systematic series of such tests.

From the internal point of view total quality costs are the best method of measuring the performance of the entire manufacturing organization. If the quality control function is effective, and assuming no radical changes of product mix, quality costs should be falling. From the external point of view an effective quality control organization is the best assurance of consistent performance for the future.

1. Introduction

A general description of the construction, application, reliability and mechanisms of failure of capacitors has been given in which it is concluded that 'The main need is for regular and systematic testing of components and the feedback of fault data to improve the control of manufacturing processes. Reliable components can only be provided by fully optimized manufacturing techniques'.[1]

This present paper discusses the organization of a production unit from the quality point of view and describes the methods employed in the control of the manufacture of metal-cased solid tantalum capacitors. Similar techniques have been applied to the other product lines such as mica, resin-dipped solid tantalum and metallized plastic film capacitors.

2. Quality and Reliability

Unfortunately quality control involves the use of a great deal of jargon, the meaning of which is by no means consistent. It is therefore necessary to define carefully the meaning of the terms to be used and their interrelation. The relationship between these terms is shown diagrammatically in Fig. 1, and some definitions are given in the Appendix.

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Quality of design is determined by the application class, according to BS 9000 as follows:

Submerged repeater Satellite Airborne	Special purpose class
Telecommunication Computer	General application class
Industrial Television and radio	Commercial application class

The performance specification defines the requirements for a particular component in terms of environmental conditions, limits of electrical characteristics, and physical attributes required for this application class. It will also define the tests to be carried out for qualification approval and for acceptance or quality assurance purposes. In the specifications with which this paper is concerned reliability is not specified quantitatively but requirements are implied by a knowledge of the application and covered as far as economically possible by the test schedules. These requirements will also determine the manufacturing conditions and degree of control of manufacture necessary to ensure that they are met.

Quality Control is primarily concerned with quality of conformance. In so far as the requirements of the performance specification cover reliability, quality

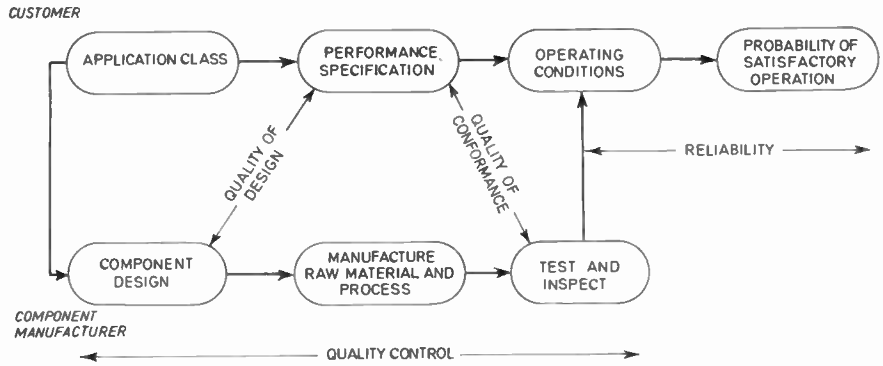


Fig. 1. Relationship between quality and reliability.

will ensure reliability. There are, furthermore, many faults which may not be detected in tests on a finished product but which may be maintained within specified limits, in process, by an effective quality control system. When, however, reliability is considered quantitatively it must be recognized that the two terms are dissimilar unless there is a complete knowledge of failure mechanisms which can be eliminated by non-destructive test methods. For the purposes of this paper we will assume that the object of quality control is to control both and that an improvement in one will normally give an improvement in the other.

There are a number of features of capacitor design which have a direct bearing on quality and reliability and which must be maintained in a state of control if satisfactory performance is to be maintained. These are discussed in the following sections.

2.1. *Capacitor Processes*

Why should quality control in capacitor production differ from that of any other product? The answer lies only in the physical nature of the product, the diversity of techniques and technologies employed, and the fact that it is made on a scale varying from small to large batch production. In general a capacitor consists of an area of the thinnest possible dielectric material commensurate with the working voltage and temperature. Basic dielectric materials are never perfect and in addition to microscopic defects such as impurities there are the macroscopic imperfections such as pinholes, inclusions, cleavages, weakspots, etc.

Raw material test methods have been evolved so as to maintain such materials within acceptable limits. The manufacturing techniques have been adapted to tolerate this background level of defects and to eliminate those capacitors which are likely to be early failures by voltage proof testing, ageing or other similar procedures. In addition, satisfactory performance depends on the control of processes such as winding, stacking, vacuum impregnation, anodization, the making of connexions etc.

2.2. *Encapsulation*

The essential physical attributes of a capacitor are provision of moisture and mechanical protection, terminations which are robust and solderable, and coding which is legible. While theoretically impervious to moisture the design of capacitor employing a metal case and glass or ceramic seals is subject to the same problems and disciplines of solderability as of other soldered assemblies. Indifferent solderability can result in reduced manufacturing rates, leaks and uneven flow of solder both inside and outside the assembly. Eliminating such defective work requires visual examination, leak tests, X-ray checks, both on a sampling basis and, if rejects are found, on a 100% basis. Although some progress has been made as scales of production have increased, these manufacturing operations still depart a long way from a 'right first time' philosophy.

Paradoxically capacitors employing resin or rubber seals are more controllable and have been used for many years. Modern production techniques using a wide variety of resin systems are now available giving high production rates and good moisture protection. Many of these devices will meet the requirements of 21 or even 56 days long-term damp heat.

2.3. *Terminations*

The solderability of terminations has been the cause of many problems, particularly in printed circuit board assembly. The advent of the solder globule test[2] gave for the first time a quantitative measurement of solderability. Wire which meets the requirements of the test after ageing has been shown by experience to solve this problem.

This test is still the subject of some controversy principally on the grounds that it sets too high a standard which can only be met at increased cost. However, there is now available a wide range of components which meet this requirement and as suitable materials become available it is to be hoped that it will gain general acceptance.

2.4. *Connexions*

It is an unfortunate fact that in the past many designs of capacitor have operated successfully employing pressure contacts. This applies particularly to the tape to foil junction in some wound capacitors. This is unfortunate because it has given rise to the belief that they are 'good enough'. Nevertheless, in the author's experience many major problems, some of them of an epidemic nature, are due to the use of this type of connexion. It is now considered that a reliable design must employ some form of fusion between all metallic circuit elements. Soldered or welded connexions meet this requirement as also do high temperature fired silver or other pastes. Conducting cements have given very variable results.

3. Quality Control Scheme

The quality control scheme described in this paper has been accepted as meeting the requirements of DEF 5001-C.[3] The inspection and control of the manufacturing process procedures have been accepted as meeting the requirements of BS 9000,[4,5] and MIL-STD-790B.[6] Where the latter specifications require specific test methods or quality assurance procedures these are, of course, employed instead of the general one described. It is envisaged that, as BS 9000 becomes more widely accepted, the scheme will be adapted where necessary to the requirement of that specification.

3.1. *Quality Policy and Organization*

In a division manufacturing several different products having widely differing requirements as regards performance, environment and reliability, it is difficult to state a single quality policy. The best statement we have found so far is:

'It is the policy of this Division to maintain a full awareness of requirements in the fields of application of our products and to ensure that they are efficiently and economically manufactured to a high standard of conformance with applicable specifications. Furthermore by continuous improvement of materials, methods, processes and techniques to ensure that such products and specifications are fully compatible with world products.'

The most important feature of such a statement is that it should be made by the General Manager and implemented by him in his day-to-day actions. Furthermore the same objective must be accepted by each of the functional managers—Chief Engineer, Production Manager, Chief Production Engineer, Quality Manager, Sales Manager.

Provided that they co-operate and recognize the interaction of their functions in achieving this objective there should be no problems that cannot be overcome and quality and reliability as well as quality costs should show continuous improvement.

The organization of manufacturing units from the quality point of view is many and various. In general they represent the evolutionary state of their quality policy. This is shown in Table 1, which represents the stages through which some of the product lines of this division have passed. The key steps are as follows:

(i) Appointment of Quality Manager;
(ii) Control charts on manufacturing operations;
(iii) 100% automatic testing introduced into production followed by acceptance sampling by Chief Inspector;

Table 1. Evolution of quality control organization

Stage	In process	Final	Carried out by	Responsible to
1	None	100% Manual	Chief Inspector	Production Manager or Chief Engineer
2	Control Charts	100% Manual	Chief Inspector	Quality Manager
3	—	100% Automatic	Production Shop	Production Manager
	Control Charts	Acceptance Sampling	Chief Inspector	Quality Manager
4	Control Charts	100% Automatic	Production Shop	Production Manager
	Quality Audit	Acceptance Sampling	Chief Inspector	Quality Manager

(iv) Hand over responsibility for detailed operation of control charts to production followed by quality audit by Chief Inspector.

Each of these changes had its own advantages. The appointment of Quality Manager permits the establishment of an overall quality policy based on a realistic assessment of quality costs. This includes the co-ordination of design and development, purchasing and production both on new and existing products. Stage (iv) has only just commenced, as described later.

It must be the design and production engineers' first responsibility to ensure that drawings and specifications are available and up to date at all times. These drawings and specifications must be mandatory. New designs must be subject to pilot production runs to bring to light production problems and to establish yield and manufacturing cost. Samples from this run should be subjected to quality assurance tests to prove the requirement specification.

Strict observance of manufacturing specifications on the shop floor is essential. The principle must be that if an engineering requirement cannot be met the specification must be amended either temporarily by a concession, but preferably permanently. If a production shop accepts a specification in the spirit 'we will do the best we can' it is almost certain that eventually the specification will be valueless and when trouble occurs there will be no firm ground on which to get back. For this reason improvisation and self help on the shop floor, however laudable and helpful they may appear, are not the right solutions to any problem. In stage (iv) it is envisaged that greater participation on the part of the production operator will be achieved by allowing him to maintain his own control charts. In this case the quality function will be able to carry out a 'quality audit' which consists of checking that these charts are being correctly maintained. Even more important, it can also consist of a detailed check that manufacturing specifications are being adhered to.

The primary responsibility for quality must rest with the Production Department. The quality function is essentially one of collection of data analysis and feedback for corrective action, and no action should be taken which weakens the responsibility of the production foreman. Nor must he be allowed to abdicate his responsibility to 'quality control'.

In the same way the setting of visual standards must be done jointly with the other functions and not unilaterally.

The quality function may be subdivided as follows:

(a) Raw material control—Supplier quality assurance and acceptance sampling of incoming raw materials and parts.

(b) Process control—Control of quality on the shop floor at each operation during manufacture, involving examination, feedback of information and corrective action when necessary.

(c) Screening tests—100% testing with rejection of all items found defective.

(d) Acceptance sampling—Sampling carried out to distinguish acceptable batches from those that are unacceptable.

(e) Quality assurance testing—Destructive testing carried out at regular intervals on samples taken at random from production to assess conformity with specification.

3.2. *Manufacturing Conditions*

In most modern factories a good standard of housekeeping is considered essential as part of the normal routine. Good discipline in such matters as cleaning of benches, routine cleaning of floors, etc., is quite adequate for many applications.

There are, however, some operations where a higher standard of cleanliness is required to prevent contamination or damage to the dielectric. This applies particularly to the winding of thin plastic films and also to the processing stages of the manufacture of electrolytic capacitors. It becomes even more important in the manufacture of capacitors for very high reliability applications such as submerged repeaters, where it is necessary to assure the reliability of each individual capacitor.

Clean manufacturing conditions can take a variety of forms:

(i) The use of enclosed machines fitted with clean (and, if necessary, dry) air.

(ii) The use of laminar flow benches. These have the great advantage of providing clean conditions in the immediate area of operation.

(iii) The establishment of enclosed areas fitted with filtered air (typically 5 μm). Access to these areas is restricted and operators wear protective clothing, such as white overalls.

(iv) The use of super-clean areas with air filtered to a very much higher standard (electrostatic precipitators). The air may be temperature and humidity controlled and access to the area restricted by pass only. Operators wear fully protective clothing, including hats and shoes.

Laminar flow benches have been found to give the same standards as the super-clean area over a limited space. For most normal capacitor manufacture using sensitive materials the clean area is adequate.

The psychological importance of the establishment of the correct manufacturing conditions must not be under-estimated. The operators take great pride in feeling that they are contributing to a product where such care is necessary.

Contamination from the operators themselves can be a cause of difficulty requiring a careful balance between technical and human requirements. Smoking is prohibited in all manufacturing areas. Eating and drinking is permitted only in the tea stations and operators are requested to wash their hands after each break. It is, however, not possible to prevent the eating of sweets nor, except for the very critical operations, the use of hand creams. Where contamination from the hands is critical, rubber fingerstalls are used and these are subject to specified washing and replacement procedures. We have not succeeded in finding headgear acceptable to the feminine taste which completely contains the hair so as to prevent dandruff contamination.

3.3. *The 'State of the Art' Process*

The manufacture of components depends on the quality of the raw materials and on control of processes. Many capacitors are of relatively new design and in the early part of their life cycle. The newer materials, such as tantalum products and the plastic films, are being improved and extended in scope with improvements of their technology. Capacitors have traditionally been made on a batch basis and many manufacturing operations are still dependent on certain operator skills. Methods improvement and mechanization are, however, gradually eliminating operator dependence coincident with an increase in the scale of manufacture.

In the early part of the life cycle it is likely that the manufacturing process itself may be 'state of the art'. That is to say that, while the basic elements will have been firmly established, the optimum conditions for any one manufacturing operation and the factors affecting yield may not have been fully evaluated. These must be identified under manufacturing conditions and using manufacturing equipment. Here we come to the dilemma of whether to insist that no changes be made or whether to accept that they may be introduced provided that control is maintained at all times.

Before we answer this we should examine conditions in a manufacturing shop. Turnover of labour is typically about 40% for women and 20% for men so that at any one time there is a proportion of trainee labour. Sickness, holidays and changes of volume of production make it essential that movement of operators from one job to another within the factory takes place. Productivity deals, cost reduction programmes, methods improvement and mechanization projects all require changes in methods of working. We must, therefore, state that a manufacturing shop is at best only in a state of dynamic equilibrium and this only by virtue of an effective quality control scheme.

Our answer on the question of change is, therefore, to accept it and to guide it firmly and surely in the direction of improved yields, reduced costs, improved quality and reliability. Furthermore we are convinced that these terms are compatible with each other.

Reference has been made several times to yield. This may be defined as

$$\text{yield} = \frac{\text{output}}{\text{throughput}} \times 100\%$$

The complement of this is called the reject rate. The term yield is a tacit acceptance of the fact that not all the throughput will meet acceptance specification requirements. At best this is a recognition of the fact that process capability is not equal to requirement. At worst it may be a statement that part of the manufacturing process is not under control. In the latter case quality and reliability are likely to be closely related. Yield improvement programmes are, therefore, a vital part of the activity of a production unit.

It is, of course, usual in specifications to require that significant changes may be made only with the approval of the appropriate authority. Our interpretation of this clause is that a significant change is one which effects interchangeability or which results in a significant change of an electrical parameter.

3.4. *Quality Improvement Teams*

Yield improvement activities, particularly in a Division with a number of different products, are organized through Quality Improvement Teams.

The key members of these teams are:

Design engineer
Production foreman
Quality control foreman
Production engineer.

The function of these teams must be as follows:

to optimize the communication between the departments;

to produce the right product, of required quality, at the right cost, on time;

to co-ordinate the function of its members to achieve overall product line objectives;

to set targets for the reduction of rejection rates, prepare action programmes and monitor performance against these targets.

The individual members are responsible, on behalf of their department, for action decided upon at the meetings. They are responsible for communicating relevant information within their department, in particular to their supervisor.

The team chairman reports at the monthly Divisional Executive Meeting on key topics and seeks advice or decisions on problems beyond the scope of the team.

3.5. *Quality Costs*

The best assessment of the performance from the quality point of view of a manufacturing organization is to determine its quality costs. These are itemized as follows:

PREVENTION:

Pre-production quality evaluation
Vendor qualification approval
Quality planning
Design and development of quality measurement and control equipment
Quality training.

APPRAISAL:

Incoming test and inspection
In-process and final test and inspection
Customer approval and acceptance
Maintenance and calibration of test and equipment
Test and inspection equipment, tools and/or gauges (depreciation)
Quality audits.

FAILURE COSTS:

Production scrap
Re-work
Complaints—warranty costs.

If these expenses are correctly analysed they give the total cost to the division of not making the product 'right first time'. In general, in the component field, the higher the failure costs the greater the appraisal costs. The objective of any quality improvement programme must be to reduce failure costs and with them the appraisal costs. This may require an increase in expenditure on prevention activities. It will be noted that all inspection and all quality assurance testing fall within the definition of appraisal costs.

3.6. *Management Reports*

If management is to be fully involved it is necessary to provide them with digested information preferably on an exceptions basis. The following reports are prepared:

(a) Quality cost report showing the amount of the quality costs referred to in Section 3.5 for the month and year to date.
(b) Weekly output report. This is a summary prepared by computer of the data obtained from production batch tickets, showing quantity made, quantity passed, quantity rejected, analysis of rejections into ten classifications, percentage rejection.
(c) Quality assurance attributes analysis. A monthly report is issued showing the parametric and catastrophic failures for each product line and for each type of test carried out. In addition to this a 13-weeks cumulative summary is maintained for comparison with the requirements of DEF 5001-C (normally 4% level S3). If the acceptance number is exceeded lot-by-lot sampling is reinstated.
(d) Customers' return reports. A report of all customers' returns stating the reason for rejection and the number of capacitors involved. The customers' returns investigation procedure includes the completion of a pro-forma which ensures that the customer is advised of the results of all investigations.
(e) Monthly report. A monthly report is prepared for each product line showing the top five quality problems on each line ranked in order of magnitude and stating the action being taken. These quality problems can be at any stage of manufacture from incoming goods to quality assurance test.

3.7. *Problem Solving*

A system has been described in the previous sections which will create pressures for quality improvement. It frequently occurs, however, that the data are sufficient to show that something is wrong but insufficient to identify the cause. This may be far removed from the point at which the faults are being found.

The carrying out of tests for correlation between results obtained at one check with those at another stage is a valuable technique which will indicate whether a greater state of control is necessary on a particular operation. A typical regression analysis is shown in Fig. 2, which shows the relationship between electrical faults at the welding stage with those at the final stage on a solid tantalum capacitor.

Another more complex regression analysis was carried out to determine which were the major factors affecting unit size in metallized plastic film capacitors. This showed that thickness of film was the major factor but there were two others in the winding operation which were important and which required tighter control.

Figure 3 shows a histogram of the initial leakage current of professional solid tantalum capacitors used on quality assurance tests of 2000 hours endurance. The black histogram shows the initial leakage current of those which failed. This indicates that initial leakage current is not an indicator of early failure.

If more data are required special investigation programmes may be prepared by the Quality Engineering section and additional data collected by inspectors of a higher grade known as fault investigators. A cause must be clearly identified before corrective

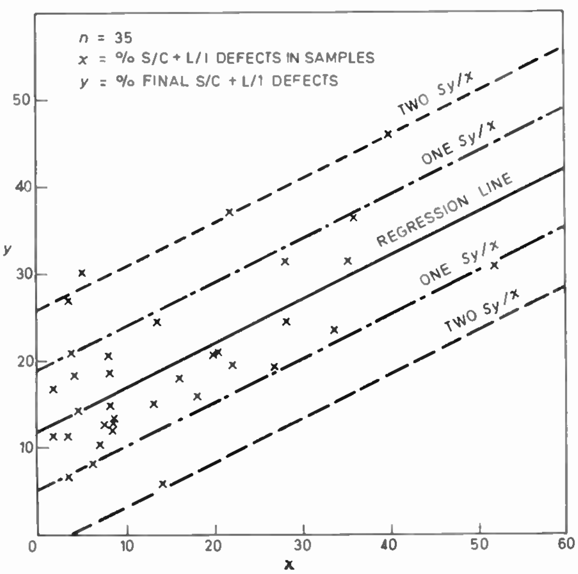


Fig. 2. Regression analysis showing relationship between electrical faults at welding with those at final stage on solid tantalum capacitor.

action is taken and the action must be clearly identified as the solution to the problem before implementation. Whenever possible this should be by laboratory tests and small scale trials. When satisfactory evidence has been obtained the following procedures apply:

(a) The processing to be carried out must be clearly defined by a temporary change note to the manufacturing specification.
(b) The quantity of components to be processed to this revised specification must be clearly defined.
(c) Responsibilities for supervising and monitoring the trial must be defined.
(d) All data must be clearly identified with the batches concerned.
(e) These data must be analysed and tested for significance compared with normal shop production, before a decision is taken to implement the change.
(f) The Chief Inspector must decide whether the components made in the trial are satisfactory for shipment, if necessary by the carrying out of additional tests. Approval of a significant change must be obtained from the Inspection Authority.

4. A Typical Quality Control System

A quality control system must cover all stages of production from raw materials to shipping and packaging. The objective must be to carry out checks at the lowest possible cost at all stages of production so as to ensure that control is maintained. It may cover any of the following:

(a) Ensure that manufacturing specifications are adhered to.
(b) Ensure that the percentage defective at a particular operation is at a satisfactory level.
(c) Ensure that before passing to the next stage the quality of the work meets a specified AQL on a batch by batch basis.
(d) Provide data on partly processed materials which enable later processes to be carried out within specified limits.
(e) Give warning that tools, process materials etc., have reached the end of their life and require replacement.
(f) Ensure that the finished product conforms to specification requirements.

On the other hand, it is quite unnecessary for every production step to be checked all the time. If a check is showing no useful information it should be discontinued in favour of more concentrated effort elsewhere. It must be emphasized that flexibility and readiness to change to meet new situations are essential.

The system to be described relates to metal-cased solid tantalum capacitors but the general principles have been applied to the other capacitor manufacturing lines.

4.1. *Solid Tantalum Capacitors*

The production of these capacitors was described by R. F. Leech[8] and while some of the details have changed, the basic principles are still the same. Figure 4, which is reproduced here from that article, shows detail of construction which may assist in the understanding of the quality control system.

A schematic diagram of the quality control system is shown in Fig. 5. This shows each main step in the manufacturing process, the materials used and the checks that are carried out at each step. All process checks are the subject of a defect prevention routine which details the checks to be made, the standards to be met and the action to be taken in the event of failure.

4.2. *Cumulative Attributes Charts*

Control charts used at the various stages vary according to the type of control required and to whom this feedback is primarily directed. Control by attributes has the merit of simplicity and is readily understood by production operators.

If it is required to determine the percentage defective at a particular operation without attempting to sentence the work, quite a low level of sampling is usually sufficient; e.g. 5 samples every four hours. Reference to the sampling tables will show that no failures in 5 samples taken from a large batch is not significant. However, in a week on day shift 50

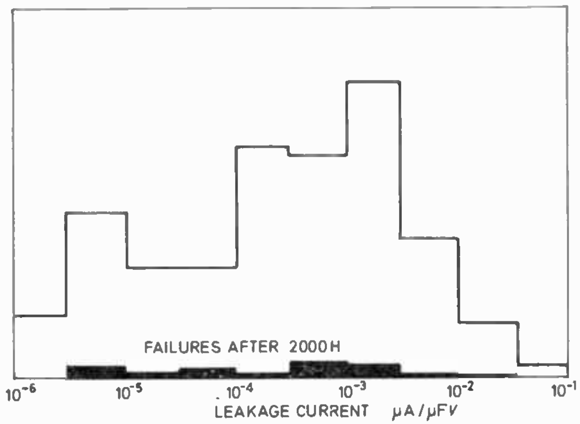


Fig. 3. Plot of distribution of initial leakage current of a particular solid tantalum capacitor.

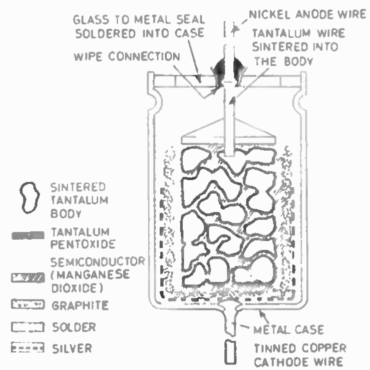


Fig. 4. Dry-sintered-anode tantalum capacitor.

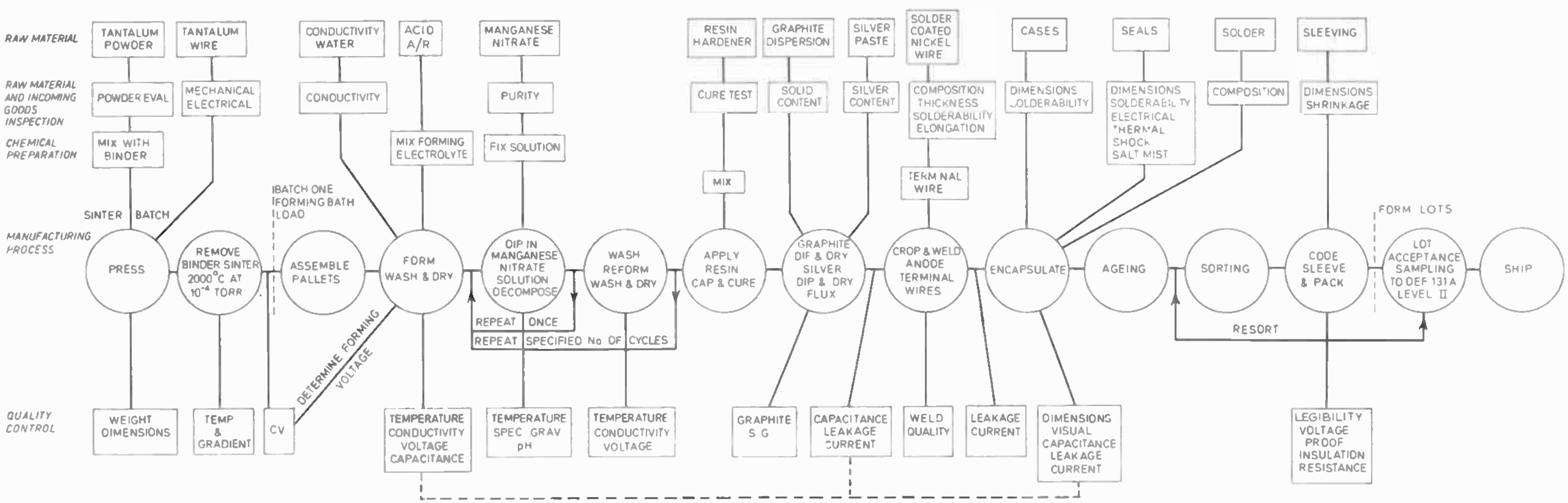


Fig. 5. Schematic diagram of quality control system.

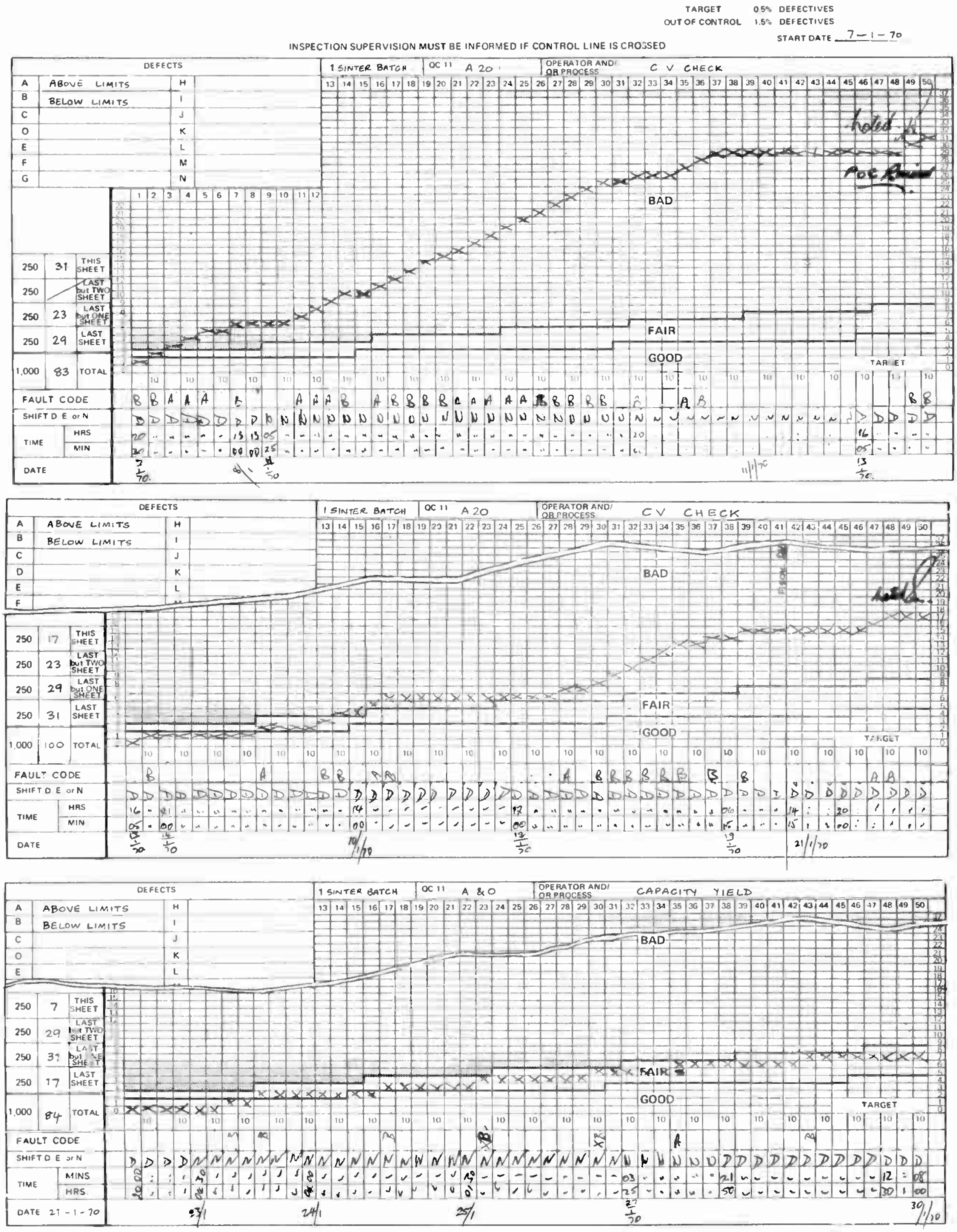


Fig. 6. Cumulative attributes charts (abridged for convenience of reproduction).

samples will have been checked. Cumulative attributes charts used for this purpose are shown in Fig. 6. These charts accommodate a total of 250 samples and show control limits for the stated percent defectives. Both target and action limits are given and added significance from the operator's point of view is given to the data by green and red colouring.

The charts shown in Fig. 6 cover the period 7th to 30th January, 1970. The first chart shows a very unsatisfactory state of affairs in the CV value of a particular anode after sintering with 31 out of 250 samples defective and a total of 83 out of 1000 on the last four sheets. Feedback for corrective action had already taken place on the earlier charts without a solution being found. However, when effective action was finally taken subsequent charts quickly returned to normal. Feedback for corrective action is dealt with in Section 4.4.

4.3. *Combined Variables and Attributes Charts*

When it is necessary to monitor a variable the conventional median and range chart is useful as giving a visual presentation of the spread of values usually for the benefit of engineers, supervisors etc. Even here, however, a cumulative attributes analysis is of value for reporting purposes and Fig. 7 shows a combined variables and attributes chart. This is of the capacitance after forming and the cumulative totals give a clear indication of the proportion defective.

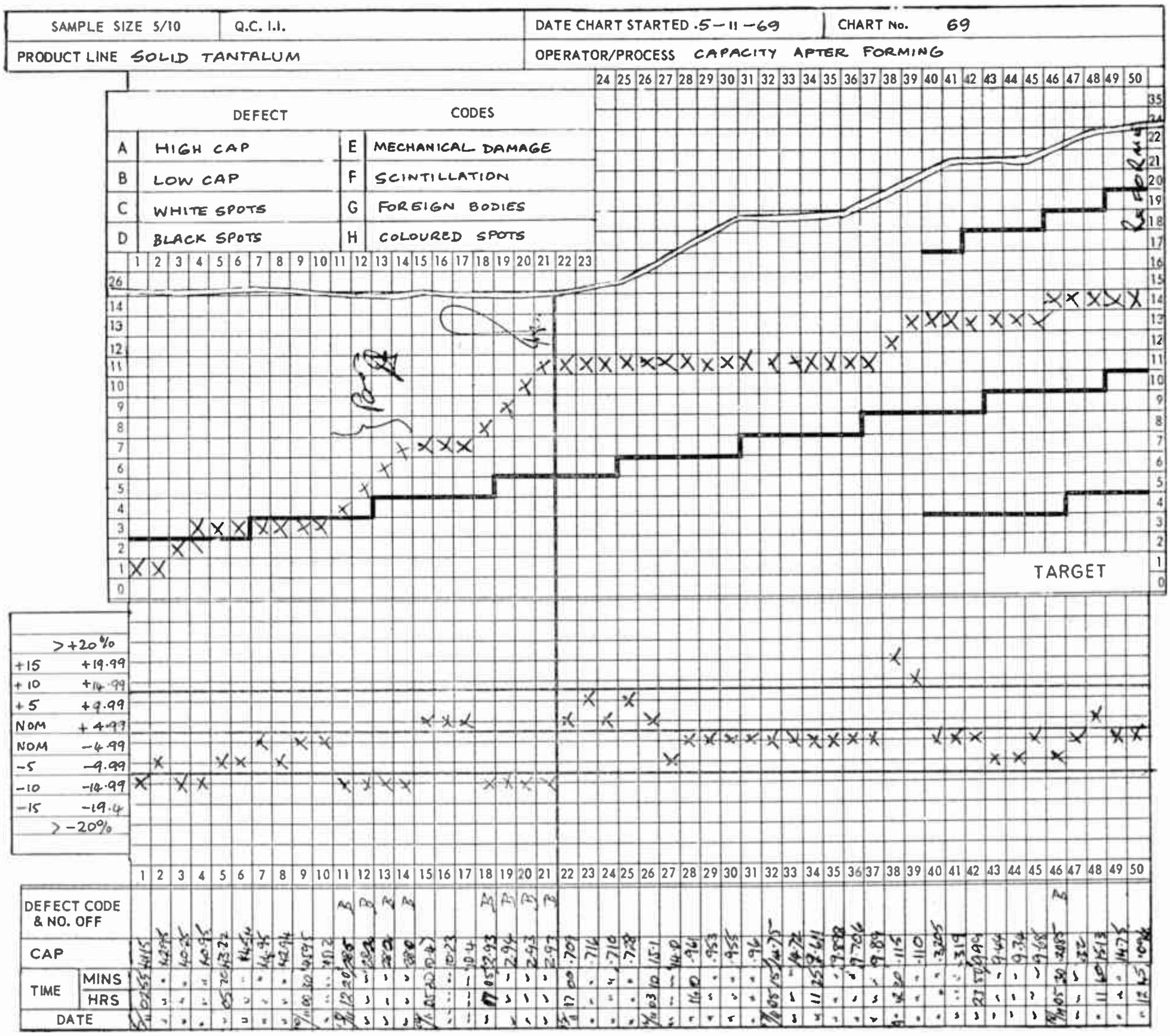


Fig. 7. Variable/cumulative attributes quality control charts (abridged).

PROCESS OUT OF CONTROL - FEEDBACK FOR CORRECTIVE ACTION

ROUTING	SIGNATURES REQUIRED	1st and 2nd APPLICATION	2nd and 3rd APPLICATION	4th and SUBSEQUENT APPLICATION	APPLICATION NO.
1	SHOP	SHOP FOREMAN	PRODUCTION SUPT. ✓	PRODUCTION MANAGER	1
2	ENG.	ENGINEER	ENG. S.H. ✓	CHIEF ENGR.	2
3	PROD. ENGR.	PRODUCTION ENGINEER	PROD.ENG. S.H. ✓	CHIEF PRODN. ENGR.	3
4	QUALITY	INSP'N FOREMAN	DEPUTY CHIEF INSP. ✓	QUALITY MANAGER	4
					5 See Note

PROCESS Solid Tantalum

INSPECTION OPERATION C V Result SPECIFICATION DPR A 20

FAULT ANALYSIS TOTAL SAMPLED 50 PER CENT DEFECT. 62%

FOR DETAILS SEE CHART ATTACHED

	RECOMMENDATIONS FOR CORRECTIVE ACTION		YES	NO
SHOP	RECOMMENDED ACTION: I understand this is being investigated by Eng. Dept. SIGNED	WILL DEFECTIVE PARTS SATISFY MANUFACTURING REQUIREMENT? THIS IS AN ENGINEERING SHOP INSPECTION, or EQUIPMENT PROBLEM.	●	
ENGINEERING	RECOMMENDED ACTION: Most of this problem has been eliminated by the new power settings. However during the present change of water supply some sinters have been at an incorrect power setting. This is now corrected. SIGNED	THIS IS AN ENGINEERING SHOP INSPECTION, or EQUIPMENT PROBLEM. CAN SPECIFICATION/DRAWING LIMITS BE RELAXED.	✓ ✓ No	✓ ✓
PROD. ENGINEERING	RECOMMENDED ACTION: Sintering control is only one of several factors affecting C.V. Further control work must wait until No. 3 furnace makes experimental time available. I agree that such work ... 9/2/70. SIGNED.	IS THE MACHINE/PROCESS CAPABLE OF MAINTAINING SPECIFICATION.		
QUALITY.	RECOMMENDED ACTION: Noted CV values now in control by adjusting temperature to suit required CV value. 9/2/70. SIGNED			

NOTE. THE OBJECT OF THIS FORM IS TO DRAW YOUR ATTENTION TO THE FAILURE OF THE SHOP TO CONFORM TO SPECIFICATION REQUIREMENT. YOUR RECOMMENDATION FOR CORRECTIVE ACTION IS REQUESTED. THE NEED FOR MORE THAN FIVE OF THESE APPLICATIONS WILL BE TAKEN AS AN INDICATION OF A MAJOR DIVISIONAL PROBLEM.

COMPLETED FORM MUST BE RETURNED TO 58420

Fig. 8. Feedback for corrective action form.

4.4. *Feedback for Corrective Action*

Control charts in operating positions are effective provided that corrective action is within the power of production operators or supervision. If, however, the problem is more complex, there must be a procedure for obtaining assistance from Engineering or higher levels of management. If on completion of the chart it has shown that the manufacturing operation is seriously out of control, a form is circulated as shown in Fig. 8. This form is self-explanatory and ensures that the following actions are taken:

Event	*Action taken by*	
Each fault	Operator's immediate supervisor.	
Action line crossed	Shop foreman signs chart.	
Chart out of limits on completion	Inspection foreman originates feedback for corrective action form.	
1st and 2nd application	Production foreman Production engineer Inspection foreman Design engineer	All are required to recommend action.
3rd and 4th applications	Shop superintendent SH production engr. Chief inspector SH engineering	All are required to recommend action.
5th and 6th applications	Production manager Chief production engr Quality manager Chief engineer	All are required to recommend action.

4.5. *Forward Control*

Most of the control of these checks is 'backward', that is to say, it is effective on subsequent batches. A case where forward control is possible on this line is in the determination of the forming voltage. Although powder weight is carefully controlled, the combination of the variability of this with variability of sintering conditions can give rise to a shift in the average capacitance yield (normally specified in μF.V). After sintering a sample is taken from each batch and the yield determined by forming to a specified voltage. This value is recorded against this sinter number and used in production to determine the exact value of forming voltage within certain prescribed limits which may be used to give the required nominal capacitance. Using this method the mean is typically within $\pm 3\%$ of nominal.

4.6. *X-ray*

One of the biggest problems in the manufacture of this capacitor is control of the final assembly. This operation is carried out semi-automatically but there are a number of features of the machine which are operator dependent. Unsatisfactory assembly was at one time a significant factor in the reliability of this capacitor particularly under vibration conditions. For this reason batch by batch sampling was introduced immediately after assembly, and the Defect Prevention Routine for X-ray examination is shown in Table 2. The number of samples was chosen so as completely to fill one X-ray plate. The operating characteristic for each case size is shown in Fig. 9.

Figure 10 shows typical major and minor defects. The introduction of this scheme revealed many problems the solution of which has resulted in a continuous improvement in performance. Typical of these were the following:

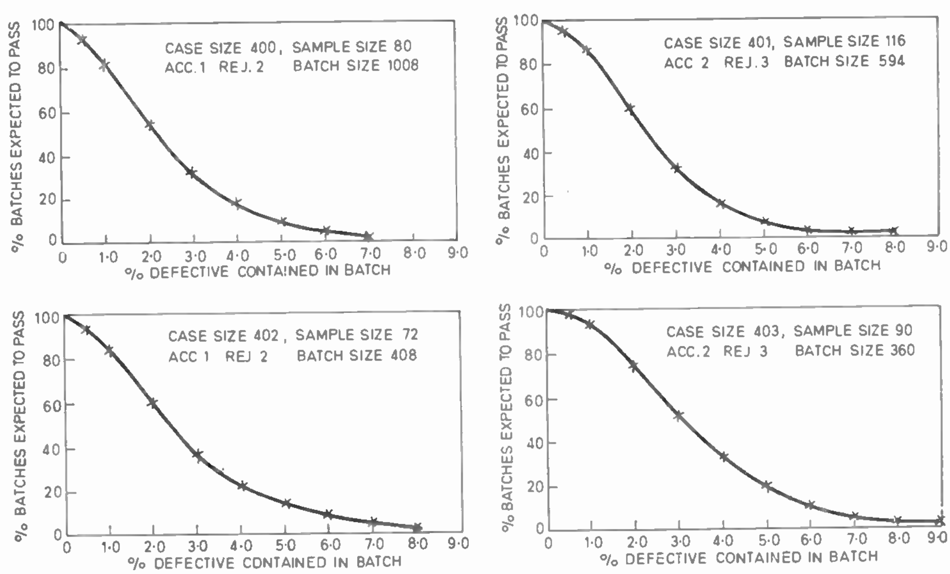


Fig. 9. Operating characteristics.

(a) Poor solderability of cases and seals. This was partly due to supplier and partly to the use of parts which have exceeded their storage life.
(b) Solder pellet wrong size.
(c) Incorrect quantity and type of flux.
(d) Welds too high up.
(e) Temperatures incorrect.
(f) Jigging did not permit correct settling of body.

4.7. *Causes of Rejections*

This type of capacitor is typically associated with a yield of 80–90%. Analysis of the rejections shows the following:

Short circuit	50%
Leakage current	7%
Capacitance	5%
Mechanical	12·5%
X-ray faults	12·5%
Miscellaneous	13%
	100%

It is interesting to analyse the causes of these rejections and consider the factors which contribute to them. This is illustrated in Fig. 11 which shows the main faults which may occur at each stage of manufacture and their effect. The large number of faults which may lead to leakage current and short circuit failures should be noted. This is a fruitful area for correlation studies.

4.8. *Ageing, Sorting and Acceptance Sampling*

This capacitor exhibits a form of self-healing characteristic which is difficult to detect but which can, under suitable conditions, stabilize the characteristics of the device. This is achieved by means of an ageing process consisting of:

(a) One day at approximately surge voltage at 85°C in a 5000-Ω circuit.

(b) One day at rated working voltage at 85°C in a low impedance circuit with fuses. Fuse failures are rejected.

Table 2. Defect prevention routine

COMPONENT/PROCESS: X-ray Examination — TYPE: Solid Tantalum Capacitors

1.0 *Introduction*

1.1 This specification describes the sampling methods and X-ray analysis in checking the Solid Tantalum encapsulation process.

2.0 *Method*

2.1 Immediately the encapsulation process has been completed, a sample taken at random from each batch will be loaded on the cards provided, ready for X-ray. The capacitors must be loaded as specified in Process Planning Instruction No. 239. The sample size will be determined by the relevant sampling plan, see paragraph 3.0.

2.2 All capacitors that have had a seal and/or pip repair must be X-rayed. *These are not to be considered as part of the sample given in paragraph 3.0.*

2.3 Before examining the X-ray the following information must be entered on the plate, Batch No., Batch Type and repairs must be identified.

2.4 Carefully examine the X-ray plate identifying all defects. See Fig. 10 for description of defects.

2.5 The inspection result and decision must be entered on the plate.

2.6 The inspection result plus the information on the plate must be recorded on Form 1164PA.

2.7 The estimated process average (E.P.A.) for A. B. & C. defects only and the E.P.A. for all defects will be calculated when the accumulative sample size is 1000 + 50. Plot this result on the chart (402PA), and pass the result to the Shop Foreman.

2.8 E.P.A. — $\dfrac{\text{Total number of defectives in sample}}{\text{Total number sampled}}$

2.9 The X-ray plate and result sheet must be filed so as to be easily available for at least 6 months, and kept for 3 years.

3.0 *Sampling Plans*

Case Size	Sample Size	Accept.	Reject.
400	80	1	2
401	116	2	3
402	72	1	2
403	90	2	3

3.1 The operating characteristic curves for these plans are given in Fig. 9.

3.2 A batch is deemed to have passed if the number of defectives in the sample is equal to or less than the accept figure.

3.3 A batch will be rejected if the number of defectives in the sample is equal to or greater than the reject figure.

3.4 A defective capacitor is one which has one or more major defects as laid down in Fig. 10.
3.5 Rejected batches must be submitted for 100% X-ray inspection. The results of these batches will be recorded on a separate sheet.

4.0 *Definitions*

4.1 *Excess Solder*

4.1.1 Rejectable (Major Fault H). Any capacitor with solder above the anode where the solder reduces the distance between anode wire and case, providing the solder has a solid appearance.

4.1.2 Acceptable (Minor defect M). Any capacitor which appears to have solder above the anode if the solder has a 'misty' appearance.

4.2 *Solder Balls*

4.2.1 Rejectable (Major Fault D). If the size of solder balls is such that they could, when side by side, bridge the gap between case and anode wire, the capacitor must be rejected.

4.2.2 Acceptable. If the solder balls are small and could not bridge gap between case and anode wire when joined in line.

They are then passed to the sorting equipment for 100% automatic test. The test equipment is shown in Fig. 12 and it carries out the tests shown in the decision chart in Fig. 13. Failures are bench tested and returned through the equipment.

After coding and sleeving lots are formed corresponding approximately to one day's production. These are then sampled in accordance with DEF 5134-A-1, Inspection level II, 1% AQL. Accepted lots are forwarded for shipping. Rejected lots are returned for re-sort.

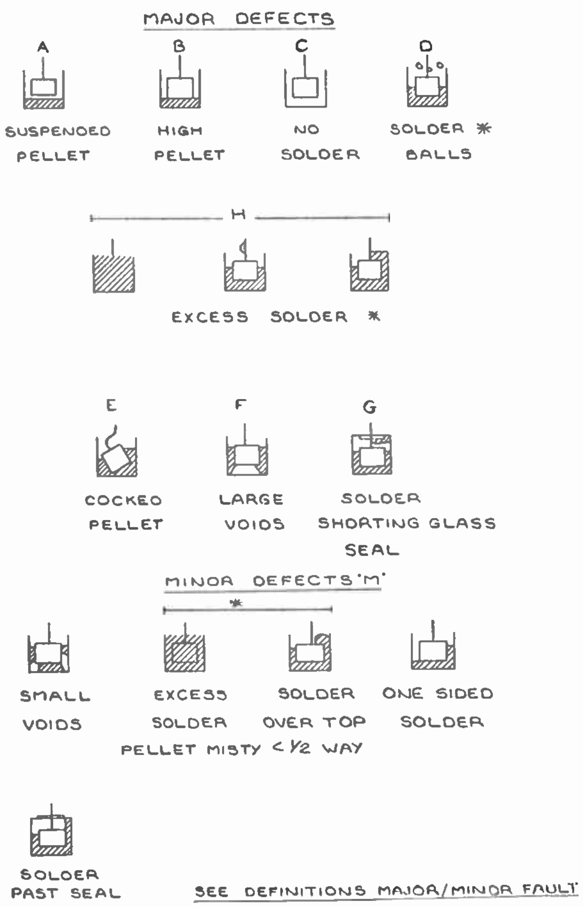


Fig. 10. Interpretation of X-rays of solid tantalum capacitors for major and minor defects.

At regular intervals samples are forwarded to the Test House for quality assurance testing as described in Section 5.0.

There are also cases where a benefit can be obtained by carrying out 'final' test in process. In this case sampling is carried out strictly in accordance with the sampling plans to determine whether batches should be accepted for a particular parameter to meet sampling requirements, results in the batch being rejected for sorting by production operators. A good example of this was the moulding of silvered mica capacitors where some difficulties were being encountered with defective mouldings. The carrying out of final test in the normal way would be one to two weeks after completion of the moulding operation. It was, therefore, impossible to identify closely conditions under which the mouldings had been made or particular moulds which had been employed. By moving final test to immediately after the moulding operation the moulding operators were able to see the faults for themselves. Defective moulds were quickly identified and a substantial improvement in performance resulted.

Fig. 12. Automatic test equipment for production sorting.

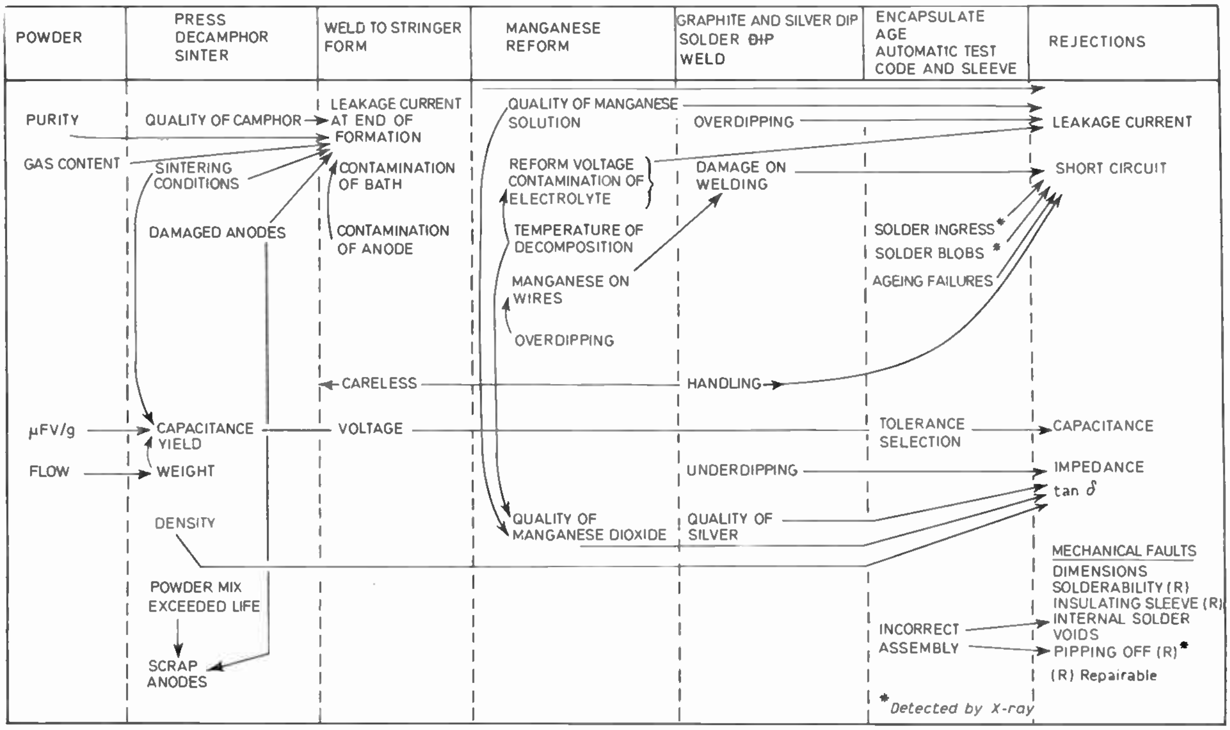


Fig. 11. Factors contributing to rejections of metal-cased solid tantalum capacitors.

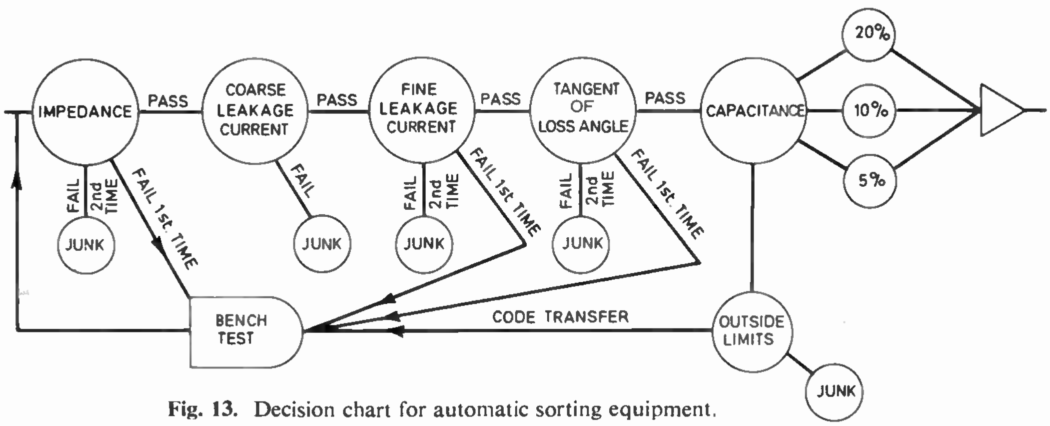


Fig. 13. Decision chart for automatic sorting equipment.

5. Quality Assurance Testing

The quality control scheme so far described ensures that the component is manufactured in accordance with the design and that the parameters fall within the specification requirements on a lot-by-lot basis.

In order to be satisfied that the component will perform satisfactorily under the wide range of environments it is likely to encounter in operation, regular quality assurance testing is essential. This phrase is used to refer to routine tests carried out either on a lot-by-lot or a periodic basis. These tests may be either destructive or non-destructive.

The range of environments likely to be encountered cover the following:

Transport and assembly
Bumping
Solderability and resistance to soldering heat
Robustness of terminations
Humidity
High or low temperatures including changes from one to the other
Mechanical stress in the equipment

Vibration
Acceleration
Shock
Operation under conditions likely to cause maximum deterioration, normally maximum stress and maximum temperature
Low air pressure
Miscellaneous—salt mist, mould growth, flammability etc.

These methods covering this range of conditions are well established and specified in BS 2011.

The wide diversity of specifications and test methods is at the present time a major embarrassment to the component manufacturer in formulating an economical quality assurance test programme. Ideally there should be one standard system of test methods recognized internationally which could be applied to the whole of the output. The lack of such standardization has the following effects:

(1) It fragments production lots resulting in unnecessarily high levels of testing.

(2) Differing test methods require additional equipment and reduce the efficiency of testing.

(3) Changes in mix result in changes of work load on the test house. It is not possible to have people available to meet peak load situations.

(4) Differences in test methods give difficulty in comparing test results with attendant difficulties in obtaining approval, particularly in other countries.

(5) Even physically identical products tested to different specifications are not interchangeable. This gives the customer problems in double sourcing.

In order to overcome some of these problems and to provide sufficient data for feedback to Engineering and Manufacturing we established a standard quality assurance test programme. This is summarized in Table 3. This programme ensures that regular testing is carried out on all the key features under maximum stringency. The object of this programme is to detect points at which the product does not conform to specification so that corrective action can be taken. The levels of testing are in accordance with the requirements of DEF 5001C. In addition to this the endurance test is extended to 10 000 hours on some product lines in order to provide limited reliability data.

5.1. *Automatic Measurement and Data Analysis*

It is not proposed to describe in detail the environmental test methods used in the test house. These are well known and have been described in the literature. Now that the manufacturer has been made fully responsible for testing the product the cost of testing is becoming a primary consideration and just as important as that of production. The need for reliability data and certified test records is likely to increase with time requiring new approaches to testing methods.

Self-balancing bridges are now available in a variety of forms utilizing digital balancing and giving a direct read-out of capacitance and tangent of loss angle, together with a binary coded output which can be used to drive a typewriter or a punched tape unit. These equipments have a rapid response and in order to utilize this it is necessary to load the components in groups and to switch automatically from one to the next. An equipment of this type is shown in Fig. 14. It consists of a self-balancing bridge and a digital voltmeter so that measurements of capacitance (accuracy 0·1%), tangent of loss angle and leakage current (insulation resistance) can be measured in one test sequence. For the latter measurement electrification times are controlled accurately to 1, 3 or 5 minutes, as specified.

Components are mounted on test jigs in groups of ten. These jigs have been designed so that they can be inserted in any of the environmental test equipments and the components remain mounted throughout. These test jigs are shown in Fig. 15. These equipments have an overall testing rate of 400 per hour. The typewritten output goes on to white gummed tape which is then stuck on to an appropriate report form giving the necessary tabulation. Reports are then issued by photocopying these sheets thus eliminating transcription errors. A punched tape unit is being fitted to this equipment to give the data in a form in which it can be fed through a reactive terminal and a data link on to a computer. Various programs have been prepared to analyse the data. An example of the various ways in which analysis can be carried out is shown in Fig. 16.

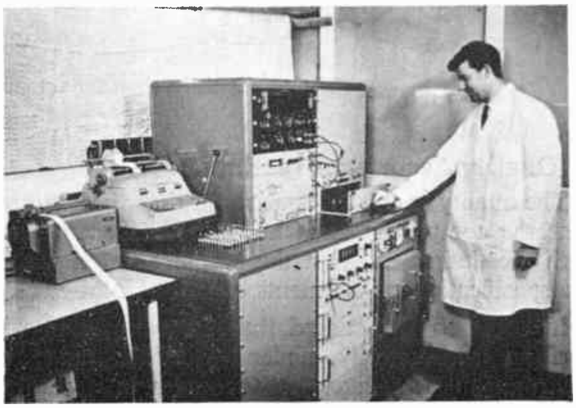


Fig. 14. Semi-automatic capacitor test equipment employing self-balancing bridge and digital voltmeter with print-out and punched tape output.

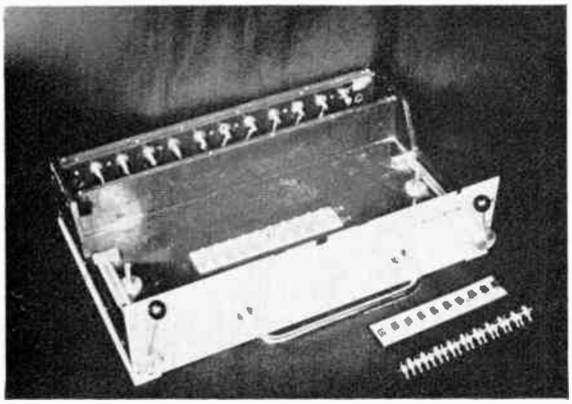
Table 3. Summary of quality assurance scheme

Group	Test	Applicable to	Frequency
A (Recorded Sample)	Capacitance Tangent of loss angle Leakage current or insulation resistance Impedance (when applicable) Solderability	All	5 per code per lot greater than 50
B	Robustness of terminations Solderability Resistance to soldering heat Initial measurements Vibration endurance (100 hours) functional (one sweep) Temperature rapid change (5 cycles) Climatic sequence Final measurements Voltage proof to destruction	Solid tantalum metal case* Solid tantalum resin dipped* Metallized plastic resin moulded* Mica	Monthly
C	Initial measurements Endurance (2000 hours) 2 acc. damp heat cycles Final measurements	Tantalum foil (three types) GPO paper Metallized plastic metal case Polystyrene*	Quarterly
E	Long term damp heat		
F	Shelf life samples from 'E' stored up to 5 years Measurements as required only		
G	Endurance 10 000 hours (continuation of 'C')	Types above marked *	As for 'C'
D	Temperature coefficient and cyclic drift	Mica Solid tantalum metal case All others	Monthly Quarterly Yearly
H	Other tests as specified	All	Every 3 years
J	Solderability after ageing	All	Monthly

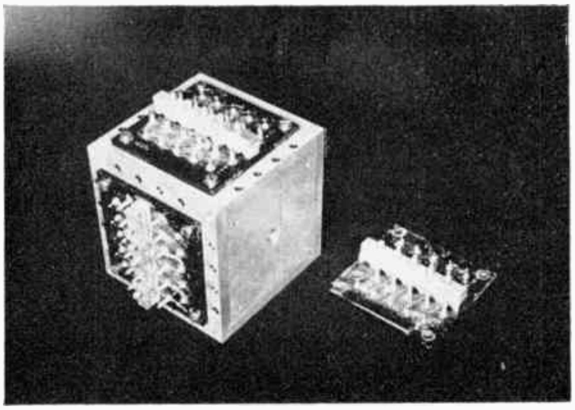
Another equipment which has proved a valuable tool for quality assurance testing was described by Beresford and Williams.[9] This is a plotting table to plot the variation of electrical parameters as a function of temperature on an automatic temperature cycling basis. A typical plot of a solid tantalum capacitor is shown in Fig. 17. For diagnostic purposes it is also possible to plot other parameters; for example, Fig. 18 shows a plot of leakage current as a function of temperature for a capacitor which exhibits an unusually high increase of current at the higher temperatures. Investigation of the causes of such changes has formed a valuable part of quality improvement programmes.

5.2. *Quality Assurance Testing and BS 9000*

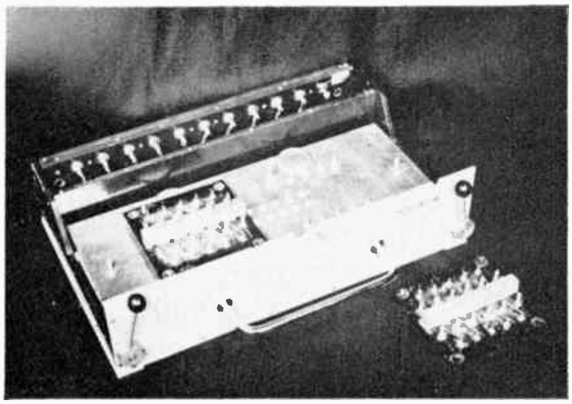
BS 9000 (Electronic Parts of Assessed Quality) was issued in 1967 and 1970 is likely to be the year when the first approved components become available. This scheme is valuable because it lays down a systematic series of tests applicable to the full range of electronic parts. These series of tests vary so as to be commensurate with the application class and the amount of testing which can economically be justified. They will be supervised by the Electrical Quality Directorate as agents of the British Standards Institution which will give the user confidence in the quality of the product, whether supplied by one manufacturer or another. This scheme should, therefore, be regarded as a major step forward in the provision of parts for the electronics industry. One of the requirements is that the manufacturer will be responsible for maintaining adequate control and recording of all manufacturing processes. Another is that the results of all specified quality assurance tests will be made available in the form of certified test records.



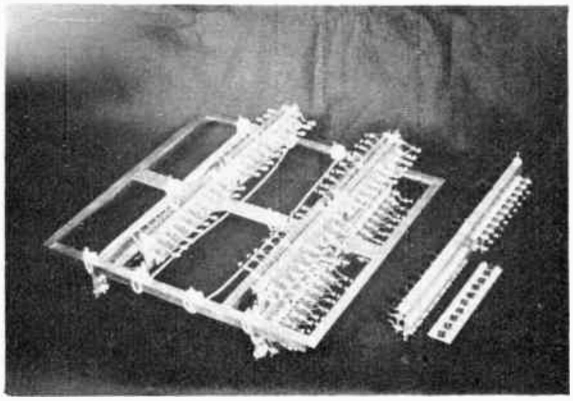
(a) Drawer fitting into the semi-automatic test equipment. The strip in the foreground is used to mount capacitors for all tests except the mechanical and climatic test sequence.



(b) Block for mounting on vibrator giving test in various planes. The plate in the foreground is used for the whole of the mechanical and climatic test sequence.



(c) Drawer containing mechanical and climatic test plates.



(d) Rack holding test strips for endurance or reliability tests. Each tray holds a total of 320 capacitors and four trays go into one oven.

Fig. 15. Test jigs.

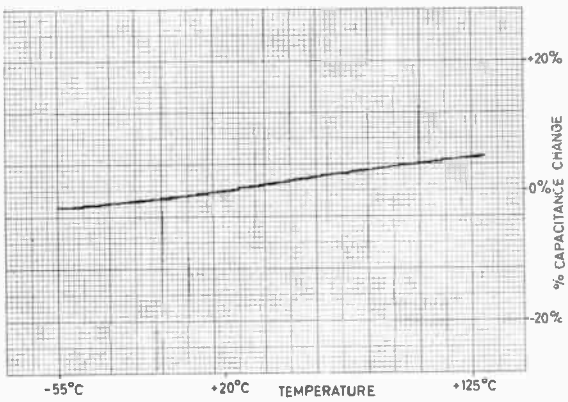


Fig. 17. Capacitance change vs. temperature for solid tantalum capacitor. (2·2 μF; 35V).

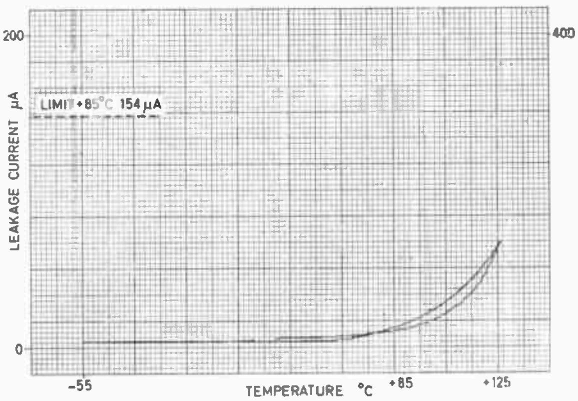


Fig. 18. Leakage current vs. temperature for solid tantalum capacitor.

```
CODE 433GAA

INITIAL MEASUREMENTS        SAMPLE C


? HAS BEEN FOUND FOR CAP. MEASUREMENT 100
THIS VALUE IGNORED, AND ALSO NEXT (PRESUMED TO BE CORRESPONDING TAN DEL.)


** CAPACITANCE ANALYSIS **


                                    HISTOGRAM   1

FREQUENCY          0   1   0   0   6   8  18  21  13   5   9   5   2   5   3   2   0   0   1   0   0   0
 ---------------------------------------------------------------------------------------------------------
EACH  * EQUALS  3 POINTS

   21                                          *
   18                                      *   *
   15                                      *   *
   12                                      *   *   *
    9                                      *   *   *       *
    6                              *   *   *   *   *       *
    3                              *   *   *   *   *   *   *   *       *   *
 ---------------------------------------------------------------------------------------------------------
CELL               1   2   3   4   5   6   7   8   9  10  11  12  13  14  15  16  17  18  19  20  21  22
                    < LOWER LIMIT                                                     UPPER LIMIT >


** LEAKAGE CURRENT ANALYSIS **


NUMBER TESTED = 100     NUMBER PASSED =  97     NUMBER FAILED =   2


MEAN =    -0.01518408          STD,DEV. =     0.03124861

CELL FREQUENCIES
LIMIT                                    <>
      84 11  1  0  0  1  0  0  0  0  2  0  0  0  0


** TAN DELTA ANALYSIS **


NUMBER TESTED = 100     NUMBER PASSED =  99     NUMBER FAILED =   0


MEAN =      0.03207316         STD,DEV. =     0.00392617

CELL FREQUENCIES
LIMIT                                    <>
        0  0  0 58 34  7  0  0  0  0  0  0  0  0  0
```

Fig. 16. Analysis of computer data.

6. Conclusions

Quality control in component manufacture must be able to deal with a technology which is still changing and which is moving towards larger scale production with mechanization. The solution to this process must be to encourage continuous improvement but to maintain strict control on changes before and during their introduction. This is best co-ordinated by yield improvement programmes.

Motivation and the part that motivational programmes can play in quality improvement has not been dealt with in this paper. Quality and Reliability Year in 1967 was the main effort in the U.K. in this direction. It was valuable in establishing an awareness at the operating level and there is no doubt that some benefits were obtained. It is by no means certain, however, that it penetrated deeply enough to have a lasting effect. In the United States the Zero Defects Campaign, which has very much the same objectives, requires even more positive participation and affirmation by personnel concerned.[10] These effects seem to have had substantial benefits in the United States. There is a need for the introduction in the U.K. of a similar philosophy adapted to the needs and to the temperament of the personnel concerned.

As the element of skill required of the operator is reduced it may be desirable to encourage participation by allowing him to carry out his own quality control checks and we expect this to be the line of development in the immediate future. In the long term, however, automatic testing methods with computer analysis of data are likely to become the primary tools of quality control.[11]

7. Acknowledgments

The permission of ITT Components Group Europe, Capacitor Division, Paignton, Devon to publish this paper is gratefully acknowledged. The author also wishes to place on record his appreciation of the assistance of his colleagues, particularly Messrs. D. A. Beresford, P. E. Dillon, R. Kingdon and T. A. Wagstaff, in the preparation and selection of drawings and examples.

8. References

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9. Appendix: Definitions

Quality The quality of a product is the degree to which it meets the requirements of the customer. With manufactured products quality is a combination of *quality of design* and *quality of conformance*. (EOQC)

Quality control A management system for programming and co-ordinating the quality maintenance and improvement efforts of the various groups in a design and/or manufacturing organization so as to enable production at the most economic levels which allow for full customer satisfaction. (EOQC)

Quality of design The value inherent in the design; a measure of the excellence of the design in relation to customer's requirements. (EOQC)

Quality of conformance A measure of the fidelity with which the product taken at the point of acceptance conforms to the design (or specification). (EOQC)

Reliability The ability of an item to perform a required function under stated conditions for a stated period of time. (BS 4700, Part 2, 1967).

In addition the following terms are used in this paper:

Performance specification Any specification defining the application class, environmental conditions and the electrical and mechanical performance of the component. It will also define tests under the following headings.

Qualification approval tests Tests carried out to a performance specification for the purpose of approving a design. Approval is granted by the customer or his representative.

Acceptance sampling tests Tests carried out on samples taken from a lot for the purpose of accepting the lot.

Quality assurance tests Tests carried out on samples taken periodically for the purpose of assuring that the requirements of the performance specifications continue to be met.

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Horizontal Aperture Equalization

By

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In any scanning process, such as a television picture or a tape or film sound track, some fine detail (that is, high frequency) information is lost since the scanning spot or slit is not infinitesimally small. This loss, or smear, is usually symmetrical in space and therefore, in the electrical output, symmetrical in time. Thus for adequate waveform response, the compensating 'aperture' equalization must have a symmetrical time response and hence a linear phase response, together with a frequency response that approximates the inverse Gaussian exp (ω^2T^2).

Various methods of achieving the desired response are examined:

(i) fixed aperture equalizers where the frequency response is corrected to a high precision and the resulting delay error is then compensated by a separate all-pass network; this leads to

(ii) another fixed equalizer which approximates the first method in a very simple manner;

(iii) the conventional variable equalizer using
  (a) two terminated delay lines, or
  (b) one delay line unterminated at the far end to produce the 'echo pair' required for linear-phase equalization;

(iv) an improved version of (iii) (b) where a cascaded pair of delay lines (or one tapped in the middle), unterminated at the far end, are used to produce two echo pairs and thus achieve much more precise equalization;

(v) versions of (iii) and (iv), simplified by replacing the delay lines with one-section and two-section linear-phase low-pass filters, without any worsening of the response.

Testing methods are described and the quantitative effect of aperture equalization on noise is demonstrated.

The devices described in the paper produce horizontal aperture equalization. The same principles can be applied to vertical aperture equalization, in those devices that involve delay lines, by substituting a delay of one line (64 μs) for each delay of one picture element (0·1 μs).

1. Introduction

Aperture equalization is, after all, only a specialized kind of equalization in which the high frequency response is increased without introducing delay distortion or, alternatively, the waveform transitions are sharpened‡ without losing symmetry. It is necessary quite often that the equalization can be varied by a simple control. This paper hopes to demonstrate some improvements in the technique and since the method of operation is best understood at some times in terms of frequency and phase response, at others in terms of transient response, it is hoped that the discussion in this paper, alternately in the frequency and time domains, may help the reader to a better insight into the overall problem.

When a television signal is produced by scanning the original image along a horizontal line, some fine detail is lost due to the finite size of the scanning spot or aperture. The aperture equalizer is the device used to correct these losses as far as possible. Aperture losses are common in most analogue (that is, proportional) transducers used to convert optical, or mechanical or magnetic, signals to an electrical form.

Early forms of aperture equalization were those used in the playback of optical sound tracks and in 'radius compensation' of disk recording. In audio practice, however, the high frequency loss is made good with high frequency lift using conventional minimum-phase networks. It is usually, but not always, considered sufficient in audio applications to correct the frequency response without consideration of the phase or transient response since the delay errors introduced in such equalizers are thought to be too short in duration to be heard. However, in television, or more properly, in any application where a *waveform* must be reproduced accurately, the equalizer must be somewhat more sophisticated.

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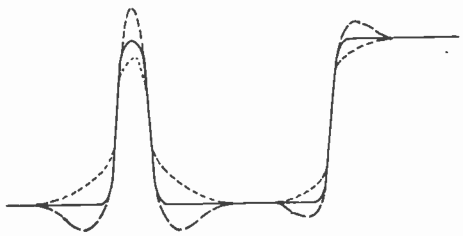
‡ The author prefers 'crispened' but the expression has acquired a special meaning in B.B.C. usage.



———— Original pulse and bar.
- - - - - - Pulse and bar, with aperture loss.
— — — — Pulse and bar, with aperture equalization.

Fig. 1. Aperture correction of a waveform.

From the manner in which they are produced, aperture losses distort waveforms symmetrically either side of a transition. Thus a sine-squared pulse input as in Fig. 1 is smeared symmetrically about its base.

This can be restated in terms of steady state response, as an attenuation of high frequencies without phase error. Both concepts, of transient and of steady state response, will be drawn upon in the work that follows. When the subject is viewed from two viewpoints at once, it may be perceived in greater depth.

2. The Shape of 'Aperture-Filtered' Response

The first step in designing an equalizer is to know what response has to be equalized. This may seem obvious enough, yet response difficulties often arise through over-simplification of the matching response and can be rectified quickly by a little attention to detail.

It has been shown elsewhere, for example, how the response of a video cable[1] is very poorly equalized by a simple high frequency lift but can be easily and accurately equalized with multiple sections. It has been shown again how the 'loss of definition' in telecines and telerecordings[2] can be due in large part to 'flare', that is, longer time symmetrical smears (but that this is far less tractable).

In the present problem, while it has been stated that the conventional type of equalizer, considered in more detail in Section 4, provides a 'suitable' shape of response, this can be very easily disproved by looking with a picture monitor at the output of a television camera trained on a resolution wedge when the aperture equalization is adjusted 'too high'. The effect usually is to make the white spaces between the tapering black bars too white, that is, higher in amplitude than the peak white of the rest of the test card, not so much at 400 lines per picture height (corresponding to approximately 5 MHz, the highest video frequency transmitted) but nearer 200 lines, that is, 2·5 MHz. Thus, the shape of conventional aperture equalization is a poor fit and for a given amount of lift at the highest frequencies, it produces too much lift at the middle frequencies. What then can be taken as a suitable shape for the equalizer?

If the scanning 'spot' is rectangular in shape, its spectrum will have a $(\sin x)/x$ shape. If, on the other hand, the distribution of electrons across the width of the scanning spot follows the Gaussian law

$$d = k_1 \exp(k_2 x^2) \qquad \ldots\ldots(1)$$

where d is the density of electrons,
x is the distance from the centre of a spot,†
and k_1 and k_2 are simply scaling factors,

then, by a rather remarkable property of the Gaussian error function, its spectrum will be a Gaussian error function also, of amplitude

$$\exp-(f/f_0)^2$$

where f_0 is the frequency where the response is down one neper (8·69 dB) and the neper (or decibel) loss is proportional to frequency squared.

The spectrum of the spot, as it has been called above, is more properly the response of the low-pass filter that would simulate the loss of high-frequency (that is, fine detail) information due to the scanning process. Thus the equalizer response should be the inverse of the spot spectrum.

To examine their properties further, the expressions are expanded:

$$(\sin x)/x = 1-(x^2/6)+(x^4/120)-(x^6/5040)+\ldots \qquad \ldots\ldots(2)$$

$$\exp(-y^2) = 1-y^2 \quad +y^4/2 \quad -y^6/6 \quad +\ldots \qquad \ldots\ldots(3)$$

and if the second order coefficients are matched, then

$$y^2 = x/6 \qquad \ldots\ldots(4)$$

and

$$\exp(-y^2) = 1-(x^2/6)+(x^4/72)-(x^6/1296)+\ldots \qquad \ldots\ldots(5)$$

which demonstrates that at the lower frequencies the curves are reasonably similar, though this may be appreciated better from Fig. 2. The deviation increases as the $\sin x/x$ function approaches zero output, that is, a notch in the response, when x is π, at which frequency the Gaussian response coinciding with it at low frequencies would be down $\pi^2/6$ nepers (14·3 dB).

Thus it is possible to nominate roughly (but not too roughly) what shape of amplitude response is required in an aperture equalizer. The shape of the delay (or phase) response is decided more easily, on the assumption that the scanning spot is symmetrical about its

† If the spot has circular symmetry, x will be in fact a radial distance, but, as in this example we are concerned most with horizontal aperture equalization, it is sufficient to consider x as distance along the line. Vertical aperture equalization is another matter again.

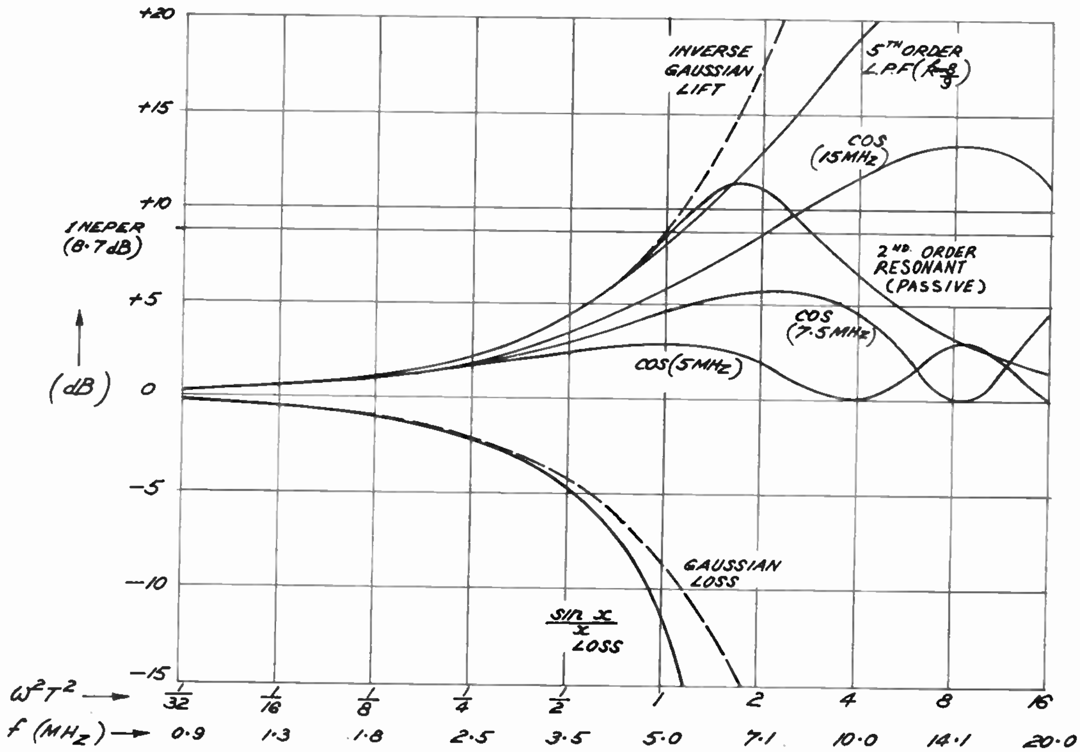


Fig. 2. Typical frequency responses of aperture equalizers.
Losses are shown for comparison.
'Cos' curves (with peaking frequencies indicated) are for conventional equalizers of Section 4.2 and Fig. 7.
'2nd-order resonant (passive)' curve is for equalizer of Section 3 and Fig. 3.
'5th-order l.p.f. ($k = 8/9$)' curve is for highest lift from the improved equalizer of Section 5.2 and Figs. 8 and 10; compare Fig. 11.

centre and hence that the resulting distortion of the output waveform is symmetrical in space and hence, in the waveform, in time. This infers that the distortion is due only to a change of amplitude response, without any change in the delay response and hence that the equalizer must change the amplitude response while maintaining a flat-delay, or linear-phase response.

The kind of equalizer response required has been deduced above on the assumption that the only loss in fine detail being compensated was due to the deviation of the practical scanning spot from an ideal point. In fact there are other sources of distortion in a scanning system due to lens aberrations, defects of prisms in multiplexers, dirt or even finger prints on either, dispersion of light in phosphors or targets, internal reflexion in glass surfaces, a whole range of distortions whose amplitude and time scales may be rather different. However, so long as the effects are symmetrical in space, so that the resulting electrical distortion is free of delay error, and so long as the time scale and shape of the correction are properly fitted to the distortion in the manner to be shown, they are amenable to this method of correction.

It will be clear from the foregoing that an aperture equalizer should be used *only* to compensate aperture-type losses. If an aperture equalizer (a non-minimum-phase network) *is* used to correct conventional minimum-phase type losses due to incorrectly adjusted amplifiers or incorrectly equalized cables,† then the resulting waveform response can be very wrong. Typically, in a pulse-and-bar waveform, a strong pre-shoot is produced when the heights of the pulse and bar are made equal.

3. Methods of Aperture Equalization Using Separate Amplitude and Delay Equalizers

The degree of correction required of a given aperture corrector will most likely change from time to time as camera tubes, lenses, or sometimes even films, are changed. Thus it will frequently be a requirement that the corrector be variable, preferably continuously variable, and it is this requirement that gives rise to the

† A cable is a special case. It is a non-minimum-phase network, but can be considered as producing
(i) a pure delay (non-minimum-phase), cascaded with
(ii) a loss characteristic of the minimum-phase kind.

most familiar form of the aperture corrector, for example, Fig. 7, but although this continuously variable form will be the prime concern in this paper, it is instructive to consider first the more elementary form of equalizer, where firstly an amplitude response of suitable shape and frequency (hence, time) scale is produced and then its delay errors are corrected by an all-pass network.

This approach leads to a network which may be novel and is extremely simple.

It should not be necessary to emphasize that this approach is unsuitable for a continuously variable type of equalizer, though one could, of course, switch networks in cascade to provide increments of lift at 5 MHz, following the progression 1, 2, 4 and 8 dB, for example. However the warning intended here is rather against the type of 'aperture equalizer' so called in at least one, older, equipment where the fixed amplitude response could be varied but the delay equalizer remained fixed!

Taking a very simple amplitude response, that obtained from the single resonant circuit of Fig. 3, then the maximally flat approximation to a 'Gaussian lift' amplitude response is that described by the expression†

$$e_{out}/e_{in} = (1+1{\cdot}9318pT+p^2T^2)/(1+0{\cdot}5176pT+p^2T^2) \qquad \ldots\ldots(6)$$

and it will be seen from Table 1 and from Fig. 2 that it gives a particularly accurate approximation over the first 8 dB of lift to a law of exp $(\omega^2T^2\sqrt{3})$.

By a fortunate coincidence, the fit is *fourth-order* maximally-flat and the actual responses are identical up to and including the terms in ω^8T^8.

If now an all-pass network is cascaded with the amplitude equalizer to make its delay response maximally flat to fourth order, then the overall response obtained by cascading the all-pass response with the original amplitude response is

$$e_{out}/e_{in} = \frac{1-1{\cdot}8768(p1{\cdot}140T)+(p1{\cdot}140T)^2}{1+1{\cdot}8768(p1{\cdot}140T)+(p1{\cdot}140T)^2} \times \frac{1+1{\cdot}9318pT+(pT)^2}{1+0{\cdot}5176pT+(pT)^2} \qquad \ldots\ldots(7)$$

In the particular example of Fig. 3, T is 22·5 ns so that the 7·4 dB lift occurs at 5·0 MHz. The phase delay error at that frequency is 3·7 ns.

It will be seen that, in equation (7), the term $1{\cdot}8768(p1{\cdot}140T)$ is similar to $1{\cdot}9318\,pT$ and $(p1{\cdot}140T)^2$ is similar to p^2T^2. Thus the equalizer response can be approximated reasonably well by the simpler expression.

$$e_{out}/e_{in} = (1-1{\cdot}9318pT+p^2T^2)/(1+0{\cdot}5176pT+p^2T^2) \qquad \ldots\ldots(8)$$

The amplitude response of equation (8) is, in fact, identical with that of equations (6) and (7); the delay response is only slightly worse. It can be realized by the network of Fig. 4. Unity coupling between the two half-windings of the inductor is assumed and it can be approximated sufficiently well by winding the two parts, one over the other, in a miniature ferrite pot core.

Bifilar winding proved unsuitable at first since it lumped so much capacitance across the emitter load resistance as to affect the response, not of the emitter voltage itself but of the portion of the signal coupled through the collector circuit. This can be avoided by taking the signal from the emitter of the first transistor to the cold end of the coil secondary through another directly-coupled emitter follower. By this procedure the cost of an extra transistor, and one resistor, is traded for ease of winding the coil. Unfortunately

† As a matter of interest, the numerator and denominator of equation (6) are both factors of the expression for the sixth-order Butterworth response. They are, in fact, the most heavily and the most lightly damped of the three factors.

Table 1.

Comparison of response of fixed passive equalizer of Fig. 3 with Gaussian equivalent

$\omega^2T^2\sqrt{12}$	$\frac{1}{4}$	$\frac{1}{2}$	1	$\sqrt{3}$	2	$2\sqrt{3}$ (max)
Response (dB)	1·09	2·16	4·33	7·41	8·45	11·44
Gaussian equivalent (dB)	1·09	2·17	4·34	7·52	8·69	15·04

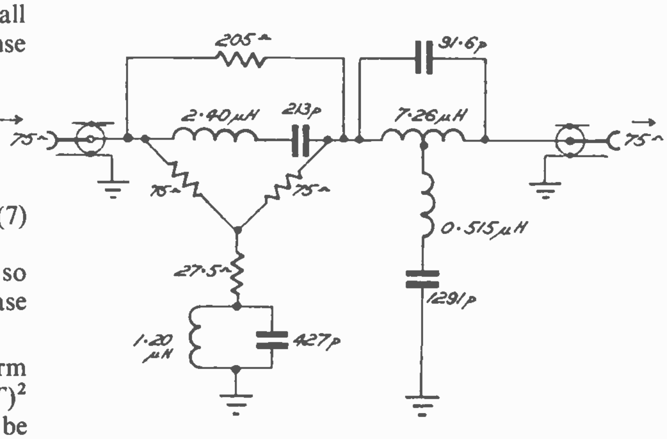


Fig. 3. Passive aperture equalizer.
Resonant equalizer for frequency response, with all-pass correction for delay response. Values are for R of 75 Ω and T of 22·5 ns, hence 7·4 dB lift at 5 MHz and peak lift of 11·4 dB at 7·1 MHz.

this network cannot be realized in an unbalanced constant-resistance form.

However, in the form of Fig. 4 where the numerical values apply to a time, T, of 22·5 ns, it can be realized very easily from an emitter follower to provide aperture equalization *without* any low-frequency loss. It would thus seem attractive for compensating finite spot size in picture monitors or even domestic television receivers and when one considers the fact that practically every television signal needs (and most usually gets) aperture correction at the source, it is surprising, at the least, that such compensation is rarely if ever provided at what could be described as the 'sink'. This network could have a special application in the viewfinders of television cameras, where a high frequency lift *is* usually applied to 'sharpen' the picture and thus facilitate focusing.

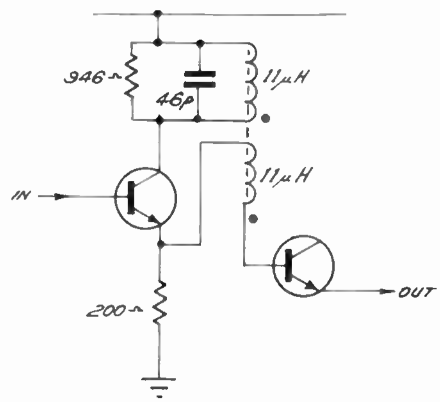
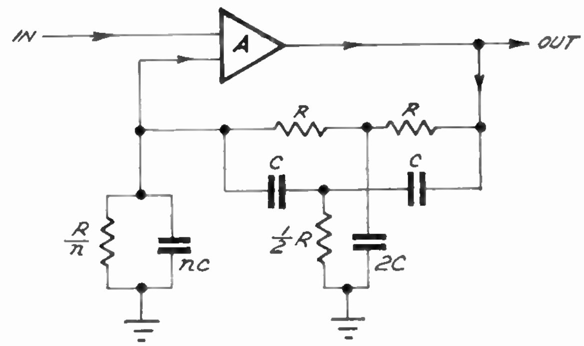


Fig. 4. Simple fixed aperture equalizer with 7·4 dB lift at 5 MHz.

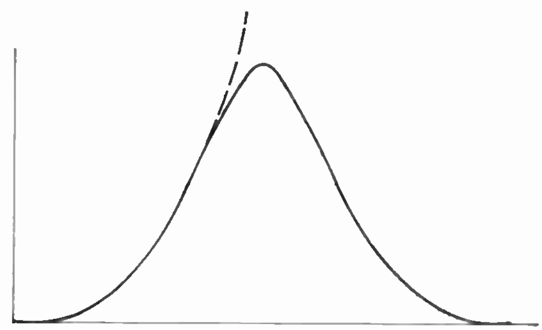
The network cannot be varied easily. However, a set of four stages could be cascaded, with each resonant frequency separated from the next by a half-octave interval, for example, with characteristic time-constants, T of 23·2 ns, 16·4 ns, 11·6 ns and 8·2 ns, and switched to provide steps of 1, 2, 4 and 8 dB equalization at 5 MHz.

The amplitude response of equation (6) can also be approximated when the feedback round an amplifier is filtered by a twin-T network, a method that is particularly suitable at audio frequencies. In Fig. 5, such a network is terminated at the output end by a low impedance, consisting of a resistance, R/n where n is large, in parallel with a capacitance, nC arranged to make the response at high frequencies symmetrical with that at low frequencies. Then the response of the feedback network, assuming that the feed impedance, that is, the output impedance of the amplifier, is small and that n is very large, is

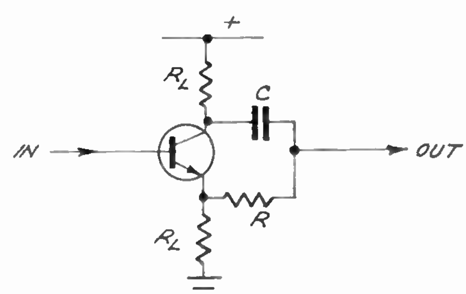
$$e_{out}/e_{in} = (1/2n)(p^2T^2+1)/(p^2T^2+2pT+1) \qquad \ldots.(9)$$



(a) Amplifier with feedback filtered by twin-T network. *RC* product is T of equations (9), (10) and (11).



(b) Shape of response, see Fig. 2 for detail, ideal response dashed.



(c) Simple all-pass network to correct delay response.

Fig. 5. Aperture equalizer suitable for audio frequencies.

where T is the product of R and C in Fig. 5(a). Then the overall response of the feedback amplifier is

$$e_{out}/e_{in} = \frac{[A/(1+A\beta)][pT+2pT+1]}{[p^2T^2+[2/(1+A\beta)]pT+1]} \qquad \ldots.(10)$$

Then, when $1+A\beta$ is made 5 so that the peak lift is 14·0 dB, the amplitude response is as given in Table 2.

Again, the amplitude response is an extremely good fit to the inverse Gaussian response required.

Table 2.
Frequency response realizable with a twin-T network in the feedback path of an amplifier

ω^2T^2	1/16	1/8	1/4	1/2	1
Response (dB)	1·0	2·1	4·2	8·3	14·0
Gaussian equivalent (dB)	1·0	2·1	4·2	8·3	16·7

The delay response of the denominator is small except for frequencies near resonance where ω is $1/T$. The delay response of the numerator, on the other hand, is equalized exactly by a first-order all-pass network, for which

$$e_{out}/e_{in} = (pT-1)/(pT+1) \qquad \ldots.(11)$$

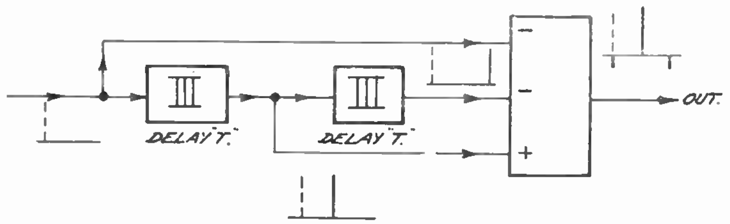
and which is realized by the elementary 'phase-shifter' of Fig. 5(c). By a slight adjustment, the overall delay response can be optimized to take account of the denominator as well at the lower frequencies. Thus a reasonably simple fixed aperture equalizer for audio frequencies is realized.

only the amplitude response is affected; the delay response is unaltered. Secondly, in another part of the circuit, the same pair of echoes is added to each other in opposite senses and when this echo pair signal is added to or subtracted from the main signal, the result is pure skew-symmetrical distortion; hence the delay response is changed without affecting the amplitude response. The overall result is that two controls are still needed for two pairs of echoes but instead of being controls for individual echoes 'leading x μs' and 'lagging x μs', they are controls for two functions 'amplitude x μs' and 'delay (or phase) x μs' and this facilitates adjustment in some applications.

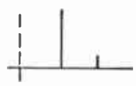
Thus an aperture equalizer can be considered, in fact, as a specialized kind of time domain equalizer, in which the correction signal is *always* a symmetrical pair of echoes taken in the same sense, and is *always* subtracted from the main signal.

In the comparatively few applications where vertical aperture equalizers have been considered justifiable in the past,[3, 4] the method of Fig. 6 is in fact used, with

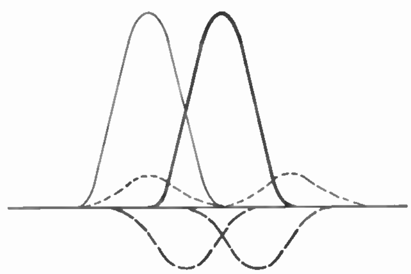
Fig. 6. Conventional aperture equalizer using terminated delay lines in the manner of time domain equalizer.



(a) (above) Block schematic with waveforms, input pulse shown dashed for reference.



(b) Alternative output waveform for pure delay correction, that is, skew symmetry.



(c) Waveforms: solid light curve: original pulse
solid heavy curve: 'main' signal, delayed $2T$
heavy dashed curve: T echo pair
light dashed curve: $2T$ echo pair
used in Section 4.4 and Fig. 8.

4. Methods of Aperture Equalization Using Delay Lines

4.1. *Method of Operation*

The shape of the distortion observed in Fig. 1 suggests a possible method of equalization. The distortion is a smear, symmetrical either side of the main pulse. Thus if a pair of pulses, symmetrical in time fore and aft, are subtracted from the main pulse, as in the heavily dashed curve of Fig. 1, the distortion will be reduced. Again, in the frequency domain, this corresponds to the fact that symmetrical echo pairs change the amplitude response without affecting the delay response.

The circuit could be realized in the manner of the conventional time domain, or 'echo' equalizer of Fig. 6(a). Note that the leading echo is obtained by delaying the main signal and taking the leading echo from the undelayed input.

In the time domain equalizer, an independent control is needed for each echo but sometimes it is more convenient to take the echoes in symmetrical pairs. Thus, firstly, the two echoes of one pair are added to each other in the same sense. When this echo pair signal is added to or subtracted from the main signal, the delay times, T, of one line duration, that is, 64 μs, with an accuracy of the order of one picture element, that is, 0·1 μs. However, delays of this order, with flat delay and amplitude responses over the whole useful bandwidth, have been expensive to produce (up to the present anyway) so that vertical aperture equalization, however desirable, has usually been considered too expensive. Nevertheless, there have been encouraging signs recently that it may become general in colour cameras.[5, 6]

In the usual application, to horizontal aperture equalization, where the correction times are only of the order of several picture elements, the circuit can be simplified to use only the one delay line with an open circuit termination at the far end from which the main signal is taken. The signal is reflected from the far end, with polarity unchanged, back to the sending end, where the termination is correct and the signal is dissipated. Thus there is present at the sending end an echo pair, consisting of the original signal and its doubly delayed copy, which are spaced symmetrically about the 'main' singly delayed, signal.

The presence of the echo pair waveform, with both parts at the one terminal, simplifies the design of the remainder of the equalizer, which is simply a method of subtracting a controllable amount of the echo pair correction waveform from the main signal. It is also desirable that when the equalization control is varied to correct the short time (high frequency) response, (for example, in a pulse and bar waveform, to correct the shape and height of the pulse), the long time (low frequency) response (for example, the height of the bar) should remain constant.

4.2. *Circuit Realization in the Conventional Method*

In the detailed design of an aperture equalizer, the quality of the delay line is important. It must have a constant delay and low loss across the useful band, so that it does not distort signals, either the 'main' signal that traverses the line once or the reflected signal that, after a double pass through the line, becomes the second signal of the echo pair. At the same time the sending end impedance must terminate the line correctly so that no signals are sent on a third and fourth pass of the line.

It might seem that the simplest and most elegant realization of the delay line would be a length of cable, but in solid polythene cable, a delay of 100 ns would require a 20 m length. Special delay cables are available which provide such delays in a most compact form and they have been used in aperture equalizers. However, they tend to attenuate the high frequencies and they also have high surge impedances which accentuate the reactive mis-match produced by the capacitive input impedances of the amplifiers bridged across the ends.

It would seem that lower losses can be achieved in delay times with lumped constants. These are basically multi-section low-pass filters, where the phase-linearity in the useful band is increased as the cut-off frequency is increased beyond it but this in turn needs more sections (that is, more components) for a given delay. Typically for a 100 ns delay and a 10 MHz cut-off, six constant-k prototype sections would be needed. If suitable mutual coupling can be arranged between inductors to produce m-derived sections with m greater than 1, typically 1·27, phase-linearity can be maintained to higher frequencies for a given cut-off frequency and fewer components used. However, some prototype half-sections will be needed in any case at points where the bridging amplifier lumps capacitance across the line, and an m-derived half-section ($m = 0{\cdot}6$) at the input end to minimize reflexions from the resistive input termination. The order of the component values can be assessed from those of a 75 Ω low pass filter with a 20 MHz cut-off frequency. The capacitance of a full prototype section is 214 pF; the inductance for a full prototype section is 1·2 μH which, in a miniature ferrite pot core, can be realized with a Q of around 100 using six turns of wire.

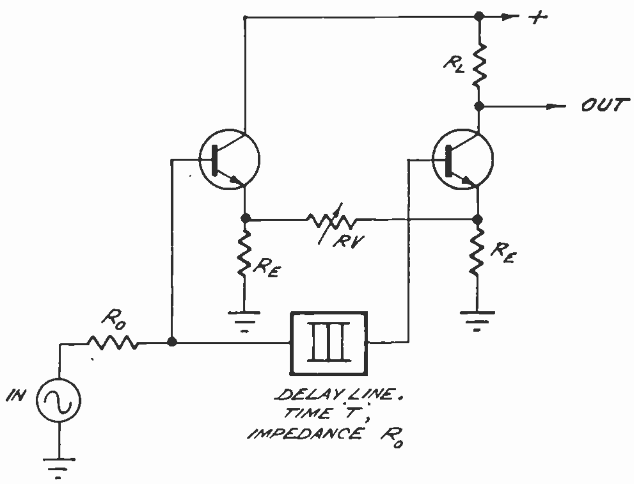


Fig. 7. Conventional aperture equalizer producing the same response as Fig. 6 but using one delay line with far end unterminated.

To realize an amplifier stage in which the two signals are subtracted, the method preferred by the author is to connect the two ends of the delay line to the bases of two transistors whose emitters are coupled via a variable resistor as in Fig. 7. The output is taken from the collector of the transistor which receives signal from the far, unterminated, end of the delay line. When the coupling resistor, RV, is made infinitely large, the only signal reaching the output collector is due to the signal delayed once by the delay line. As the value of RV is decreased, more and more signal is coupled in from the input end of the delay line comprising the echo pair of undelayed and doubly delayed signals.

Since, at low frequencies, the signals at the two emitters are of identical amplitude and phase, no current flows in the coupling resistor, whatever its setting. Hence the low frequency response at the output collector remains constant regardless of control setting. Thus the requirements of the variable subtracting stage are satisfied in a very simple manner.

Several precautions must be observed if precise results are to be obtained.

Firstly, the low-frequency signal voltages are identical at the two emitters only if the g_m, or y_{fe}, of the two transistors, their collector resistances (both internal and in the external load) and the external emitter resistances are all the same. Even if the transistors are identical, the presence of a load, R_L, in one collector, but not in the other, would unbalance the signals to some extent. However, this can be taken up by connecting a small resistor, with the value $R_L/(\mu+1)$ between the emitter of the unloaded transistor and the junction between its emitter resistor and the variable control. This, however, is less important in transistors than in valves, where μ is much lower.

Secondly, it is desirable to use a Darlington pair for each emitter follower so that a high bridging impedance is presented to the delay line by both parts of the subtracting stage. The addition of two extra transistors causes little extra complication or expense.

It is sometimes preferred to connect the two emitters together and to couple the first collector into the load resistor along with the second collector in such a manner that the sum of the two voltage drops, due to each collector current flowing through its portion of the load resistance, is constant. However, as this requires a ganged pair of variable resistors and has no discernible advantage in transistor circuits, it is not considered further here.

4.3. *Frequency Response of the Conventional Method*

It can be shown that the response of the aperture equalizer derived in the manner above is, neglecting constant multipliers which take account of gain and polarity,

$$e_{out}/e_{in} = (1-n\cos\omega T)/(1-n) \qquad \ldots\ldots(12)$$

where T is the delay time of the line and n expresses the proportion of echo pair correction signal added to the 'main' signal. It is usually explained that this response is a 'reasonably good' compensation for a Gaussian loss but it is a matter of common experience that the shape is not really satisfactory. Watching the wedge portion of a test pattern on a good quality picture monitor as the equalization increases, the white sections of the wedge are seen to become whiter at the high frequencies. However, with the conventional method, before the section of the wedge around 400 lines per picture height (corresponding to 5 MHz) is fully corrected, the section around 200 lines per picture height (corresponding to 2·5 MHz) is over accentuated, showing 'whiter than white'. Even if it is objected that such a test attempts to equalize the camera and monitor as a unit, the incorrect shape of the equalization is still demonstrated, since their overall response will be Gaussian also. For the loss due to the finite width of the scanning spot of the monitor will be near Gaussian, similar to that of the camera, and when two Gaussian responses are convoluted (that is, connected in cascade) the result is again Gaussian, the resultant rise-time being the r.m.s. addition of the two component rise-times, or reciprocal cut off frequencies.

In any case, the result can be demonstrated by calculating a typical response from equation (12).

For illustration, an aperture loss of Gaussian shape and 1 neper (8·69 dB) down at the system cut-off frequency of 5 MHz will be taken as typical. Not only does this correspond to a typical case (5 MHz response 37% of the low frequency response), it also simplifies computation, since this 5 MHz amplitude is e^{-1} times the low frequency amplitude. The response of the actual equalizer will be made to match that of the ideal response at the lowest frequencies and then the discrepancy between the ideal and the actual at the cut-off frequency, f_C, (5 MHz) will be found.

To match the response at the lowest frequencies, equation (12) is expanded to

$$e_{out}/e_{in} = 1+\{\omega^2T^2n/2!(1-n)\}- \\ -\{\omega^4T^4[n/4!(1-n)]\}+\ldots \qquad \ldots\ldots(13)$$

and its second term matched to the corresponding term of the inverse Gaussian expression

$$\exp(y^2) = 1+y^2+(y^4/2!)+(y^6/3!)+\ldots \qquad \ldots\ldots(14)$$

and since, in the particular example, y^2 is taken as unity, the second term will be unity when ω corresponds to the cut-off frequency f_C.

If the peak of the response (which occurs when ωT is π) coincides with f_C, then the amplitude at this point will be $1+(4/\pi^2)$, corresponding to 3·0 dB. In other words, if a cosine type equalizer of this kind is made to match accurately, at the lower frequencies, an inverse Gaussian response corresponding to 1Np (8·7 dB) loss at f_C, it will actually produce only 3·0 dB lift at f_C.

The degrees of lift obtainable for other peaking frequencies are set down in Table 3 and three of the curves are plotted in Fig. 2. These show clearly how desirable it is that the peaking frequency should be well above f_C. On the other hand, most equalizers in use today are peaked around this figure, usually between 5 MHz and 7·5 MHz.† Note that the lift can never exceed 2 times (6·0 dB) and in the best feasible case, when the response is peaked at 10 MHz to 15 MHz, the lift obtained will still be 3 dB to 3½ dB short of the 8·7 dB required.

Table 3.

Response of simple cosine equalizers matched at low frequencies to ideal, inverse Gaussian, lift of 1Np (8·7 dB) at 5 MHz

Frequency of peak response (MHz)	5	7·5	10	15	∞
ωT	π	$2\pi/3$	$\pi/2$	$\pi/3$	∞
Lift at peak (dB)	3·0	5·6	8·4	13·3	∞
Lift at 5 MHz (dB)	3·0	4·5	5·2	5·6	6·0

In practice, of course, the picture is not quite as grim as this. A slightly excessive lift *will* be tolerated at lower frequencies to allow a greater lift to be achieved at 5 MHz. Nevertheless, this cannot be carried too far, since much smaller response errors can be tolerated at low frequencies than at high frequencies.[7]

In any case, the losses that need to be compensated are often much greater than 1Np. This is a typical figure for vidicon tubes but in telecine cameras it is desirable also to compensate for the loss of fine detail information (that is, high frequency response) in the film itself. Studies in the A.B.C. Engineering Laboratory with a micro-densitometer have shown that good response can be obtained from some direct camera negatives or reversal films, even a rising response being observed in one case. However, in continuous contact prints, which are of poorer quality but are more quickly and cheaply produced and are therefore most often used, modulation depths between 80% and 15% (that is, between 2 dB and 16 dB down) have been observed at 28 lines per mm, corresponding in 16 mm film to 390 television lines per picture height, that is, 5 MHz in the Australian system.‡ This applies of course to a single copying process. Later-generation copies are correspondingly worse.

Again there is a considerable loss due to the finite size of the scanning spot in the monitor picture tube. One hesitates to suggest pre-compensation at the transmitting end for losses in the receiver, particularly in the absence of information on the minimum loss to be expected there. Nevertheless, it is obvious that at the very least, the studio should aim at producing a signal as free from faults as possible.

It is not sufficient, as has been pointed out before, to demonstrate that the insertion of a given fault produces no visible effect on a monitor picture. There is a 'threshold of visual impairment'[7] which makes a single fault with a K factor below 3% or even 5%, practically invisible but when several such errors occur in tandem, the impairment can be severe. For these reasons it is desirable that the aperture compensation at the studio should be as accurate as possible.

The kinds of aperture equalizer generally used up to the present lie between 'cosine (peaked at 5 MHz)' and 'cosine (peaked at 7·5 MHz)' in Fig. 2. They produce a poor fit to the ideal (dashed) curve. However, it will be shown, as the paper proceeds, that other kinds can be devised which fit much more closely.

4.4. *An Improvement on the Conventional Method*

The response above can be improved in a remarkably simple manner. Just as in an echo equalizer a more refined kind of equalization is obtained when additional echo pairs (separated further from the main pulse) are added, so in an aperture equalizer the response can be refined further by adding an extra delay line and providing amplifier stages (as in Fig. 8) which take the 'main' signal from the far, unterminated and (delayed $2T$) and feed it to the output while subtracting the echo pair (delayed T and $3T$) from the mid-point and *adding* the echo pair (delayed zero and $4T$) from the input end. All that is required is

† The reason often given is that this minimizes noise outside the useful band but it is seen in Fig. 2 that this is true only in the crudest way. What is really needed is an equalizer which gives the required inverse Gaussian response regardless (more or less) of its out-of-hand response, cascaded with a fixed linear-phase low-pass filter which attenuates the signal sharply beyond the band-limit, f_C, and it is virtually impossible to produce such an overall response in a single device. Limitation of the pass band occurs anyway in the transmitter (and in the receiver and links, videotape machines and so on), but it is necessary also at an early stage in the studio equipment to prevent out-of-band noise components beating with each other, due to unavoidable non-linearities, and producing additional in-band noise components. For this reason, any equipment where non-linearity is produced deliberately, in particular a gamma control, should be preceded by a band-limiting filter.

‡ In television measurements, using a pattern of lines, a black line followed by a white line is counted as two lines, for example, 390 lines per picture height is equivalent to 520 lines per actual picture width of 52 μs and hence to 10 lines per μs. In photographic measurements, on the other hand, only the black lines (or alternatively, only the white lines) are counted. Thus 260 (½ × 520) lines per active picture width of 9·4 mm is equivalent to 28 lines per mm.

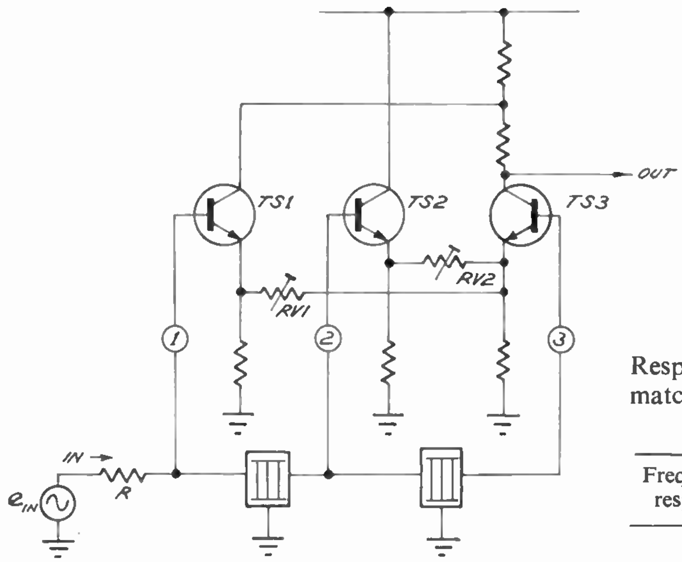


Fig. 8. Improved aperture equalizer with two delay lines, each of time, T, and surge impedance, R_0.

Table 4.

Response of cosine equalizers with two coefficients, matched at low frequencies to an ideal, inverse Gaussian lift of 1Np (8·7 dB) at f_c, 5 MHz

Frequency of peak response (MHz)	5	7·5	10	15	∞
ωT	π	$2\pi/3$	$\pi/2$	$\pi/3$	∞
Lift at peak (dB)	4·2	8·4	13·0	22·0	∞
Lift at 5 MHz (dB)	4·2	6·4	7·1	7·6	8·0

(i) a further delay line of time, T, capable of providing the required delay with negligible attenuation,

(ii) an extra transistor adding stage and

(iii) a method of controlling the proportions of the *two* correcting signals.

In fact, the first two requirements are readily achieved. The third requirement is complicated by the need that, when the equalization is varied, the two-echo pair signals must contribute in proportions which vary with respect to each other and it might seem at first that a considerable problem must be faced in ganging two controls. However, further consideration shows that while the proportions needed for Gaussian response must be known, to establish the design parameters, it is preferable to keep the two controls independent so that any deviation from the exact Gaussian shape (for example, towards sin x/x shape) may be compensated more exactly.

The fit of the equalizer to the inverse Gaussian loss curve is now much better than in the previous case. To fit the expression $\exp(k\omega^2T^2)$ at the lower frequencies, the equalizer response is

$$e_{out}/e_{in} = [1+(5k/2)+3k^2]-[(8k/3)+4k^2]\cos\omega T + \\ +[(k/6)+k^2]\cos 2\omega T \quad \ldots\ldots(15)$$

Then the properties of such equalizers, simulating the typical 1Np Gaussian lift at 5 MHz, as the peak frequency is moved from 5 MHz upwards, is shown in Table 4.

Note that the maximum possible response at f_c is now 2·5 times (8·0 dB). Again, if the response is peaked near 5 MHz the fit is comparatively poor but, if it is peaked in the 10 MHz to 15 MHz region the response is only 1 dB to 1·5 dB below the ideal. And again since these figures are based on a maximally-flat fit at low frequencies, an even better result can be obtained if a small amount of ripple is tolerated, as will happen when the two controls are adjusted independently on an experimental basis.

It is quite feasible to go on from here to a three-section delay line and thus achieve a good fit up to greater magnitudes of equalization. The computations have been done and the design is straightforward but it seems that the added expense and complexity of control is not justified for present applications, remembering that the feasible maximum of equalization is limited in any case by the increase in noise, as demonstrated in Table 10, Section 8.

It will be obvious that an improved *vertical* aperture equalizer can be realized in a similar manner, using four delay lines, each of 64 μs, instead of the conventional two.

5. Simulation of the Delay Line with a Linear-Phase Low-Pass Filter

5.1. *Using a Third-order Filter*

A conventional delay line is fairly cumbersome to construct and adjust, particularly with the low losses needed to ensure that the 'trailing' echo is sufficiently similar to the 'leading' echo. Besides, there is always the problem of internal reflexions from discontinuities within the line. Thus the question arises whether one can be simulated in a simpler manner, and in fact it can be, by a low-pass filter that has a linear-phase response overall when terminated in resistance at only one end. The network of Fig. 9 has a response at its far unloaded end, which has for its denominator the third-order Bessel polynomial

$$e_3/e_{in} = 1/[1+pT+(2/5)p^2T^2+(1/15)p^3T^3] \quad \ldots\ldots(16)$$

and is thus the linear-phase, or maximally flat delay, response of third-order.

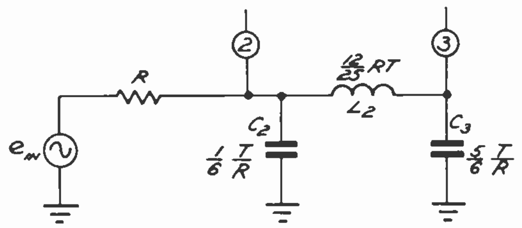


Fig. 9. Third-order linear phase low-pass filter unloaded at far end, simulating delay line of time, T, in Fig. 8.

The response at the input end, that is, across C_2 is

$$e_2/e_{in} = [1+(2/5)p^2T^2]/[1+pT+(2/5)p^2T^2+ +(1/15)p^3T^3] \quad \ldots\ldots(17)$$

Note that the numerator of equation (17) contains only terms which are of even order in pT, a general property of such filters terminated in resistance at only one end, so long as the dissipation in the reactive components is negligible. Hence the delay response of e_2 is the same as that of e_3 and so is that of any voltage derived by adding or subtracting them in any numerical proportion. For if they are added in the proportion $1:n$, the resulting output voltage is given by

$$e_{out}/e_{in} = \frac{1-(2/5)[n/(1-n)]p^2T^2}{1+pT+(2/5)p^2T^2+(1/15)p^3T^3} \quad \ldots\ldots(18)$$

The general form of this kind of response is the same as that of Thomson's sine-squared filter,[8, 9] except that where Thomson used numerators with factors of the form $1-\omega^2T^2$ to place notches in the response and thereby effectively terminate it at frequencies within the linear phase limit of the denominator, the factors in the present application are of the form $1+\omega^2T_r^2$.

Thus, in the response of equation (16), the amplitude initially *rises* with frequency, while the delay response is flat up to the frequency which we will call the 'limit of phase-linearity' beyond which the delay response drops (that is, deteriorates) rapidly. It is sufficient that this limit be at 5 MHz or above. In the third-order filter, the limit occurs when ω^2T^2 is about 2, that is, T must be 45 ns or less. The amplitude continues to rise somewhat beyond this point and then, after passing through a broad maximum, falls with an ultimate slope of 6 dB per octave. This response is similar to that of the 'cosine' type derived from the analogous delay line and fits the Gaussian curve as well as the comparable device using a delay line.

$$|e_{out}/e_{in}|^2 = [1+\omega^2T^2(0{\cdot}1+0{\cdot}5k)]^2/|E(j\omega T)_{T_3}|^2 \quad \ldots\ldots(19)$$

Table 5.

Response of equalizers with third-order low-pass filters for which equation (19) is true, matched lower frequencies to an ideal inverse Gaussian lift of 1Np (8·7 dB) at f_c (5 MHz)

Frequency of peak response (MHz)	9	13	20	∞
k	1	2	4	∞
Lift at peak (dB)	7	12	17	∞
Lift at 5 MHz (dB)	5·0	5·5	5·8	6·0

This can be seen when Table 5† is compared with Table 3. The numerical values of k in Table 5 were chosen in steps of 2 times, so that the frequencies for 1Np lift in the ideal matching curves are half an octave apart, that is, are separated by ratios of 2 in ω^2T^2. Again, as in the case of the simple cosine equalizer, the amplitude of the highest lift that can be obtained, corresponding to an ideal 1Np lift, is 2 times, that is, 6·0 dB.

The equalizer is realized by connecting points 2 and 3 in Fig. 9 to the corresponding points in Fig. 7 in place of the delay line.

An apparent difficulty with this circuit turns out in fact to be an advantage. Because the 'main' signal is taken from the output of a low pass filter with a gentle Gaussian slope, the 'flat' position of the equalizer is achieved only by ensuring that the 'echo pair' signal, as we called it when a true delay line was used, is still present to some extent. In fact, in equation (19), if the minimum value of k is zero, when n is 0·2, then the response is 0·1 dB down at 5 MHz and 0·3 dB down at the limit of phase linearity but in fact this non-zero value for n is an advantage when the control is a variable resistor RV2 coupling the two emitters, for it is much easier to provide a finite maximum resistance at the end of its range than to provide a complete open circuit.

† Equation (19) in Table 5 is another version of equation (18). Firstly, it is in the form of the modulus squared for purposes of computation. Hence the denominator is converted from the operational 'pT' form of equation (16) to the steady state '$j\omega T$' form and its squared modulus taken. Secondly, the numerator is rewritten in terms of k, which is equivalent to $(n-0{\cdot}2)/(1-n)$, and is the coefficient in the equivalent ideal response, exp $(k\omega^2T^2)$.

5.2. *Using a Fifth-order Filter*

With the single delay line simplified in this way and remembering the improved performance obtained from a two-section delay line, it is a small step to substitute a fifth-order linear-phase filter as shown in Fig. 10 for the two delay lines in Fig. 8.

Thus the denominator is made the fifth-order Bessel polynomial

$$E(p)_{T_5} = 1+pT+(4/9)p^2T^2+(p^3T^3/9)+ +(p^4T^4/63)+(p^5T^5/945) \quad ......(20)$$

Then the voltages at the different tapping points are described by

$$e_1/e_{in} = [1+(4/9)p^2T^2+(p^4T^4/63)]/E(p)_{T_5} \quad ......(21)$$

$$e_2/e_{in} = [1+(26/99)p^2T^2]/E(p)_{T_5} \quad ......(22)$$

$$e_3/e_{in} = 1/E(p)_{T_5} \quad ......(23)$$

As before, the numerators contain only even order powers of pT and thus, when e_1, e_2 and e_3 are added together in any proportion, the delay response will always be that of the fifth-order Bessel polynomial which retains its phase linearity up to the limit defined arbitrarily by the group delay error reaching 0·02 T, that is, the frequency where ω^2T^2 is 9.

The required response is obtained by adding the three voltages, e_1, e_2 and e_3, in suitable proportions, n_1, n_2 and n_3, so that their sum approximates the required inverse Gaussian lift. Thus if

$$e_{out} = n_1e_1+n_2e_2+n_3e_3 \quad ......(24)$$

then

$$e_{out}/e_{in} = \{n_1[1-(4/9)\omega^2T^2+(\omega^4T^4/63)]+ +n_2[1-(26/99)\omega^2T^2]+n_3\}/E(j\omega T)_{T_5} \quad ......(25)$$

where

$$E(j\,\omega T)_{T_5} = 1+(j\,\omega T)+(4/9)(j\,\omega T)^2+(1/9)(j\,\omega T^3)+ +(1/63)(j\,\omega T)^4+(1/945)(j\,\omega T)^5 \quad ......(26)$$

that is, the fifth-order Bessel polynomial used by Thomson to produce low-pass responses with maximally-flat delay.

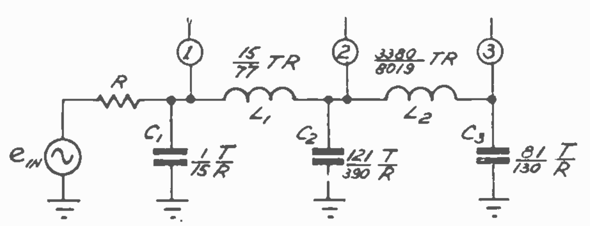


Fig. 10. Fifth-order linear phase low-pass filter unloaded at far end, simulating the pair of delay lines in Fig. 8.

The conditions for matching its response to the first two terms of the ideal response

$$|e_{out}/e_{in}|^2 = \exp(k\omega^2T^2) = 1+k\omega^2T^2+(k^2\omega^4T^4/2!)+ +(k^3\omega^6T^6/3!)+\ldots \quad ......(27)$$

are fulfilled by the expressions

$$n_1 = (13+182k+819k^2)/104 \quad ......(28)$$

$$n_2 = -(44+506k+1386k^2)/104 \quad ......(29)$$

$$n_3 = (135+324k+567k^2)/104 \quad ......(30)$$

the resulting response being

$$e_{out}/e_{in} = \{1+\omega^2T^2[(1/18)+(k/2)]+ +\omega^4T^4[(1/504)+(k/36)+ +(k^2/8)]\}/E(j\,\omega T)_{T_5} \quad ......(31)$$

For the purpose of realizing the circuit of Fig. 12, the coefficients, n_1, n_2 and n_3, are separated into component parts as shown in Table 6.

Table 6.

Components of n_1, n_2 and n_3 (multiplied by 104) for realization of Fig. 12

n_1	n_2	n_3
+52		+52
$+22+182k+819k^2$	$-22-182k-819k^2$	
	$-22-324k-567k^2$	$+22+324k+567k^2$
-61		$+61$

Note that in equations (28) and (30) the coefficients, n_1 and n_3, are always positive and in equation (29) the coefficient, n_2, is always negative.

Now consider two gain stages that have equal emitter (or cathode) resistors and have a different input into each, say e_2 and e_3, the output being taken from the collector of the stage fed by e_3. Then, when a coupling resistor is connected between the two emitters, an increment of gain is added to the e_3 stage whose ratio to the original gain is equal to the ratio of the emitter resistor to the coupling resistor, while the signal from e_2 is coupled to the e_3 output with a gain in the e_3 stage. As explained in Section 4.2, this is an approximation which is valid for transistors but less so for valves.

Thus the coefficients required by Table 6 are realized:

(i) By adding signals from the collectors of the stages with inputs fed from e_1 and e_3 in a common load resistor. Since the emitter resistors, R_E, are equal, the contribution from each will be equal, that is, each contributes 52/104 of the total.

(ii) By taking the terms containing k and k^2 in n_1 and n_3, in each case opposite their negatives in n_2. The constant part of n_2 is then split equally (11/104 each) and allotted suitably. Using the principle enumerated above, the coupling resistor between emitter 1 and emitter 2 will then be

$$52R_E/(22+182k+819k^2)$$

Similarly the coupling resistor between emitter 2 and emitter 3 will be

$$52R_E/(22+324k+567k^2).$$

In this way, the coupling resistors are made equal at their maximum value and in fact they track reasonably well throughout their useful range in Table 7. Note that the equal and opposite pairs of contributions, say in n_2 and n_2, arise from collector 1 being connected to the load while collector 2 is not.

(iii) By making up the remaining $\pm 61/104$ by connecting a resistor 52/61 times R_E between emitter 1 and emitter 3. But since a *negative* contribution is required to n_1, the coupling resistor must be connected to the emitter of the stage whose base is connected to e_1 but whose collector is *not* connected to the output load, that is, a separate emitter follower is required, taking signal from e_1.

In Table 7 the values of Table 6 are expressed numerically for specific values of k which were chosen for steps of 2 in the corresponding ideal responses. Thus when plotted in Fig. 11, where the abscissa is scaled in ωT *squared*, the ideal curves are each separated by *half* an octave. From the scale of group delay error it appears that the limit of phase linearity occurs when $\omega^2T^2/9$ is unity, that is, when ω is $3/T$, and that in the 'flat' position when k is zero, the response is only 0·4 dB down at this limit. Thus if a high degree of aperture lift were required, the limit of phase linearity could be set at 5 MHz.

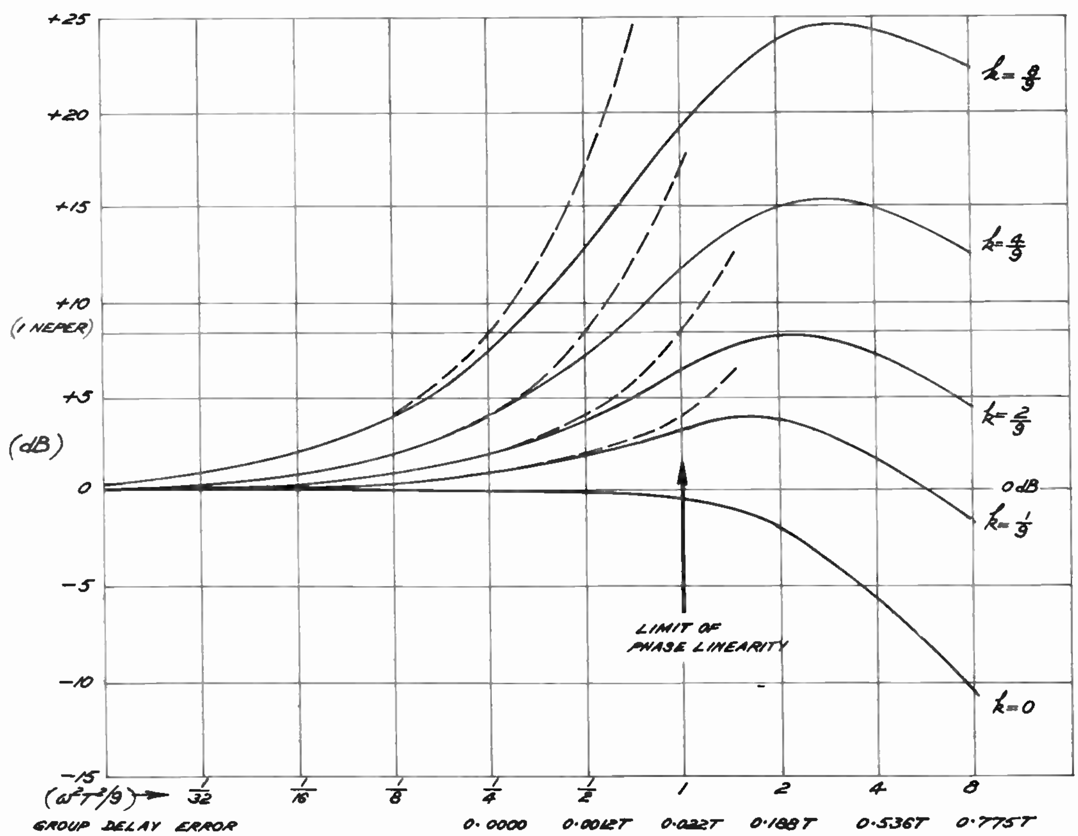


Fig. 11. Family of responses achieved with improved equalizer using principles of Figs. 8 and 10 and realized in Fig. 12. Solid curves: actual responses. Dashed curves: ideal responses, matched to actual responses at low frequencies.

Table 7.

Numerical values for couplings (and hence normalized coupling conductances) in aperture equalizer with fifth-order linear-phase low-pass filter

	n_1	n_2	n_3
Basic to all	+0·50		+0·50
	−0·59		+0·59
When $k = 0$	+0·21	−0·21	
		−0·21	+0·21
	+0·12	−0·42	+1·30
When $k = 1/9$	+0·51	−0·51	
		−0·62	+0·62
	+0·42	−1·13	+1·71
When $k = 2/9$	+0·99	−0·99	
		−1·17	+1·17
	+0·90	−2·16	+2·26
When $k = 4/9$	+2·54	−2·54	
		−2·67	+2·67
	+2·45	−5·21	+3·76
When $k = 8/9$	+7·99	−7·99	
		−7·29	+7·29
	+7·90	−15·28	+8·38

However, in setting up the practical circuit of Fig. 12, the limit was set half an octave higher, at 7·1 MHz, to ensure a better curve shape inside the useful band and flatness in the zero position to better than 0·1 dB. Then T, which is the insertion delay (identical for group delay *and* phase delay) of the filter and also determines the component values of the low-pass filter in Fig. 10, is 67·5 ns. It is interesting to compare this value with the total delay, $2T$, of 66·7 ns in the two-section delay line, peaked at 15 MHz in the two coefficient cosine type, which produces a very similar response (compare Tables 4 and 8). In fact the present equalizer combines the greatly improved response of the two-delay line equalizer with the simplicity of a low-pass filter, as may be seen from Fig. 11.

$$|e_{\text{out}}/e_{\text{in}}|^2 = \{1+\omega^2T^2[(1/18)+(k/2)]+ +\omega^4T^4[(1/504)+(k/36)+(k^2/8)]\}^2/|E(\mathrm{j}\,\omega T)_{T_5}|^2$$

Table 8.

Response of equalizers with fifth-order low-pass filter for which the above equation is true, matched at the lower frequencies to an ideal inverse Gaussian lift of 1Np (8·7 dB) at f_C (5 MHz)

Frequency of peak response (MHz)	7	11	17	∞
k	2/9	4/9	8/9	∞
Lift at peak (dB)	8	15	24	∞
Lift at 5 MHz (dB)	6·5	7·4	7·7	8·0

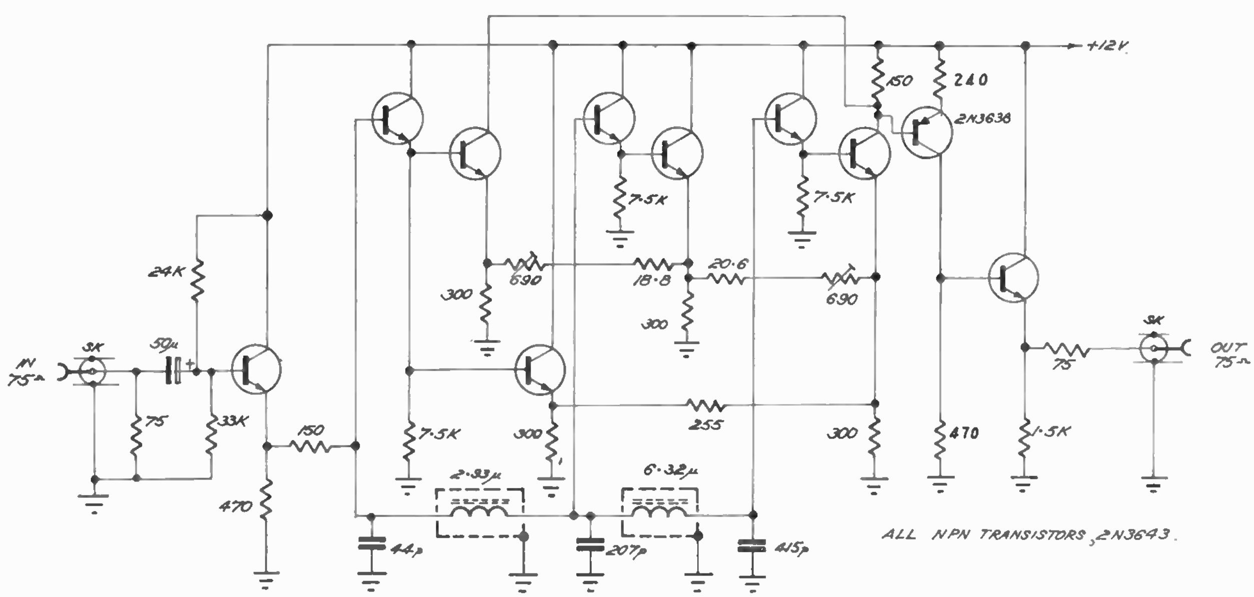


Fig. 12. Realization of improved aperture equalizer. Responses as Fig. 11.

Although the component values of the two coupling resistors track reasonably well for the various values of k, the two controls were made independent, firstly because the calculated tracking is not perfect, secondly because of the difficulties in tracking practical controls over the wide range of resistances and thirdly to allow greater flexibility in matching curves slightly different from Gaussian.

Table 9.

Comparison of responses of aperture equalizers using fifth-order l.p. filters ($k = 8/9$) matched at low frequencies to Gaussian and sin x/x responses

$4\omega^2T^2/9$	1/8	1/4	1/2	1	2	4	8	16
Matched to:								
Gaussian (dB)	1·1	2·1	4·2	7·7	13·0	19·1	23·6	24·1
Ideal Gaussian (dB)	1·1	2·2	4·3	8·7	17·4	34·7		
Matched to:								
sin x/x	1·1	2·3	4·6	8·9	15·1	22·1	27·1	28·1
Ideal sin x/x (dB)	1·1	2·3	4·9	11·7	23·8			

↑ Notch between, when $4\omega^2T^2/9 = \pi^2/6$

Table 9 shows, for example, how the response can be changed from one that matches a Gaussian response, actually the $k = 8/9$ response in Table 8, to one that matches a sin x/x response similar to the Gaussian at low frequencies. It also illustrates, by the way, how the curve originally intended to fit the sin x/x response (fourth row) actually fits the ideal Gaussian response (third row) much better than the original Gaussian matching curve (second row), *so long as* a small over-compensation, up to 0·3 dB, can be tolerated at the lower frequencies.

In Fig. 12, the Darlington connected transistors, with emitter followers feeding emitter followers, serve two purposes, firstly to prevent the transistor inputs loading the low-pass filter, which ideally must have no dissipation in its reactive elements, and secondly to prevent the impedance offered by the filter to the bases of the transistors affecting the impedance seen at their emitters and hence the proportions in which their signals add.

Since the shunt capacitances of the amplifier input form part of the filter, it must be aligned before use. Firstly, all the component values are bridged as accurately as possible after making allowance for the anticipated values of transistor input capacitance. The emitter coupling resistors, including the fixed one, are temporarily disconnected and readings taken of frequency response at their emitters. From equations (21), (22) and (23), it will be seen that while the frequency response of e_3 is a slowly falling Gaussian approximation, the response of e_2 has a notch when when ω^2T^2 is 99/26 (that is, 4·69 MHz when T is 67·5 ns). Again e_1 has two notches which occur when

$$(\omega^4T^4/63-(4\omega^2T^2/9)+1 = 0 \quad \ldots\ldots(32)$$

that is, 3·70 MHz and 11·91 MHz when T is 67·5 ns. Since the capacitances are comparatively large, these notch frequencies can be used to set the values of the inductors. However, the smallest, and therefore most vulnerable, capacitance is C_1. Fortunately any shunt *resistance* here can be taken up as part of the input resistance. However, the accuracy of its capacitance is important. It can be checked by temporarily disconnecting the 150 Ω feed resistor from C1 and connecting it to C3. The voltage at C3 will have two notches determined by

$$(26\omega^4T^4/15309)-(877\omega^2T^2/5103)+1 = 0 \quad \ldots\ldots(33)$$

that is, 5·87 MHz and 22·97 MHz when T is 67·5 ns. Alternatively, if measurements cannot be taken conveniently above 15 MHz, C2 may be temporarily disconnected also. Then a single notch will be seen at C3 when ω^2T^2 is (15309/629), that is, at 11·63 MHz when T is 67·5 ns.

When the equalizer was built, the frequency response was as predicted. The linearity of its phase response was checked by the symmetry of its waveform output with a $2T$ pulse-and-bar input signal and found to be satisfactory also.

6. Use of a Delay Line and Low-pass Filter

Another kind of aperture equalizer intended for the special case of 'flare' equalization, that is, one providing a linear-phase 'shelf'-type lift at comparatively low frequencies, has been described before[2] but is included here for completeness, since it is usable generally in the case where the delay line produces significant attenuation in the useful band, thus making the amplitudes of the signals in the 'echo pair' unequal and also degrading the 'main' signal. This is done as in Fig. 13 by passing the signal through a linear-phase low-pass filter and also through a delay network of high transmission quality, for example, a 489 m length of UR57 cable with an active equalizer in the case quoted, where T was 2·5 μs, the delay time being equal to the nominal delay of the low-pass filter, that is, the time at which its transition reaches 50% amplitude. The filtered signal is then subtracted from the delayed signal. The explanation given in the original paper in terms of the bar transition does not seem to have satisfied some readers and the following explanation in terms of the pulse response may be more satisfactory.

The distorted pulse, with its base smeared symmetrically as in Fig. 13(a) can be thought of as the sum of an undistorted pulse with another pulse which has been distorted by passing it through a linear-phase low-pass filter whence its height is reduced and its width increased in the same proportion. Assuming that the energy in the distortion pulse component is small, the distortion is corrected by (low-pass) filtering the signal to produce a signal similar to the distortion component and subtracting that from the main signal.

Alternatively, the (low-pass) filtered, and thus broadened, pulse may be thought of as comprising three pulses side by side, with a shortened version of the main pulse in the middle flanked by a shorter pair of pulses, the 'echo' pair. When this is subtracted from the main signal, the result is the same as with the earlier types of delay line equalizer, except that a copy of the undistorted pulse is subtracted from it as well, causing a loss of height that must be made good by increased gain. Again, if the subtracting amplifier is of the type described earlier with the signals applied at the same low frequency amplitude to the bases of two identical transistors and with a variable coupling resistance between their emitters, then the amplitude of the resulting output signal will again be constant at low frequencies.

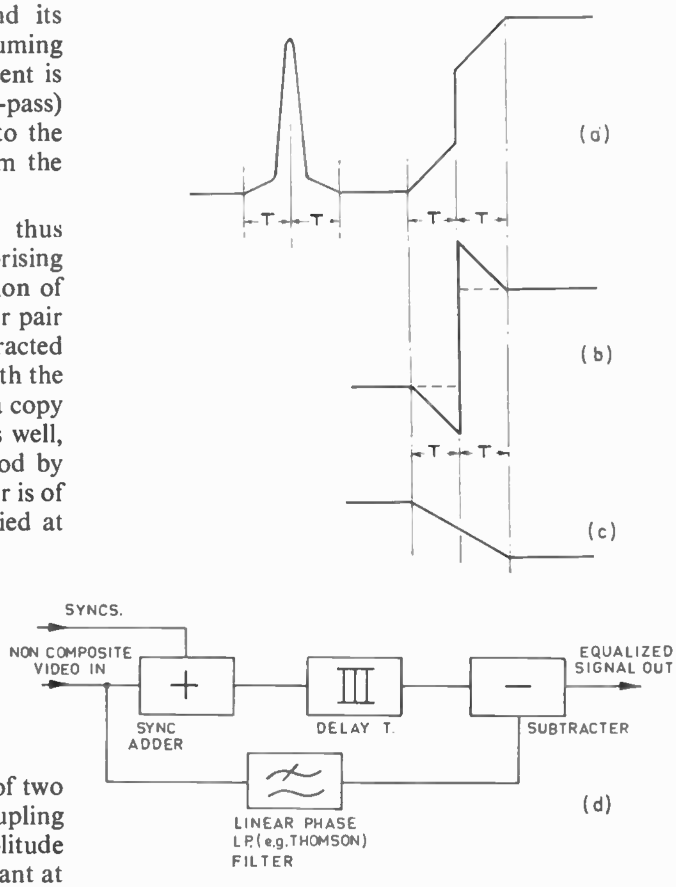


Fig. 13. Equalizer for 'flare'.

The only practical difficulty in using such an equalizer with the long delays suggested in reference 2 is that the synchronizing information must be removed and reinserted in an undistorted form. Otherwise the waveform distortion that corrects the picture information may distort the sync. pulses so badly that they cannot perform their synchronizing function effectively. This applies to some degree in all aperture equalizers but is less noticeable in the conventional short-time kind, since most are incorporated in equipment where the sync. pulses are reinserted later anyway and, even if they were not, most sync. separators can cope with short time distortion of the sync. information.

7. Testing Methods

Since the purpose of an aperture equalizer is to compensate the effect of errors in the scanning process, the best kind of input signal for testing it is a picture, that is, an optical input, and Seyler's pulse-and-bar pattern[10] is well suited to the purpose. Its rectangular $2T$ pulse seems the most suitable if the test slide is to be reproduced photographically in high-contrast material with a minimum of critical control, even though a sine-squared pulse would be preferred on other grounds.

Mr. J. M. Albiston† has pointed out that, while the 'optical input' waveform is excellent for its original purpose, namely checking the performance of vidicons, it is not so suitable for testing image orthicon cameras because their redistribution effects produce 'edging' errors in the output signal, ('i.o. halo') which vary with exposure. The author agrees but also feels that, if a testing procedure shows up the faults, so much the worse for the equipment tested. If the edging effects of an i.o. camera are considered desirable and the author considers that they are, as a crude, ill-controlled form of equalization of flare[2] in the rest of the television system, especially the receiving picture tube, then it might be feasible to read the electrical output of an i.o. camera through an electrical filter producing a 'standard' flare, for example, a circuit similar to Fig. 13 but with the subtractor replaced by an adder.

Again the pulse and bar electrical waveform is a very useful method of testing video responses. However, with an aperture equalizer, its main use is checking for symmetry of the downward-going 'ears', produced fore and aft of the pulse waveform, this indicating a linear phase response. To check the shape, a

† Convention discussion.

specially drawn template would be needed for each amplitude of lift and this would detract from one of the chief advantages of the pulse and bar waveform, that its wave shape is simple. In any case, aperture equalization tends to accentuate any short time faults, such as ringing after the pulse, in the testing waveform itself. Thus electrical tests of an aperture equalizer are best taken with an electrically simulated aperture *loss* in cascade. Then when the equalizer is correctly adjusted, the pulse and bar waveform are undistorted. In general any equalizer is tested most satisfactorily in tandem with the equipment it is intended to complement, for example, a cable equalizer in cascade with its cable.

It is fortunate that a filter of suitable response is available. The T Thomson filter is 6·0 dB down at 5 MHz and has a response below this that simulates the Gaussian response closely. For this reason, it can be used to good effect in checking the response of variable aperture equalizers when their intended correction is 6 dB at 5 MHz. However, this is only a limited application. Ideally one would like a sequence of 'aperture loss' filters that could be cascaded in, say the familiar 1, 2, 2, 5 sequence of dB loss.

However, the Thomson filter is not a constant resistance device and hence cannot be used with complete certainty, except between purely resistive terminations. Hence, for example, Thomson filters cannot be cascaded with each other. At the moment the use of isolating amplifiers between filters seems too cumbersome. It is hoped to report further work at a later time.

The Thomson filter might also be used in reading the output of Seyler's optical (rectangular) pulse and bar. Consider the case where an ideal $2T$ rectangular pulse is passed through an ideal T sine-squared filter. The output is derived by considering the positive-going $2T$ rectangular pulse as a positive-going step transition followed, at a time $2T$ later, by a negative-going step transition.

The output of the T Thomson filter due to the positive-going step transition is a positive-going waveform with a shape approximating the first half of a $2T$ sine-squared pulse. Actually, it consists of a ramp function going linearly from time zero to time $2T$ with a negative sinewave of period $1/2T$ superimposed. However, it reaches half amplitude at time T and just reaches its final value, the same as the original step height, at time $2T$. At this instant the negative-going step follows in the input $2T$ rectangular pulse waveform. Thus the output waveform returns to zero, its trailing edge symmetrical with its leading edge. Although the resulting pulse is actually a triangular pulse of $4T$ base width, with two cycles of $2T$ period sinewave subtracted, it approximates the sinesquared shape closely. Further, its half amplitude duration of $2T$ and its height are identical with the original $2T$ rectangular pulse.

In a linear system, the order in which the various filters are connected is unimportant. Thus in a linear system, it would not matter whether the $2T$ rectangular, optical, pulse was cascaded with a T sine-squared optical filter before it reached the camera or whether, on the other hand, the $2T$ rectangular optical pulse was fed directly to the camera, while the T sine-squared filter was connected, as an electrical device, as the last link in the chain before the measuring oscilloscope. The latter would provide a very simple procedure for testing the complete optical and electrical transfer of a television camera by a sine-squared pulse and bar method, since the devices it requires are readily available.

In fact, of course, the system is *not* linear; the overall transfer aimed at has a gamma of $\frac{1}{2}$, that is, it can be described by the expression

$$v_{out} = V_o(b_{in}/B_o)^{\frac{1}{2}} + V_{os} \qquad ......(34)$$

where v_{out} is the output voltage, b_{in} is the brightness of the incoming scene, V_o and B_o are quantities of voltage and brightness which set the respective scales and V_{os} is an 'off-setting' voltage which takes account of the fact that when the input brightness is zero, the output voltage is not zero.† In fact the signal goes through successive stages of non-linear transfer ('gamma') and frequency-response ('aperture' loss and correction, together with band-limiting).

Thus, while a testing method using an optical $2T$ rectangular pulse (and bar) as an input and reading the electrical output on an oscilloscope through a T Thomson filter, has obvious attractions, its theoretical limitations must be appreciated. The present intention is to investigate whether the method is valid with the degrees of non-linearity met in practice. In any case, the limitations do not apply to cameras using plumbicons. These have an intrinsic gamma of unity and thus should present no problems due to non-linearity so long as readings are taken with the electrical gamma control set to unity.

8. Effect of Aperture Equalization on Noise

Since aperture equalization increases the amplitudes of the higher frequency components relative to the low frequency components, it also increases the relative amplitude of noise. The extent to which the noise is

† In fact, if the system is arranged so that a 40 : 1 range of input brightness takes the output voltage from 100 IRE units at peak white to 10 IRE units ('set up') at minimum brightness (1/40 of peak white), then zero input brightness (in an ideal system with a gamma of $\frac{1}{2}$) would produce an output signal of $-6{\cdot}9$ IRE units. With no set-up and a 40 : 1 brightness range producing signals between 100 IRE units and 0 IRE units, zero input brightness would produce $-18{\cdot}8$ IRE units output!

increased can be computed by a straightforward exercise in integration.†

Since the bandwidth of a television signal is limited at various points in the system to a cut-off frequency, f_C (5 MHz), the mathematical functions describing the noise must be taken as definite integrals between the limits 0 and f_C. The effect then is as if the filtering were ideal, that is, the gain being constant at frequencies between 0 and f_C and zero above f_C.

Thus the 'noise bandwidth' of white noise is

$$f_{\mathrm{BN(W)}} = \int_0^{f_C} 1\,\mathrm{d}f = f_C \qquad \ldots\ldots(35)$$

In the case of 'triangular', or blue, noise on the other hand, whose energy per cycle rises 6 dB with every octave, the noise bandwidth must be given in terms of a 'transition' frequency, f_T, virtually a scaling factor with the dimension of frequency. Thus

$$f_{\mathrm{BN(B)}} = \int_0^{f_C} (f^2/f_T^2)\mathrm{d}f = f_C^3/3f_T^2 \qquad \ldots\ldots(36)$$

Thus when white noise is processed by an aperture equalizer in which

$$|e_{\mathrm{out}}/e_{\mathrm{in}}|^2 = \exp(f^2/f_E^2) \qquad \ldots\ldots(37)$$

f_E being the 'equalization' frequency, that is, the frequency at which the equalization lift is $\frac{1}{2}$Np (4·34 dB), then the bandwidth of the resulting white, equalized, noise is

$$\begin{aligned} f_{\mathrm{BN(WE)}} &= \int_0^{f_C} \exp(f^2/f_E^2)\,\mathrm{d}_j \\ &= f_C\{1+[a^2/3]+[a^4/(2!5)]+[a^6/(3!7)]+ \\ &\qquad +[a^8/(4!9)]+\ldots\} \qquad \ldots\ldots(38) \end{aligned}$$

where a is the ratio (f_C/f_E) which varies with equalization and is 1 for $\frac{1}{2}$Np (4·3 dB) lift at f, 2 for 1Np (8·7 dB) lift at f, 3 for 1·5Np (13·0 dB) lift at f and so on.

† Though in the absence of a more elegant method, functions involving $\exp(f^2/f_E^2)$ must first be expanded as a series, for example,

$$\exp(f^2/f_E^2) = 1+(f^2/f_E^2)+(1/2!)(f^4/f_E^4)+(1/3!)(f^6f^6)+\ldots$$

and then integrated term by term.

Similarly, in the case of weighted noise, the expansion is

$$\frac{\exp(f^2/f_E^2)}{1+(f^2/f_W^2)} = \frac{1}{1+(f^2/f_W^2)} + \frac{(f^2/f_E^2)}{1+(f^2/f_W^2)} + \frac{(f^4/f_E^4)}{2![1+(f^2/f_W^2)]} + \\ + \frac{(f^2/f_E^6)}{3![1+(f^2/f_W^2)]} + \ldots$$

Also

$$\int \frac{\mathrm{d}_j}{1+(f^2/f_W^2)} = f_W \tan^{-1}(f/f_W)$$

$$\int \frac{f^2\mathrm{d}f}{1+(f^2/f_W^2)} = ff_W^2 - f_W^3 \tan^{-1}(f/f_W)$$

$$\int \frac{f^n\mathrm{d}f}{1+(f^2/f_W^2)} = [f_W f^{n-1}/(n-1)] - f_W^2 \int \frac{f^{n-2}\mathrm{d}f}{1+(f^2/f_W^2)}$$

When the same function, equation (37), equalizes blue noise, the bandwidth of the resulting blue, equalized, noise is

$$\begin{aligned} f_{\mathrm{BN(BE)}} &= \int_0^{f_C} (f^2/f_T^2)\exp(f^2/f_E^2)\,\mathrm{d}f \\ &= (f_C^3/f_T^2)\{[1/3]+[a^2/5]+[a^4/(2!7)]+ \\ &\qquad +[a^6/(3!9)]+[a^8/(4!11)]+\ldots\} \qquad \ldots\ldots(39) \end{aligned}$$

The change seen after the application of aperture equalization is expressed most meaningfully as a 'noise factor', that is, the ratio of the amplitude of noise after equalization to that before equalization. Thus in Table 10 the middle columns, for 'flat', unweighted noise, are found by dividing equation (38) by (35) and equation (39) by (36).

These expressions derived so far are for noise as read directly on an oscilloscope. To measure the *subjective* effect of noise, it is usual to pass the signal through a simple first-order low-pass ('noise weighting') filter. This simulates to a worthwhile extent the effect produced on the viewer, since the eye is less sensitive to fine, high frequency, detail. However, the accuracy of a particular characteristic may be arguable. For example, the time constant, T, in the transfer function of the 'weighting filter'

$$e_{\mathrm{out}}/e_{\mathrm{in}} = 1/(1+pT) \qquad \ldots\ldots(40)$$

in the C.C.I.R. standard for all 625 line, 50 field/s 5 MHz to 6 MHz systems had been defined previously[11] as $\frac{1}{3}$ μs but had been criticized as being too long. On the other hand, in the standards adopted recently[12] for the new British 625 line, 50 field, 5·5 MHz system (C.C.I.R. system I), the weighting time constant is 0·2 μs. In the work that follows, the latter time constant will be taken as being the more realistic. Hence the characteristic 'weighting' frequency, f_W, at which $2\pi f_W T$ is unity, is 792 kHz. Then when white noise is weighted, its noise bandwidth is

$$f_{\mathrm{BN(WW)}} = \int_0^{f_C} \frac{\mathrm{d}f}{1+(f^2/f_W^2)} = f_W \tan^{-1}(f_C/f_W) \qquad \ldots\ldots(41)$$

Note that when f_C is much greater than f_W, then

$$\tan^{-1}(f_C/f_W) \simeq (\pi/2)-(f_W/f_C) \qquad \ldots\ldots(42)$$

to a good approximation.

When blue noise is weighted, its noise bandwidth is

$$\begin{aligned} f_{\mathrm{BN(BW)}} &= \int_0^{f_C} \frac{(f^2/f_T^2)\,\mathrm{d}f}{1+(f^2/f_W^2)} \\ &= (f_W^2/f_T^2)[f_C - f_W \tan^{-1}(f_C/f_W)] \\ &= (f_W^2/f_T^2)[f_{\mathrm{BN(W)}} - f_{\mathrm{BN(WW)}}] \end{aligned} \qquad \ldots\ldots(43)$$

It is found from equations (35), (36), (41) and (43) that a weighting filter of 0·2 μs time constant (f_w of

792 kHz), in a bandwidth, f_c, of 5 MHz, reduces white noise by 6·5 dB and triangular noise by 12·3 dB.

When white noise is processed by aperture equalization in the manner of equation (37) and is then weighted, its noise bandwidth is

$$
\begin{aligned}
f_{\mathrm{BN(WEW)}} &= \int_0^{f_C} \frac{\exp(f^2/f_E^2)\,df}{1+(f^2/f_W^2)} \\
&= f_W \tan^{-1}(f_C/f_W)\exp[-a^2(f_W^2/f_C^2)] + \\
&+ f_c\{(a^2 f_W^2/f_C^2)[1+(a^2/2!3)+(a^4/3!5)+(a^6/4!7)+ \\
&\qquad +(a^8/5!9)+\ldots] - \\
&-(a^4 f_W^4/f_C^4)[(1/2!)+(a^2/3!3)+(a^4/4!5)+\ldots]+ \\
&+(a^6 f_W^6/f_C^6)[(1/3!)+(a^2/4!3)+(a^4/5!5)+\ldots]- \\
&-(a^8 f_W^8/f_C^8)[(1/4!)+(a^2/5!3)+(a^4/6!5)+\ldots]+\ldots\}
\end{aligned}
\qquad \ldots\ldots(44)
$$

where a is the ratio f_C/f_E, as in equation (38).

When blue noise is processed by aperture equalization and weighted, its noise bandwidth is

$$f_{\mathrm{BN(BEW)}} = (f_W^2/f_T^2)[f_{\mathrm{BN(WE)}} - f_{\mathrm{BN(WEW)}}] \qquad \ldots\ldots(45)$$

The expression for $f_{\mathrm{BN(WEW)}}$ may look formidable but when written in the form of equation (44) it converges rapidly. Thus the expressions for the effect of equalization on weighted noise can be computed readily for the right hand columns of Table 10 by taking the ratios of equation (44) to equation (41) and of equation (45) to equation (43). It is interesting how, in equations (43) and (45), the bandwidths of weighted *blue* noise, both with and without aperture equalization, are derived from the relevant noise bandwidths for *white* noise by subtracting the weighted bandwidth from the unweighted.

The degrees of equalization in Table 10 were chosen in steps of ½ neper for ease of computation but they do correspond well in practice to 'low', 'medium' and 'high' equalization. It will be seen that, when the measurement is on a flat basis, the increase due to aperture equalization of blue noise is little different from that of white noise, say 1·5 times (in dB). However, when they are measured on a weighted basis or are perceived by the viewer, the difference is substantial, nearly 4 times (in dB).

This demonstrates, if demonstration is needed, the wisdom of

(i) peaking input circuits with a 'Percival coil'[13, 14] whenever possible, for example, in a vidicon or plumbicon camera, thereby turning a blue spectrum into something nearer white, and

(ii) specifying and measuring noise on a weighted basis whenever possible.

It must be remembered of course that the figures in Table 10 are for an ideal, inverse Gaussian, shape of equalization and will therefore differ somewhat from those achieved in practice. Nevertheless, the object of this paper has been to point out how that ideal can be much better approximated and the better the approximation, the more valid will be Table 10.

Table 10.
Increase of noise due to aperture equalization

Degree of equalization at f_C (5 MHz)		Measured flat		Measured weighted	
(Np)	(dB)	White (dB)	Blue (dB)	White (dB)	Blue (dB)
0·5	4·3	1·7	2·8	0·5	1·9
1·0	8·7	3·7	5·8	1·1	4·3
1·5	13·0	6·3	9·0	2·1	7·0

9. Conclusion

It is believed that the new kind of aperture equalizer discussed in Section 5 simplifies the design of variable aperture equalizers considerably and at the same time makes possible a much higher degree of precision.

The methods described in Section 3 may also be of interest. For while the method of Fig. 3, using a resonant equalizer with phase corrector, may not be novel, the fact that such precise results can be obtained may be as surprising to the reader as it was to the author.

The circuit of Fig. 4 may not be novel. When the 946 Ω resistor is reduced to 400 Ω, the circuit degenerates into an all-pass network. About eleven years ago, the author encountered a circuit in a German television receiver which was similar in principle but was realized in a different manner. He assumed then that it was an all-pass network[15, 16] but is not so sure now. However, the absence of any other known description, together with its obvious simplicity and effectiveness, are believed to justify its inclusion.

Sufficient work has been done on methods of testing aperture equalizers to indicate that there is still much more useful work to be done.

10. Acknowledgment

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STANDARD FREQUENCY TRANSMISSIONS—September 1970

(Communication from the National Physical Laboratory)

Sept. 1970	Deviation from nominal frequency in parts in 10^{10} (24-hour mean centred on 0300 UT)			Relative phase readings in microseconds N.P.L.—Station (Readings at 1500 UT)		Sept. 1970	Deviation from nominal frequency in parts in 10^{10} (24-hour mean centred on 0300 UT)			Relative phase readings in microseconds N.P.L.—Station (Readings at 1500 UT)	
	GBR 16 kHz	MSF 60 kHz	Droitwich 200 kHz	*GBR 16 kHz	†MSF 60 kHz		GBR 16 kHz	MSF 60 kHz	Droitwich 200 kHz	*GBR 16 kHz	†MSF 60 kHz
1	− 299·9	− 0·1	+ 0·1	683	625·0	**17**	− 300·1	− 0·1	+ 0·1	698	638·7
2	− 300·1	0	+ 0·1	684	625·2	**18**	− 300·0	− 0·1	+ 0·1	698	639·3
3	− 300·0	− 0·1	+ 0·1	684	626·3	**19**	− 300·0	− 0·1	+ 0·1	699	640·1
4	− 300·0	− 0·1	+ 0·1	684	627·2	**20**	− 300·1	− 0·1	+ 0·2	700	641·0
5	− 300·1	0	+ 0·1	685	627·4	**21**	− 299·9	0	+ 0·1	699	641·0
6	− 300·1	− 0·1	+ 0·1	686	627·9	**22**	− 300·1	− 0·1	+ 0·2	700	641·5
7	− 300·0	− 0·1	+ 0·1	686	629·0	**23**	− 300·0	0	+ 0·2	700	641.9
8	− 300·1	− 0·1	0	687	629·9	**24**	− 300·1	− 0·1	+ 0·1	701	642·5
9	− 300·2	− 0·1	0	689	631·1	**25**	− 300·0	− 0·1	+ 0.1	701	643·0
10	− 300·1	− 0·1	0	690	632·3	**26**	− 300·1	− 0·1	+ 0·1	702	644·0
11	− 300·1	− 0·2	0	691	633·9	**27**	− 300·1	− 0·1	+ 0·1	703	644·7
12	− 300·2	− 0·1	0	693	635·1	**28**	− 300·0	− 0·1	+ 0·1	703	645·8
13	− 300·1	− 0·1	0	694	635·9	**29**	− 300·1	+ 0.2	+ 0·1	704	643·9
14	− 300·1	− 0·1	0	695	636·6	**30**	− 299·9	+ 0·1	+ 0·1	703	642·8
15	− 300·2	− 0·1	0	697	637·2						
16	− 300·0	− 0·1	+ 0·1	697	637·9						

All measurements in terms of H.P. Caesium Standard No. 334, which agrees with the N.P.L. Caesium Standard to 1 part in 10^{11}.

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

†Relative to AT Scale; (AT_{NPL} − Station) = + 468·6 at 1500 UT 31st December 1968.

Joint Conference on 'Laboratory Automation'

(I.E.R.E.-I.E.E.-I.P.P.S.-I.M.C.-R.I.C.)

Middlesex Hospital Medical School, London, W.1

Tuesday, 10th November to Thursday, 12th November, 1970

PROGRAMME

Tuesday, 10th November. 10 a.m. to 1.15 p.m. and 2.30 p.m. to 5.30 p.m.

Introductory Address by Dr. I. Maddock, c.b., o.b.e., f.r.s., *Controller Industrial Technology, Department of Trade and Industry.*

Session on 'AUTOMATION IN CHEMICAL, BIOCHEMICAL AND NUCLEAR LABORATORIES'

Chairman: D. A. Patient

'Some Applications of Automation to Analytical Chemistry,' K. C. Steed and F. Trowell, *A.W.R.E.*

'Applications of Ion-Selective Sensors to Automatic Analysis,' D. E. Collis, *Electronic Instruments Ltd.*

'Autolab—A System for Automatic Chemical Analysis Using Discrete Samples,' Sven Gardell, *University of Lund*, and Sven Linstedt, *University of Gothenburg.*

'Automation of Routine Analysis in Works Laboratories,' G. W. De Vries and L. S. Verhoeven, *N.V. Unilever.*

'Productivity in the Analytical Laboratory—A Rational Approach to Automation,' S. R. Gilford, *Gilford Instrument Laboratories Inc.*

Chairman: F. E. Whiteway

'Data Processing at Harwell,' W. M. Currie, *A.E.R.E.*

'Exploiting Small Computers for On-Line Applications,' B. O. Wade, *A.E.R.E.*

'Laboratory Analysis with a Sealed-Tube Neutron Generator,' T. B. Pierce, *A.E.R.E.*

'Control and Monitoring of Neutron Beam Experiments Using Data Processors,' E. E. Maslin, *A.W.R.E.*

'Interface Equipment for the Simultaneous Control of Three Neutron Experiments by a Small Computer,' W. Hutchins and G. Beamont, *A.W.R.E.*

6 p.m. to 7.15 p.m. **Conference Cocktail Party at University College London.**

Wednesday, 11th November. 10.15 a.m. to 1.15 p.m. and 2.30 p.m. to 5 p.m.

Session on 'AUTOMATIC METHODS IN SPECTROMETRY' (including electro-magnetic, nuclear and mass spectrometry and gas chromatography)

Chairman: L. W. Price

'Computer Applications in the Field of Spectroscopy,' R. T. De La Pena, *Honeywell Ltd.*

'A Digital Control System for an Emission Spectrograph,' L. J. Wood and T. R. Grist, *Rank Precision Industries.*

'A Computer Orientated Microdensitometer for Spectrographic Plates,' J. Farren, T. L. Jones, K. J. Reavill and A. Tupman, *A.E.R.E.*

'A Data Acquisition System for Use in Time-Dependent Fourier Transform Spectrometry,' A. E. Costley, T. J. Bonney, P. Garnett and J. E. G. Wheaton, *National Physical Laboratory.*

'Use of an On-Line Digital Computer for Enhancement and Integration of Nuclear Magnetic Resonance Spectra,' R. D. B. Waymark and D. R. Bowman, *University College London.*

Chairman: J. K. Foreman

'A Low Cost High Performance Mass Spectral Data System,' A. Carrick, *National Physical Laboratory.*

'A Manual and Punched Tape Programmer for an X-ray Spectrometer,' J. Harrison, D. Hinkley, J. Watling and J. D. Wilson, *A.E.R.E.*

'Some Hardware Aspects of Computer-Aided Gas Chromatography,' L. M. COLLYER, L. C. H. HAWKINS and G. H. THOMSON, *B.P. Chemicals Limited.*

'Continuous Gas Chromatography,' K. C. NG and G. C. MOSS, *University of Warwick.*

Thursday, 12th November. 9.45 a.m. to 1.15 p.m. and 2.30 p.m. to 5.45 p.m.

Session on 'AUTOMATION IN MECHANICAL, ELECTRONIC, ELECTRICAL AND OTHER LABORATORIES AND OBSERVATORIES'

Chairman: S. FEDIDA

'The Use of a Laboratory Computer as Signal Generator and Correlator,' J. D. BOYLE and A. PALMER, *University of Exeter.*

'Automated Ultrasonic Radar Simulator,' J. R. STAVELEY, *E.M.I. Electronics Ltd.*

'The Automatic Recording and Analysis of Magnetic Fields,' E. LLOYD THOMAS and J. E. LE WARNE, *Admiralty Engineering Laboratory.*

'Low Level Programming for the On-Line Correction of Microwave Measurements,' H. V. SHURMER, *University of Warwick.*

'The Development of Automatic Analysis Equipment at the Coal Research Establishment of the National Coal Board,' J. A. HARRISON and P. H. WITHERIDGE, *National Coal Board.*

'Acquisition and Reduction of Acoustical Noise Data,' M. NATH and I. N. BIRCHALL, *Ruston Paxman Diesels Ltd.*

Chairman: G. S. EVANS

'Random Loading Fatigue Machines On-Line to a Digital Computer,' F. SHERRATT, *University of Warwick.*

'Use of a Process Control Computer in the Automation of Materials Testing and Analytical Laboratories,' T. K. WHITE, *Henry Wiggin and Co. Ltd.*

'Computer Control of Laboratory Experiments and Test Rigs,' G. B. MARSON, *Kent Instruments Ltd.*

'Automatic Temperature Monitoring in a Large Creep Laboratory,' D. M. BRADLEY and I. C. CRAIK, *National Engineering Laboratory.*

'A Multi-Channel Analogue Recording System with Computer Interfaced Data Processing Facilities,' D. W. BAKER and D. R. GAGER, *Marconi Company Limited.*

'Automated Methods in Optical Astronomy,' G. J. CARPENTER, *Royal Observatory, Edinburgh.*

Chairman's Summing-up.

The registration charges for the Conference are £15 for members of the sponsoring Institutions and £18 for non-members; these include the volume of preprints of papers, and lunch and refreshments on the three days of the Conference, as well as a ticket for the Conference Cocktail Party. Registration forms may be obtained on request from the I.E.R.E., 9 Bedford Square, London WC1B 3RG (Telephone 01-637 2771).

Electronic Engineering in Ocean Technology—*continued*

power complementary-symmetry m.o.s. integrated circuits now becoming available are a real boon to the underwater electronic engineer. A surprising number of British ships now carry moderate-sized digital computers, and this is beginning to have a profound effect on both navigation and data-gathering techniques. And finally, Britain has the men, techniques and organization for a good marine instrument industry. We have the tradition and home interest. Though possibly not the foremost nation in the field, we are up amongst the world leaders and should be able to stay there if we handle the situation correctly.

M. J. TUCKER

A complete list of the papers included in the Conference Proceedings is given in an announcement by the I.E.R.E. Publications Department elsewhere in this issue; copies of the volume cost £6 10s. each.

Electronic Engineering in Ocean Technology

A Report on the Joint I.E.R.E.-I.E.E. Conference held at the University College of Swansea, 21st to 24th September 1970.

There has been widespread discussion recently about how much of the nation's resources should be devoted to developing what sort of marine technology. This discussion does not usually cover defence or the construction of merchant ships, which are considered to be separate problems. There is little disagreement in principle about fisheries, where the home industry is large and important and the potential export markets for ships, equipment and services are enormous. The controversy has largely centred on the offshore oil and gas industry where we still depend very largely on U.S. and Dutch expertise; on the exploitation of other offshore minerals; and on fields such as diving and submersibles where the prestige and hobby contents are liable to make hard-headed assessment more difficult.

It is rather surprising that we have heard rather less about the marine instrument industry, although the recent conference in Swansea highlighted the fact that it is an important one. Its turnover is already many millions of pounds sterling per annum, and a large proportion of this is exported: as much as 90% in the case of some of the larger firms. There is a feeling in some quarters that it does not always get the support from government policy which it deserves.

Of the 39 papers presented at the conference, 33 were by British authors. In view of the above comments, it is interesting to note the applications of the instruments and techniques discussed in these papers. These were: Hydrographic surveying (6 papers plus a special address by the Hydrographer of the Navy, Rear-Admiral G. S. Ritchie, C.B., D.S.C., F.R.I.C.S.*); Fishing (4); Oceanographic research (4); Pollution monitoring (3); Defence (3); Geology (2 for research, one for offshore oil and gas operations); Submersibles (2) and 8 on general topics or techniques of wide application.

A little more should be said about hydrographic surveying, which comes top of the above list. In this context I use the word in the restricted sense of charting the shape of the sea bed primarily for the purpose of navigation. It is, of course, an activity of long standing, but the recent development of large bulk carriers, and in particular of super-tankers, has raised new problems, which were discussed by Admiral Ritchie in his address at the Conference Dinner. These ships have draughts which can be in excess of 25m, and therefore have difficulty in navigating in the southern North Sea and certain other areas of the World. Since, for example, in the case of a 200,000 ton tanker each centimetre of lost draught costs something approaching £1,000 per annum, they operate with clearances as low as 1m. The consequences of a grounding can be catastrophic, as the *Torrey Canyon* disaster showed. Thus, the tanker routes must be surveyed in great detail, and at quite frequent intervals since the sea bed is often unstable. Conventional techniques are inadequate, and a much faster surveying system must be developed. During the survey the soundings must be corrected for tidal height, and subsequently the ship operator must know the tidal height in order to calculate the clearance below his keel. A suitable offshore tide gauge is therefore essential. That described in a paper by Collar and Spencer† at the conference is not immediately applicable to this purpose, since it is intended as a research tool and stores its information until it is recovered, whereas the survey gauge must be capable of reading out its information at the end of each day without being moved.

Of the six overseas papers at the conference, three were French, two were from the U.S.A. and one was from Canada Five were on instruments for oceanographic research, and one was a general paper. Of the 240 (approximately) participants at the conference, 22 were from overseas, coming from 10 different countries. This was a much smaller proportional overseas representation than that at the Conference on Electronic Engineering in Oceanography organized by the Institution at Southampton in 1966, and it is worth asking why this should be. It is at least partly because of the increase in the number of conferences: there were at least two other conferences on similar subjects being held at the same time as this one, one in the U.S.A. and one in Japan, and another is being held in Germany two months later. The organizing committee was not aware of these when the date was originally chosen, and it proved impracticable to alter the date later. In the upshot this proved to be something of a blessing, since otherwise the conference might have been overwhelmed. The number of papers offered was considerably more than could be accepted, and even with a certain amount of strictness about dates so that all papers could be pre-printed, the final number was as many as could reasonably be presented in the time available. The number of delegates was ideal: enough to be technically representative, but not so large as to overflow the largest lecture hall available.

Some rather random thoughts and impressions generated by the conference were as follows. The President of the I.E.R.E., H. F. Schwarz, who opened the conference, leads one of the most successful marine instrument engineering companies in the world. Modern electronics is so compact and, above all, reliable, that one can now put sophisticated electronic systems underwater with a reasonable chance of success. D. C. Webb‡ pointed out that since the supply of power is one of the main problems in self-contained underwater equipment, the very low

* To be published in the November issue of *The Radio and Electronic Engineer*.

† Collar, P. G., and Spencer, R., 'A digitally recording offshore tide gauge', I.E.R.E. Conference Proceedings No. 19, pp. 341–52.

‡ Webb, D. C., *et al.* 'A new instrument for the measurement of vertical currents in the ocean', I.E.R.E. Conference Proceedings No. 19, pp. 323–31.

(*Continued on page 214*)

Contributors to this issue

Mr. John Forrester obtained a B.Sc. degree in physics at Battersea College of Technology in 1966. He then joined Muirhead and Co. of Beckenham where he worked on solid-state power conditioning systems. The correlation work described in the paper results from his project and thesis during an M.Sc. course in space electronics and communications at the University of Kent at Canterbury. He is currently a senior engineer in the spacecraft department of G.E.C.-A.E.I. Electronics, Portsmouth, Hants.

Mr. D. S. Girling received his technical education at Loughborough College of Engineering and at the School of Signals, Catterick during the war. He joined the capacitor laboratory of Standard Telephones and Cables in 1946, took charge of the laboratory in 1948 and became chief engineer of the capacitor division in 1957. During this time his main interests were the development of tantalum electrolytic and metallized plastic film capacitors, film circuits, mechanisms of failure, reliability and quality control. In January 1965 he was appointed Quality Manager and he is also Group Quality Manager of ITT Components Group.

Mr. Girling has been active in standardization work through both the Radio and Electronic Component Manufacturers Federation and the British Standards Institution, and he is currently chairman of the B.S.I. Capacitor Sub-Committee (TLE/4/4).

Dr. J. A. M. McDonnell obtained his physics degree at Manchester University in 1960 and during his Ph.D. studies developed space instrumentation for sounding rockets and the *Ariel II* satellite at Nuffield Radio Astronomy Laboratories, Jodrell Bank. He continued development of space instrumentation and cosmic dust simulation techniques at NASA Goddard Space Flight Centre, Greenbelt, Md., U.S.A., from 1965 to 1967. In 1967 he returned to take up a lecturership at the University of Kent at Canterbury. He is currently organizing a M.Sc. course on advances in communications and digital systems.

Mr. Neville Thiele was educated at Queensland University and the University of Sydney, graduating as Bachelor of Engineering in 1952. In 1952 he joined the staff of E.M.I. Australia Ltd. as a development engineer in the Special Products Division. In 1955 he spent six months in England, Europe and the U.S.A. and on return was responsible for the development of E.M.I. Australia's television receiver. In 1957 he was appointed advance development engineer.

In 1962 Mr. Thiele joined the Australian Broadcasting Commission, and is now senior engineer, design and development, responsible for the Federal Engineering Laboratory, and concerned with design and investigation of equipment and systems for sound and vision broadcasting.

Dr. D. C. Cooper is a senior lecturer in the Department of Electronic and Electrical Engineering at the University of Birmingham. A note on his career was published in the December 1968 issue of *The Radio and Electronic Engineer*.