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A New Deal for Technicians

THE increasing importance of the technician in our modern technological society can hardly be underestimated and will be fully appreciated by readers of this Journal: the figure of well over three-quarters of a million in Great Britain in 1971 for manufacturing industry, the public sector of industry, in construction and in central and local government is impressive, as is the statistic of well over 100,000 in research and development. Set up by the Labour Government in 1967, the Haslegrave Committee on Technician Courses and Examinations reported just over three years ago* and its principal recommendation for the establishment of a Technician Education Council has now been implemented. On 14th March the Education Secretary, Mrs. Margaret Thatcher, stated that:

'The negotiations between my Department and the City and Guilds of London Institute about the administrative servicing of the Council have been successfully completed, and other valuable discussions have been held with professional and other interests involved in Joint Committees for national certificate and diploma awards. I am satisfied that the new Council will be able to count on the broad support of these several interests in its work.

'The Council will be concerned in the development of policies for schemes of technical education for persons at all levels of technician occupations in industry and elsewhere. To this end it will, as proposed in the Haslegrave Report, plan, administer and keep under review the development of a unified national system of courses for such people; and will devise or approve suitable courses, establish and assess standards of performance, and award certificates and diplomas as appropriate. I shall want the Council to be associated with the courses and examinations administered by Joint Committees for national certificates and diplomas and with the technician level examinations of the CGLI until all of these are replaced. In all its work I see the Council proceeding with the fullest possible involvement and co-operation of industry, the professional bodies, and the education service.'

The Council will have as its part-time Chairman Mr. A. L. Burton, present Chairman of the Furniture and Timber Industrial Training Board and 23 members drawn from education, industry and technical and professional bodies.

The long-term aim of the Council will be to unify the present great diversity of educational qualifications in different disciplines for technician engineers and technicians. It will fall into a logical place between the professional engineer and the craftsman and it is to be hoped that unification of qualifications and courses will not lose that flexibility which is desirable if craftsmen are to progress to technicians and technicians to become professional engineers.

This Institution was among those who submitted evidence to the Haslegrave Committee and in general welcomed its recommendations. One reservation, on which the Engineers Registration Board made similar comments, related to the name for the proposed Council. It was strongly felt that 'Technical Education Council' would be a more suitable title, since the activities of the Council would not be limited to education at what ERB consider to be Technician level but would also include Technician Engineer courses and examinations. These representations were not accepted by the Department of Education and Science, whose authority is confined to England and Wales: however, the Scottish Education Department has set up a Technical Education Council!

These semantic criticisms apart, assuming that the Council's terms of reference and determination of its policy will follow the broadly accepted recommendations of the original Haslegrave Report, the new organization will be welcomed as long overdue. If we may quote from a paper given at the 1966 Commonwealth Conference on the Education and Training of Technicians 'the technician is neither a superior tradesman nor a depressed technologist' and the building of courses and schemes of training for the technician must bear this in mind.

* 'Report of the Committee on Technician Courses and Examinations' HMSO, 1969, price 62½p. (SBN 270143 4).

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* See also pages 252 and 256.

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Theory of cylindrical antennas with lumped impedance loadings

B. D. POPOVIĆ, D.Sc.*

SUMMARY

A method is presented for the analysis of a thin symmetrical cylindrical antenna with lumped impedance loadings along its length. The loadings can be resistive, inductive, capacitive, or mixed, and their physical size is assumed to be of the order of the antenna radius. The values of admittances of such structures obtained by the theory are in good agreement with available experimental results.

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1 Introduction

Cylindrical antennas with lumped loadings of appropriate magnitudes and at appropriate positions along the antenna length are known to offer interesting possibilities. Altshuler[1] obtained an essentially travelling-wave cylindrical monopole antenna by introducing a resistive loading a quarter-wavelength from its end. Hallén[2] noticed that a travelling-wave antenna can also be obtained by making a monopole antenna in the form of a row of small conducting cylinders between which dielectric disks of increasing thickness toward the monopole end are sandwiched. A similar structure was analysed experimentally by Rao *et al.*[3] Nyquist and Chen[4,5] proposed a travelling-wave antenna similar to that of Altshuler, but with reactive instead of with resistive loading. Relatively recently, two methods were described for analysis of loaded cylindrical antennas, the loading assumed to be continuous.[6,7] Finally, measurements on a simple construction of monopole with tapered[8] and constant[9] quasi-distributed capacitive loading were reported. These results demonstrated a satisfactory accuracy of the continuous-loading theory[7] when applied to monopoles with lumped loadings, provided these are not too far apart from one another ($\ngtr 0{\cdot}15\ \lambda$).

The present paper is aimed at describing a novel general method for analysis of thin symmetrical metallic antennas with lumped impedance loadings. The method lends itself particularly to cases when the loadings are electrically small, such as narrow air-gaps between conducting cylinders making a capacitively loaded antenna described in References 8 and 9. Essentially, the proposed method represents a generalization of the polynomial approach for analysis of perfectly conducting dipole antennas,[10] as current distribution between adjacent loadings is approximated also by a polynomial of arbitrary order with complex coefficients. The approximation for current along the antenna is therefore a piecewise polynomial approximation. The often used (though in methods differing from the present one) step-function and piecewise linear approximations,[11–13] or the parabolic approximation[14] are obtained in the present method as special cases.

2 Outline of the Method

2.1 *Integral Equation for Current along Antenna with Lumped Loadings*

Consider a perfectly conducting cylindrical dipole of radius a and length $2h$ ($h \gg a$), centre-driven by a delta-function generator of voltage V and angular frequency ω. Let the dipole be situated in a vacuum, and let $a/\lambda \ll 1$, λ representing the free-space wavelength of the radiated wave. We assume the dipole to be loaded with lumped loadings $Z_2, Z_3, \ldots, Z_{n+1}$, at distances $\pm z_2, \pm z_3, \ldots, \pm z_{n+1}$ ($z_{n+1} < h$) from the generator, which is positioned at x = 0 (Fig. 1). For convenience in mathematical formulation, which will be apparent later, we let $Z_1 = 0$ at $z_1 = 0+$, and put $z_{n+2} = h$. Assuming the loadings to be of vanishing sizes, we have the following equation which must be satisfied on the dipole surface:

$$E_z(z) = -V\delta(z) + {}^{'}\sum_{i=1}^{n+1} Z_i I(z_i)\{\delta(z-z_i)+\delta(z+z_i)\}. \qquad (1)$$

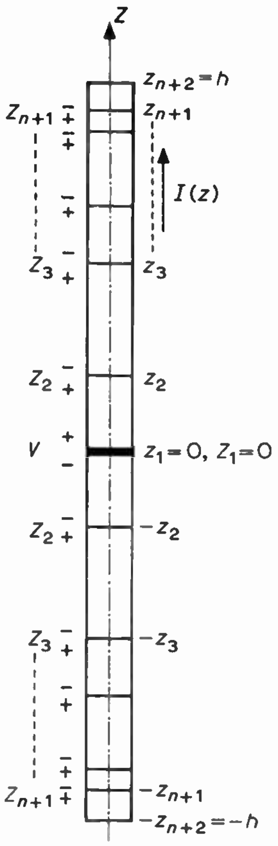

Fig. 1. Dipole with lumped loadings.

In this equation, $E_z(z)$ is the z-component of the electric field strength on the antenna surface, $\delta(z)$ is the Dirac delta-function, and $I(z)$ the current along the dipole.

In terms of the magnetic vector potential $A_z(z)$, $E_z(z)$ can be expressed as

$$E_z(z) = -\mathrm{j}\omega\left(1+\frac{1}{\beta^2}\frac{\partial^2}{\partial z^2}\right)A_z(z) \tag{2}$$

where $\beta = \omega(\epsilon_0\mu_0)^{1/2}$ is the free-space propagation coefficient. Combining eqns. (1) and (2) we get

$$\mathrm{j}\omega\left(1+\frac{1}{\beta^2}\frac{\partial^2}{\partial z^2}\right)A_z(z) = V\delta(z) - \\ -\sum_{i=1}^{n+1} Z_i I(z_i)\{\delta(z-z_i)+\delta(z+z_i)\} \tag{3}$$

Owing to the structure symmetry, $I(-z) = I(z)$. Therefore the magnetic vector potential $A_z(z)$ and the electric scalar potential $\phi(z)$ on the antenna surface must satisfy the following requirements:

$$A_z(-z) = A_z(z) \tag{4}$$

$$\phi(0+)-\phi(0-) = V \tag{5}$$

$$\phi(\pm z_i-0)-\phi(\pm z_i+0) = Z_i I(\pm z_i), \\ i = 2,3,\ldots,(n+1) \tag{6}$$

With these conditions, and noting that

$$\phi(z) = \frac{\mathrm{j}\omega}{\beta^2}\frac{\partial A_z(z)}{\partial z} \tag{7}$$

(the Lorentz condition), the solution of eqn. (3) is

$$A_z(z) = C\cos\beta z + \frac{\beta V}{2\mathrm{j}\omega}\sin\beta|z| - \\ -\frac{\beta}{2\mathrm{j}\omega}\sum_{i=1}^{n+1} Z_i I(z_i)\{\sin\beta|z-z_i| + \sin\beta|z+z_i|\} \tag{8}$$

where C is a complex constant to be determined.

The magnetic vector potential can also be expressed in terms of the current $I(z)$,

$$A_z(z) = \frac{\mu_0}{4\pi}\int_{-h}^{h}\frac{I(s)\exp(-\mathrm{j}\beta r)}{r}\,\mathrm{d}s \tag{9}$$

where

$$r = \{(z-s)^2+a^2\}^{1/2} \tag{10}$$

Combining eqns. (8) and (9) we obtain the integral equation for current $I(z)$ along the loaded dipole of Fig. 1:

$$\int_{-h}^{h} I(s)\frac{\exp(-\mathrm{j}\beta r)}{r} = C_1\cos\beta z + \frac{2\pi}{\mathrm{j}\zeta_0}\Big[V\sin\beta|z| - \\ -\sum_{i=1}^{n+1} Z_i I(z_i)\{\sin\beta|z-z_i| + \sin\beta|z+z_i|\}\Big] \tag{11}$$

where $C_1 = 4\pi C/\mu_0$, and $\xi_0 = (\mu_0/\epsilon_0)^{1/2}$ is the intrinsic impedance of a vacuum. This equation must be satisfied for any z, and thus also for $z = 0$. For $z = 0$ it becomes

$$\int_{-h}^{h} I(s)\frac{\exp(-\mathrm{j}\beta r_0)}{r_0}\,\mathrm{d}s \\ = C_1 - \frac{2\pi}{\mathrm{j}\zeta_0}\sum_{i=1}^{n+1} 2Z_i I(z_i)\sin\beta z_i \tag{12}$$

where

$$r_0 = (s^2+a^2)^{1/2} \tag{13}$$

from which the unknown constant C_1 can be expressed in terms of $I(z)$. From eqns. (11) and (12) we finally obtain

$$\int_{-h}^{h} I(s)\left\{\frac{\exp(-\mathrm{j}\beta r)}{r} - \cos\beta z\frac{\exp(-\mathrm{j}\beta r_0)}{r_0}\right\}\mathrm{d}s \\ = \frac{2\pi}{\mathrm{j}\zeta_0}\Big[V\sin\beta|z| - \sum_{i=1}^{n+1} Z_i I(z_i)\{\sin\beta|z-z_i| + \\ +\sin\beta|z+z_i| - 2\cos\beta z\sin\beta z_i\}\Big] \tag{14}$$

This equation can be written in a shorter form,

$$\int_{-h}^{h} I(s)F(z,s)\,\mathrm{d}s + \sum_{i=1}^{n+1} I(z_i)H(z,z_i) = G(z) \tag{15}$$

where

$$F(z,s) = \frac{\exp(-\mathrm{j}\beta r)}{r} - \cos\beta z\frac{\exp(-\mathrm{j}\beta r_0)}{r_0} \tag{16}$$

$$G(z) = -\mathrm{j}\frac{2\pi V}{\zeta_0}\sin\beta|z| \tag{17}$$

and

$$H(z,z_i) = \mathrm{j}\frac{2\pi Z_i}{\zeta_0}\{2\cos\beta z\sin\beta z_i - \\ -\sin\beta|z-z_i| - \sin\beta|z+z_i|\}. \tag{18}$$

2.2 *Approximate Solution of the Integral Equation for Current*

As in the case of unloaded dipoles, a solution for the integral equation (15) in a closed form is not known. An approximate solution can be obtained by assuming the current to be in the form of a finite functional series with unknown complex coefficients, and calculating coefficients by stipulating the integral equation to hold at a sufficient number of points along the antenna (the so-called point-matching method). In order that the method be successful, the functional series must be chosen so that it can approximate current distribution accurately. However, the current distribution is not known, and consequently as flexible as possible a functional series is required for approximation of current. The simple power series appears to be very convenient for that purpose.

Let us therefore approximate current distribution on the segment $z_i \leqslant z \leqslant z_{i+1}$ by

$$I(z) = \sum_{k=1}^{n_i} I_{i,k}\left|\frac{z}{z_{i+1}}\right|^{k-1}, \quad z_i \leqslant z \leqslant z_{i+1},\ i = 1,2,\ldots,(n+1) \qquad (19)$$

where (n_i-1) is the desired order of the polynomial approximation for current along the ith segment (i.e. between points z_i and z_{i+1}), and $I_{i,k}$ are complex constants to be determined. Substituting $I(z)$ as given by eqn. (19) into eqn. (15) we get

$$\sum_{i=1}^{n+1}\left\{\int_{z_i}^{z_{i+1}} - \int_{-z_i}^{-z_{i+1}}\right\}\sum_{k=1}^{n_i} I_{i,k}\left|\frac{s}{z_{i+1}}\right|^{k-1} F(z,s)\,ds + \sum_{i=1}^{n+1}\sum_{k=1}^{n_i} I_{i,k}\left|\frac{z_i}{z_{i+1}}\right|^{k-1} H(z,z_i) = G(z) \qquad (20)$$

This can be put in the form

$$\sum_{i=1}^{n+1}\sum_{k=1}^{n_i} I_{i,k}\,P(z,i,k) = G(z) \qquad (21)$$

where

$$P(z,i,k) = Q(z,i,k) + \left|\frac{z_i}{z_{i+1}}\right|^{k-1} H(z,z_i) \qquad (22)$$

and

$$Q(z,i,k) = \frac{1}{(z_{i+1})^{k-1}}\int_{z_i}^{z_{i+1}} s^{k-1}\{F(z,s)+F(z,-s)\}\,ds \qquad (23)$$

The complex coefficients $I_{i,k}$ must also satisfy the following $(n+1)$ equations:

$$\sum_{k=1}^{n_i-1} I_{i-1,k} = \sum_{k=1}^{n_i} I_{i,k}\left(\frac{z_i}{z_{i+1}}\right)^{k-1}, \quad i = 2,3,\ldots,(n+1) \qquad (24)$$

$$\sum_{k=1}^{n_{n+1}} I_{n+1,k} = 0 \qquad (25)$$

These equations express continuity of current at points z_i, $i = 2,3,\ldots,(n+1)$, and the fact that $I(\pm h) = 0$, respectively. The total number of unknown complex coefficients $I_{i,k}$ being equal to $N = (n_1+n_2+\ldots+n_{n+1})$, we must write another $(N-n-1)$ equations using eqn. (21). To do this, we stipulate that eqn. (21) be satisfied at $(N-n-1)$ points along the antenna. We choose

$$z_p = z_i + \frac{z_{i+1}-z_i}{n_i-1}m, \qquad m = 1,2,\ldots,(n_i-1), \quad i = 1,2,\ldots,(n+1) \qquad (26)$$

and stipulate that

$$\sum_{i=1}^{n+1}\sum_{k=1}^{n_i} I_{i,k}\,P(z_p,i,k) = G(z_p), \quad p = 1,2,\ldots,(N-n-1) \qquad (27)$$

Equations (24), (25) and (27) constitute a system of N linear complex equations in N complex unknowns $I_{i,k}$. The integrals $Q(z,i,k)$ given by eqn. (23) cannot be integrated explicitly, but it is possible to integrate them numerically, and then solve the equations (24), (25) and (27) for the unknowns $I_{i,k}$. However, care must be exercised in numerically evaluating the integrals $Q(z,i,k)$. Since $z_{i+1} > z_i \geqslant 0$, the integrand can, in general, have two sharp peaks in the range of integration (at $s = 0$ and at $s = z$), which are relatively difficult to integrate accurately. A simpler integration is obtained if the two terms of $F(z,s)$ given by eqn. (16) are integrated separately.

Once the coefficients $I_{i,k}$ are determined the current distribution along the antenna is given by eqn. (19), and the antenna driving-point admittance is simply

$$Y = I(0+)/V = I_{1,1}/V \qquad (28)$$

The current distribution along the antenna being known, the radiation pattern of the antenna can be calculated easily.

3 Numerical Results

3.1 *The Altshuler Antenna*

Altshuler[1] started from a transmission line analogy, and predicted that a travelling wave could be maintained along a part of a thin cylindrical monopole, if a resistive loading of appropriate magnitude is inserted $\lambda/4$ from the monopole end. He verified this conclusion experimentally, on a monopole of radius $a = 0{\cdot}3175$ cm, at a frequency of $f = 600$ MHz. The outer radius of the coaxial feeder to the monopole was equal to $8a$. Altshuler found that a resistance of 240 Ω at $\lambda/4$ from the monopole end was the optimal loading, resulting in an essentially travelling current wave from the excitation point to the load, and a standing wave from the load to the monopole end. The monopole exhibited fairly constant input admittance for practically any length of the monopole section from the ground plane to the load.

The present theory was applied first to the Altshuler antenna. Figure 2 shows dependence of real (G) and imaginary (B) parts of the monopole admittance, versus the normalized length h/λ of the monopole, as measured by Altshuler, and as calculated by the present theory. Agreement in conductance is seen to be very satisfactory. The theoretical susceptance is higher than the experimental, although the shape of the susceptance curves in the two cases is very much alike. This discrepancy is due to poor approximation of the real driving conditions (coaxial line with outer-to-inner-diameter ratio as large as 8) by the theoretical model of the excitation zone (the delta-function generator). The theoretical results were obtained by dividing the monopole into segments of

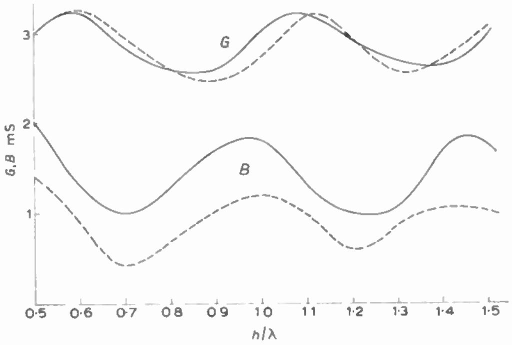

Fig. 2. Real (G) and imaginary (B) parts of admittance of the Altschuler antenna.

$a = 0{\cdot}3175$ cm, $f = 600$ MHz, $Z_{load} = 240\,\Omega$ at $z = \pm(h-\lambda/4)$.
—— Present theory. ---- Experimental (Altshuler[1]).

lengths less than $\lambda/2$, and using the second-order polynomial approximation for current (i.e. $n_i = 3$) along such segments.

As an example of current distribution, Fig. 3 shows experimental and theoretical current distribution for $h/\lambda = 5/8$. It is seen that agreement between theoretical and experimental curves is excellent.

3.2 *Antennas with Lumped Capacitive Loadings*

Assuming that, approximately, $I(z)$ and $A_z(z)$ along a thin cylindrical antenna are proportional, Nyquist and Chen[4, 5] reached a conclusion that an appropriately positioned purely reactive loading should also result in a travelling current wave up to the loading. Their experiments confirmed that prediction qualitatively, but the particular type of the loading (parallel connexion of a gap, equivalent to a capacitor, with a short-circuited section of a coaxial line, equivalent to an inductance) did not allow them to determine the value of the optimal load experimentally. Of course, this structure was much more frequency sensitive than the Altshuler antenna.

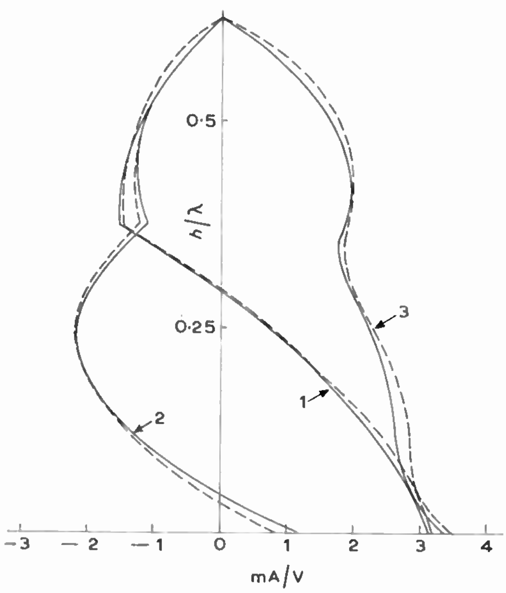

Fig. 3. Real part (1), imaginary part (2) and magnitude (3) of current along the Altshuler antenna.

$a = 0{\cdot}3175$ cm, $h = 31{\cdot}25$ cm, $f = 600$ MHz, $Z_{load} = 240\,\Omega$ at $z_2 = \pm(h-\lambda/4) = \pm 18{\cdot}75$ cm.
—— Present theory, $n_1 = n_2 = 3$.
---- Experimental (Altshuler[1]).

According to Nyquist-Chen experiments,[5] a dipole of half-length $h = 1$ m and radius $a = 0{\cdot}3175$ cm should sustain at 600 MHz an essentially travelling wave up to the load, and have an admittance of about $(2{\cdot}80 + \mathrm{j}1{\cdot}65)$ millisiemens, if loaded symmetrically with appropriate reactances (theoretically, $-\mathrm{j}363\,\Omega$) at $z = \pm 81{\cdot}1$ cm. The results obtained by the present theory indicate that the optimum load (i.e. for which the standing-wave ratio is minimal) is approximately $-\mathrm{j}475\,\Omega$. To this loading corresponds the dipole admittance of about $(2{\cdot}70 + \mathrm{j}1{\cdot}52)$ mS, which is in good agreement with the result of Nyquist and Chen. Current distribution in the optimal and a close nonoptimal cases is shown in Fig. 4. It is seen that in the optimal case current up to the load is approximately a travelling wave, of the shape quite similar to that obtained experimentally by Nyquist and Chen.

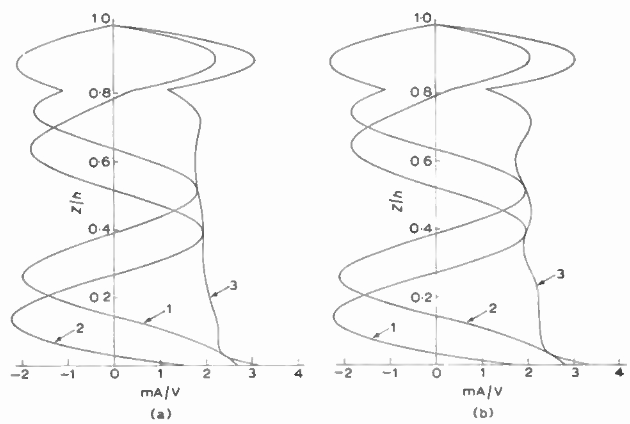

Fig. 4. Theoretical current distribution along dipole of length $2h = 2$ m, with capacitive loadings at $z_2 = \pm 0{\cdot}811$ m from the centre; $a = 0{\cdot}3175$ cm, $f = 600$ MHz, $n_1 = 9$, $n_2 = 3$, (1) rea part, (2) imaginary part, (3) magnitude.

(a) Optimal loading ($Z_2 = -\mathrm{j}475\,\Omega$).
(b) Loading larger than optimal ($Z_2 = -\mathrm{j}550\,\Omega$).

The present theory was next applied to the case of broadband capacitively loaded antenna of the type described in Reference 8. An antenna was made in the form of 15 rings of lengths $(3{\cdot}2 - 0{\cdot}2 \times n)$ cm, $n = 1, 2, \ldots, 15$, the protrusion of the inner coaxial line conductor above the ground plane being $3{\cdot}2$ cm. The ring and inner coaxial line conductor diameter was $0{\cdot}6$ cm, diameter of the outer cable conductor was $1{\cdot}4$ cm, and the glass tube onto which the rings were mounted had a diameter of $0{\cdot}5$ cm. The width of all the 15 gaps was about $0{\cdot}2$ mm. Experimental and theoretical values of susceptance and conductance for such a monopole antenna are shown in Fig. 5. The gaps were assumed to be equivalent to loadings of $-\mathrm{j}250\,\Omega$ each at 1000 MHz, and piecewise

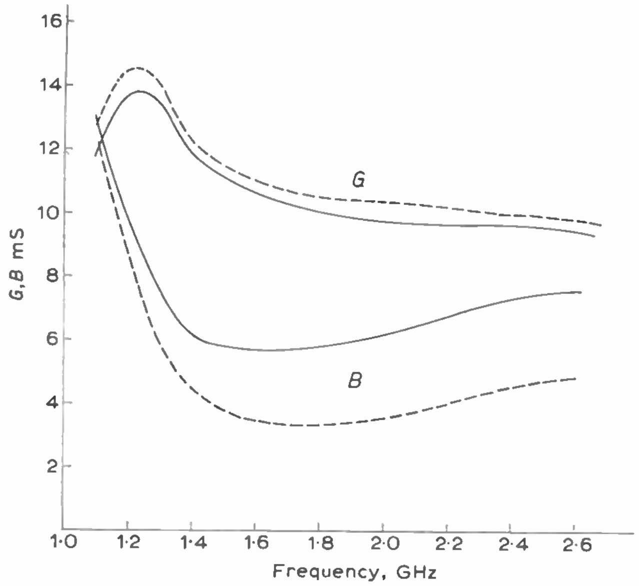

Fig. 5. Conductance (G) and susceptance (B) of monopole with tapered capacitive loading (for description of the monopole see Sect. 3.2); $a = 0{\cdot}3$ cm, $h = 27{\cdot}5$ cm.

— — Present theory, $Z_2 = Z_3 = \ldots = Z_{16} = -j250\Omega$ at 1 GHz, $n_1 = n_2 = \ldots = n_{16} = 3$.

---- Experimental.

parabolic approximation of current ($n_i = 3$) was adopted along all segments. Agreement in conductance is seen to be satisfactorily close. The theoretical susceptance is larger than experimental, which can again be explained by different driving mechanisms in the two cases.

As an example of current distribution, Fig. 6 shows theoretical real and imaginary parts of current, as well as its magnitude, along the antenna described in the preceding paragraph, at a frequency of 2 GHz. Note the overall decaying travelling-wave behaviour of the current distribution, and the standing-wave behaviour between each two adjacent loadings.

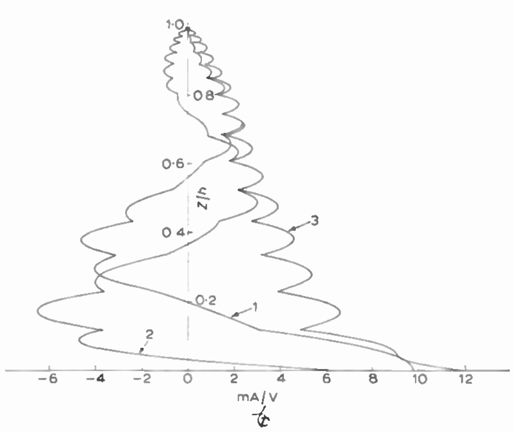

Fig. 6. Current distribution along monopole with tapered capacitive loading (for full description of the monopole see Sect. 3.2); $a = 0{\cdot}3$ cm, $h = 27{\cdot}5$ cm, $f = 2$ GHz. (1) Real part, (2) imaginary part, (3) magnitude.

3.3 *Theoretical Comparison between Lumped and Continuous Loadings*

As the final example, let us compare theoretical admittance of and current distribution along a loaded dipole in the following two cases: (1) the total loading is evenly distributed along the antenna length, and (2) the total loading of equal magnitude is divided into several lumped loadings. We shall analyse the case of resistive antenna with constant continuous loading, for which experimental results are available.[15]

Consider a resistive dipole of half-length $h = 0{\cdot}226$ m and radius $a = 0{\cdot}3175$ cm. Let the total loading along half of the dipole be $317\,\Omega$, which is $1400\,\Omega$/m, and let the frequency be $f = 663$ MHz, i.e. $\beta h = \pi$. For the case of such continuous loading experimental values of admittances are given in Reference 15, and theoretical results are presented in Reference 16. For convenience, these are given again in Table 1.

Let now the total loading be concentrated at $n = 1, 2, 3$ and 4 equidistant points along the dipole arms. Values of G and B corresponding to these cases are also shown in Table 1. It is seen that as little as four loadings are sufficient to substitute the continuous loading, at least as far as dipole conductance is concerned. (Theoretical susceptance is, of course, higher in the lumped loading case, as relatively high, second-order polynomial approximation for current is used for a small segment of the antenna in the vicinity of the delta-function generator.)

Table 1

Comparison of admittances of resistive dipoles with continuous and lumped equidistant loadings; $a = 0{\cdot}3175$ cm, $h = 0{\cdot}226$ m, $f = 663$ MHz, total loading along one dipole arm $= 317\,\Omega$ (i.e. $1400\,\Omega$/m). In the case of lumped loadings, piecewise parabolic approximation of current has been adopted.

Type of loading	Admittance, mS
Single loading at $\pm h/2$	$3{\cdot}70 + j2{\cdot}11$
Double loading, at $\pm h/3$ and $\pm 2h/3$	$2{\cdot}48 + j2{\cdot}73$
Triple loading, at $\pm h/4$, $\pm 2h/4$ and $\pm 3h/4$	$2{\cdot}15 + j2{\cdot}70$
Quadruple loading, at $\pm h/5$, $\pm 2h/5$, $\pm 3h/5$ and $\pm 4h/5$	$2{\cdot}04 + j2{\cdot}67$
Continuous loading, theoretical, 3rd order polynomial approximation for current[16]	$1{\cdot}90 + j1{\cdot}91$
Continuous loading, experimental (Shen[15])	$1{\cdot}9 \ + j2{\cdot}2$

Compared in Fig. 7 is the theoretical current distribution in the case of continuously loaded dipole of electrical half-length $\beta h = \pi$, with those corresponding to 2 and 4 lumped loadings. It is evident that as little as 4 lumped loadings can approximate the continuous loading quite accurately.

4 Conclusion

An approximate method is presented for determining current distribution along thin cylindrical dipoles with

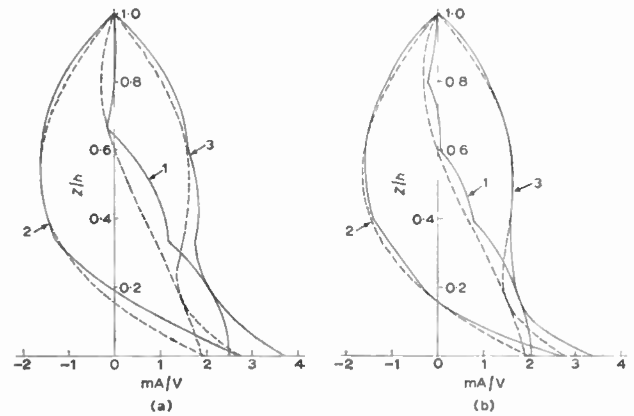

Fig. 7. Theoretical current distribution along resistive dipole; $a = 0{\cdot}3175\,\text{cm}$, $h = 0{\cdot}226\,\text{m}$, $f = 663\,\text{MHz}$, total loading along half of the dipole $= 317\,\Omega$, i.e. $1400\,\Omega/\text{m}$.

—— Lumped equidistant loadings, piecewise parabolic approximation of current.

---- Continuous loading, 3rd order polynomial approximation of current.[16]

(1) Real part, (2) imaginary part, (3) magnitude.

(a) Two lumped loadings. (b) Four lumped loadings.

lumped loadings along their length. Only the case of centre-fed dipoles with pairs of identical loadings positioned symmetrically with respect to the generator is presented, but with certain modifications the theory can be extended to the asymmetrical case.

Essentially, the method consists in approximating the current distribution between adjacent loadings by a polynomial of any desired order, that is, in approximating current along the antenna by a piecewise polynomial function. The number of loadings and the order of the polynomial approximation are limited only by the word length and the storage capacity of the computer available. The advantage of the present method over the existing methods is its conceptual simplicity and modest computer storage requirement, combined with satisfactory accuracy.

The accuracy of the theory was checked by computing admittances of dipoles with resistive and capacitive loadings for which experimental data are available. In all cases agreement between theoretical and experimental results was found to be good.

5 Acknowledgments

This work was in part supported by the Serbian National Research Foundation. The examples of Section 3.2 were computed during the author's stay with the Laboratory of Electromagnetic Theory, The Technical University of Denmark, Lyngby. Experimental results shown in Fig. 5 were obtained by the author's colleague M. B. Dragović.

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The enhancement of spurious signals in non-linear frequency multipliers

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SUMMARY

An inherent disadvantage of non-linear frequency multipliers when fed from a signal source containing spurious signals, is the enhancement of these signals in the multiplier pass-band as a function of the degree of multiplication. An analysis showing the enhancement effect is given, and some experimental results are described.

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1 Introduction

Where a requirement exists for rapid microwave frequency changing, such as in a frequency agile coherent radar, it is expedient to carry out the switching operation at low frequency and then multiply up to the transmitter frequency.

Most considerations of non-linear frequency multipliers assume that the frequency to be multiplied is pure; in every case, however, the purity is a matter of degree. For example, should the frequency be obtained from a high-grade signal generator the spurious signals would consist of the inherent noise of the source, probably 120–160 dB below the level of the signal frequency. If, however, the signal is derived from a frequency modulator or mixing circuit, the spurious signals in the multiplier pass-band will consist of sidebands as well as harmonics of the local oscillator and of the signal frequencies. In practice the amplitude for these frequencies would be 30–40 dB below the signal frequency level.

2 Analysis

Frequency multipliers in general employ non-linear input/output characteristics to produce harmonics of the input signal frequency. The characteristic is of the form:

$$E_{\text{out}} = aV_{\text{in}} + bV_{\text{in}}^2 + cV_{\text{in}}^3 \ldots$$

The coefficients a, b, c in the expression represent respectively 1st, 2nd and 3rd harmonic frequency contents, and the desired order of multiplication is selected by filtering out the unwanted harmonics.

Consider for example a frequency doubler.

Let V_{in}, the input, consist of two frequencies separated *by a small* amount. $E_1 \sin \omega_1 t$ is the main signal and $E_2 \sin \omega_2 t$ the unwanted or spurious signal, so that

$$V_{\text{in}} = E_1 \sin \omega_1 t + E_2 \sin \omega_2 t.$$

A simple analysis considering only those terms up to and including 2nd order shows:

$$\begin{aligned}
E_{\text{out}} &= a[E_1 \sin \omega_1 t + E_2 \sin \omega_2 t] + \\
&\qquad + b[E_1 \sin \omega_1 t + E_2 \sin \omega_2 t]^2 \\
&= a[E_1 \sin \omega_1 t + E_2 \sin \omega_2 t] + \\
&\qquad + bE_1^2 \sin^2 \omega_1 t + 2bE_1 E_2 \sin \omega_1 t . \sin \omega_2 t + \\
&\qquad + bE_2^2 \sin^2 \omega_2 t \\
&= bE_1^2 \sin^2 \omega_1 t + bE_2^2 \sin^2 \omega_2 t + \\
&\qquad + 2bE_1 E_2 \sin \omega_1 t . \sin \omega_2 t + F_1 \\
&= bE_1^2 \left(\frac{1-\cos 2\omega_1 t}{2}\right) + bE_2^2 \left(\frac{1-\cos 2\omega_2 t}{2}\right) + \\
&\qquad + 2bE_1 E_2 \sin \omega_1 t . \sin \omega_2 t + F_1 \\
&= \frac{b}{2}(E_1^2 + E_2^2) - \frac{b}{2}(E_1^2 \cos 2\omega_1 t + E_2^2 \cos 2\omega_2 t) + \\
&\qquad + 2bE_1 E_2 \sin \omega_1 t . \sin \omega_2 t + F_1
\end{aligned}$$

where

$$F_1 = a(E_1 \sin \omega_1 t + E_2 \sin \omega_2 t)$$

Since

$$\sin \omega_1 t \sin \omega_2 t = \frac{\cos(\omega_1-\omega_2)t}{2} - \frac{\cos(\omega_1+\omega_2)t}{2}$$

$$E_{\text{out}} = \frac{b}{2}(E_1^2+E_2^2) - \frac{b}{2}(E_1^2\cos 2\omega_1 t + E_2^2 \cos 2\omega_2 t) +$$
$$+ bE_1 E_2[\cos(\omega_1-\omega_2)t - \cos(\omega_1+\omega_2)t] + F_1.$$

From the above it is seen that the output consists of several terms, besides the first-order term. The significance of each is as follows:

(i) $\frac{1}{2}b(E_1^2+E_2^2)$ is a fixed term which can be ignored as it is not related to frequency.

(ii) $\frac{1}{2}b(E_2^2\cos 2\omega_1 t)$ is the 2nd harmonic of the input signal frequency, i.e. the input frequency $\times 2$, this being the *wanted* frequency.

(iii) $\frac{1}{2}b(E_2^2\cos 2\omega_2 t)$ is the 2nd harmonic of the input spurious frequency; this lies close to the wanted frequency, but its relative amplitude is greatly reduced compared with the input conditions. This ratio of input signal to the spurious input equals E_1/E_2 and ratio of the *wanted* 2nd harmonic signal to spurious 2nd harmonic is E_1^2/E_2^2.

(iv) $bE_1E_2\cos(\omega_1-\omega_2)t$ is the difference frequency. In the case of a frequency multiplier containing 1st and 2nd harmonics only, the difference frequency component lies well outside the operating band and therefore has little effect; however, some of the difference frequency components of higher harmonics fall into the wanted band and have significant effect as explained later.

(v) $bE_1E_2\cos(\omega_1+\omega_2)t$ is the sum frequency. ω_1 and ω_2 are close together and their sum lies close to the frequency we are considering, the actual separation being $2\omega_1-(\omega_1+\omega_2) = \omega_1-\omega_2$.

The amplitude coefficient of the 2nd harmonic term of the signal frequency is $\frac{1}{2}bE^2$.

The amplitude coefficient of the frequency summation term is bE_1E_2.

The ratio of these two amplitude terms is $E_1/2E_2$. But the ratio of the original signal amplitudes is E_1/E_2.

Thus at the multiplier output the required 2nd harmonic is produced as expected and also an unwanted frequency which is separated from the wanted one by the original frequency difference. Furthermore the amplitude of the output unwanted signals has increased by 6 dB relative to the input conditions.[2]

If the coefficient b should not be the same for the two terms because of the difference in signal levels, the 6 dB figure would be modified; experimental evidence, however, has confirmed this figure.

Analysing the expression further to include the cubic terms, i.e. 3rd harmonic content, we find the coefficients are:

3rd harmonic coefficient $cE_1^3/4$,

3rd order summation coefficient $3cE_1^2E_2/4$,

the ratio of these two terms becomes $E_1/3E_2$ for a tripler.

It is apparent that in this case the unwanted signal amplitude is enhanced by a factor of three, i.e. 9·5 dB. It can also be proved that the 4th order harmonic terms resolve down to $E_1/4E_2$, i.e. an enhancement of 12 dB for a quadrupler.

So far only summation terms have been considered to be significant spurious signals within the band. Difference terms, however, also fall within the required band when they exist as differences between the higher harmonics of the fundamental and the spurious input frequency or its harmonics.

In general the summation and difference frequencies are of the form $n\omega_1 \pm m\omega_2$: and these will appear respectively on either side of any one of the output harmonics of the fundamental.

The number of sum and difference terms increases in proportion to the number of harmonics the multiplier is capable of producing. For example, the frequency difference associated with the cubic term is $2\omega_1-\omega_2$ which results in a spurious frequency component separated from the main fundamental by the same amount as the spurious input, but appearing on the other side of the fundamental. Similarly, the spurious signals on either side of the 2nd harmonic are $\omega_1+\omega_2$ and $3\omega_1-\omega_2$, etc. (see Fig. 4).

The analysis has shown the amplitude coefficient of the sum and difference terms to be the same. Therefore the difference term frequency component will also be enhanced by 6 dB per octave of multiplication.

This note does not attempt a complete analysis of the non-linear characteristics of frequency multipliers but in general it can be stated that the unwanted frequencies in the pass-band of a multiplier increase by 6 dB per octave of multiplication, and the frequency difference remains the same. Also the higher the order of harmonics selected (i.e. the greater the multiplication factor) the greater the number of other frequencies produced from the original input spurious signals and these are separated by multiples of the original frequency differences.

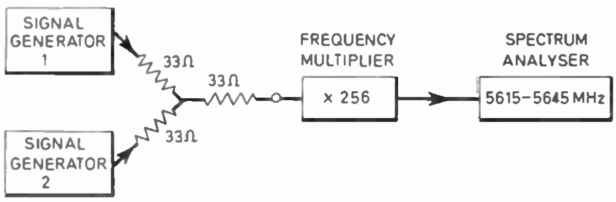
Further analysis also shows that slight changes of the ratio $E_1/2E_2$ are caused by contributions from the higher even-order terms, i.e. V_{in}^6, V_{in}^8, etc., should these exist in the multiplier.

3 Experimental Results

The following experimental results were obtained using two types of frequency multiplier. The first consisted of eight separate doubler stages, the output of each stage being filtered and the second harmonic only passed to the input of the succeeding stage. The first two of the eight stages were overdriven transistor amplifiers and the remaining six were varactor stages with mixed self and external bias.

The second type of multiplier was a single step-recovery diode capable of a multiplication of $\times 15$. The first three harmonics plus the fundamental only are shown in Fig. 4.

Reference to Fig. 1 shows the output frequency spectrum of an eight-octave (2^8) frequency multiplier



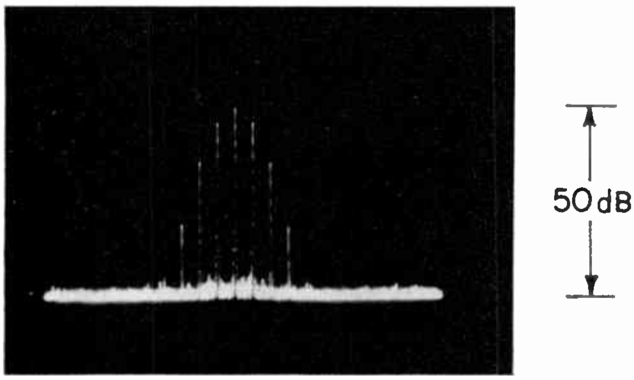


Fig. 1. Eight-octave multiplier fed with two signals.

Main signal in	22 MHz	Amplitude −6 dB
Spurious signal in	23 MHz	Amplitude −56 dB
Main signal out	5632 MHz	Amplitude +50 dB
Spurious signal out	5632 ± 1 MHz	Amplitude +48 dB

when fed with two frequencies as shown. As predicted by the previous analysis the spurious signal enhancement is 48 dB, i.e. 6 dB per octave. Some of the other spurious frequency components are also evident.

Reference to Fig. 2 shows the frequency multiplier output spectrum with no spurious input present. The absence of spurious signals in the output is evident.

There is, however, noise generated in the multiplier and Fig. 3 shows the multiplier output spectrum again without the spurious frequency input but with the amplitude level of the input reduced by 21 dB. The effect

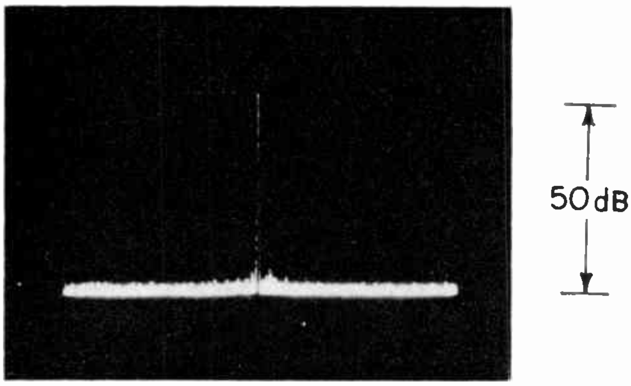


Fig. 2. Eight-octave multiplier without spurious signals input.

Main signal in	22 MHz	Amplitude −6 dB
Spurious signal in	Nil	
Main signal out	5632 MHz	Amplitude +50 dB
Spurious signal out	Nil	

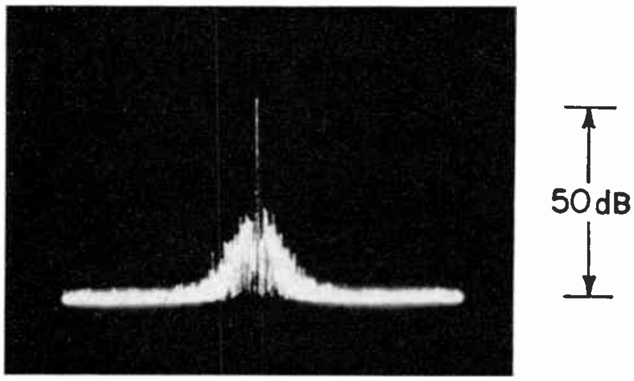


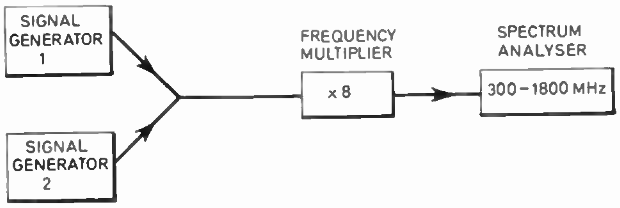
Fig. 3. Eight-octave multiplier showing enhancement of multiplier noise.

Main signal in	22 MHz	Amplitude −27 dB
Spurious signal in	nil	
Main signal out	5632 MHz	Amplitude +50 dB
Spurious signal out	Enhanced multiplier noise	Amplitude +20 dB

of this is to reduce the signal/noise ratio. Due to the 48 dB enhancement of the multiplier, the noise now appears as the spurious signals on the output spectrum.

Reference to Fig. 4 shows the output frequency spectrum of a step-recovery diode multiplier when fed with two frequencies as shown. Three of the main signal harmonics plus the fundamental, together with the spurious frequencies are shown. The main input frequency is 400 MHz and the input spurious signals 340 MHz.

The sum and difference term frequencies are seen on either side of the main harmonic frequencies and, from



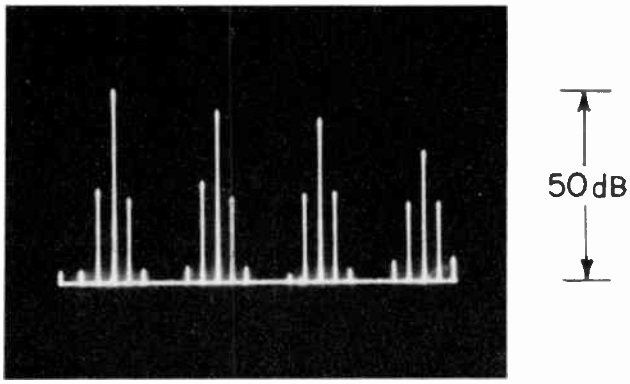


Fig. 4. Step recovery diode multiplier.

Main signal in	400 MHz	
Spurious signal in	340 MHz	Amplitude 25 dB below main signal
4th harmonic of main signal	1600 MHz	Amplitude +34 dB
Spurious signal accompanying 4th harmonic	1600 MHz ±60 MHz	Amplitude + 21 dB or 13 dB below main signal 4th harmonic

inspection, these clearly show an increase relative to the main harmonic they accompany, as the harmonic order increases. Measurement of the relative amplitudes indicates an approximate 6 dB per octave enhancement.

4 Conclusions

In general any form of frequency multiplier poses this problem of spurious signals enhancement. If the spurious signals lie outside the multiplier pass-band they can be filtered out but nothing can reduce the in-band unwanted signals. If a multiplier output spectrum of high purity is required it is necessary (i) to start with a very low level of input spurious signals, and (ii) to keep the order of multiplication as low as possible.

If the multiplier consists of low harmonic stages cascaded with inter-stage filters the spurious signals outside the wanted band are filtered out at each stage and therefore prevented from generating additional sum and difference terms which may appear as in-band spurious signals: such a system is believed to be preferable to one in which all the filtering is carried out at the multiplier output only.

Increasing use is being made of step-recovery diodes as frequency multipliers, the advantage being the efficiency with which they produce high-order harmonics, hence a high multiplication factor can be achieved in one device. The exploitation of this advantage precludes the use of inter-stage filters. Also many applications involving frequency multiplication require the multiplier to have considerable bandwidth (10–20%), in which case, of course, very high-order harmonics cannot be used with the step recovery diode. If any spurious signals lie within the required bandwidth, filtering is of no avail and it then becomes necessary to resort to either (i) or (ii) or both of the remedies mentioned above.

As an example a radar system may require a transmitter output spectrum with spurious signals at least 50 dB below the signal frequency level. If the signal frequency is to be derived from an 8-octave multiplier ($\times 256$), the level of the input spurious signals must be $50+(8\times 6\text{ dB}) = 98$ dB below the level of the main frequency.

Modern practice is to start the multiplier with as high a frequency as possible, but other precautions will become apparent in use such as the great care necessary in anti-shock mounting of the signal source particularly if this is a crystal.[1]

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A method of measuring negative impedance

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and

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SUMMARY

A method of measuring the components of a negative impedance has been developed. The device considered possesses an S-type negative impedance as opposed to an N-type, the tunnel diode being an example of the latter. The method of measurement is illustrated by measurements taken on a negative impedance booster used for compensating the losses of a wire transmission line. The technique of measurement is fairly simple and the accuracy of measurement depends upon the grade of standard components used for comparison and possibly also on the magnitudes of the negative impedance components to be measured. For values of a series positive resistance of 60 Ω and a parallel circuit composed of a negative resistance of 160 Ω and negative capacitance of 0·013 μF, measurements were possible with uncertainties not exceeding ±2·15%, ±0·75% and ±1·85% respectively using 1·0% grade capacitors and 0·1% grade decade resistors.

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List of symbols

C	capacitor, used in the test circuit, to provide a low impedance termination for the n.i.b. to make it oscillate
C_-	parallel negative capacitance component of the a.c. equivalent circuit of the negative impedance device
C1	capacitor, fixed, of large value, to make the device oscillate, in the first circuit configuration of measurement
C2	capacitor, variable, of known accuracy, to make the device oscillate in second circuit configuration of measurement
C_{-s}	equivalent series capacitance of C_- at a given frequency of oscillation
C_{-s_1}	equivalent series capacitance of C_- in parallel with R_- at frequency f_1
C_{-s_2}	equivalent series capacitance of C_- in parallel with R_- at frequency f_2
f_1	frequency of oscillation in circuit configuration 1
f_2	frequency of oscillation in circuit configuration 2
R_+	series positive resistance in the a.c. equivalent circuit of the negative impedance device
R_-	parallel negative resistance component of the a.c. equivalent circuit of the negative impedance device
R_{-s}	equivalent series resistance of the negative resistance R_- at a given frequency of oscillation
R_{-s_1}	equivalent series resistance of R_- in parallel with C_- at frequency f_1
R_{-s_2}	equivalent series resistance of R_- in parallel with C_- at frequency f_2
X_{f_1}	reactance of negative capacitance C_- at frequency f_1
X_{f_2}	reactance of negative capacitance C_- at frequency f_2

1 Introduction

Two types of negative impedance devices exist: the N-type which has a d.c. (I–V) characteristic similar to the letter 'N', e.g. the tunnel diode, and the S-type, an example of the latter being the negative impedance booster (n.i.b.).†

During the course of work on negative impedance boosters, hereafter referred to as the device or n.i.b., a method was developed to measure accurately the impedance components of the equivalent circuit of such a device. The method may be used to measure a negative impedance of any S-type device with suitable values of standard components used for comparison.

A basic circuit of the n.i.b. is shown in Fig. 1 and its a.c. equivalent circuit is shown in Fig. 2 where R_- and C_- have negative values. The effect of the capacitor C_0 at the terminals of the n.i.b. is to place an apparent

† Meacham, L. A., 'Negative impedance boosting', *Bell Syst. Tech. J.*, 47, pp. 1019–41, 1968.

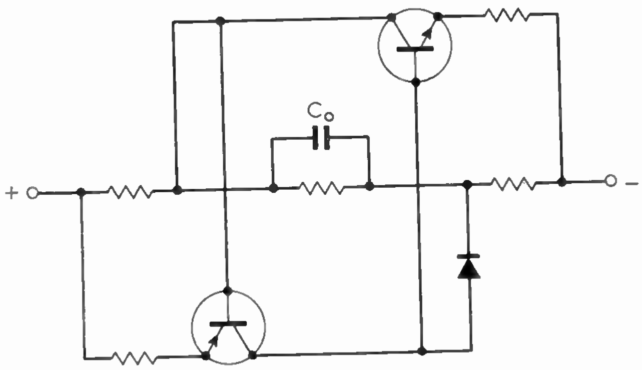

Fig. 1. Circuit diagram of a negative impedance booster.

negative capacitance C_- in parallel with a negative resistance R_-. The biasing resistors of the n.i.b. appear as a positive resistance, R_+.

The d.c. characteristic curve of the device has the shape shown in Fig. 3 and the portion of the characteristic which is of interest is the region of negative slope. This paper deals with the measurement of R_+, R_-, C_- and the slope resistance $(R_+ - R_-)$ in the region of negative slope.

2 Measurement Theory

An N-type negative impedance device is unstable when it is terminated by a high impedance. The S-type negative impedance device, on the other hand, is unstable when terminated by a low impedance. This instability manifests itself in the form of oscillations. By suitable adjustments of the component values of an external circuit connected to the n.i.b., the threshold of oscillation can be found.

The frequency of oscillation and the external component values are noted and based on the simple concept that on the threshold of oscillation the net resistance of the circuit is zero and the effective series reactance of the n.i.b. and the reactance of the added capacitor are equal in magnitude and opposite in sign, the values of the negative impedance components can be calculated.

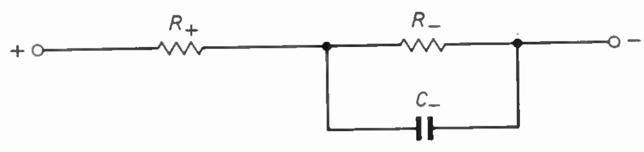

Fig. 2. Equivalent a.c. circuit of an n.i.b. on application of a suitable direct current.

These conditions are achieved in the arrangement shown in Fig. 4. The n.i.b. is fed from a constant current generator and the external terminating components R and C provide the low terminating impedance to make the n.i.b. oscillate. The oscillation may be started and sustained by the critical adjustment of R or C or both.

When the oscillations are just sustained (and the frequency of oscillation noted), the circuit of Fig. 4 may be represented by that shown in Fig. 5. R_{-s} and C_{-s} are the equivalent series values of R_- and C_- of Fig. 4 at the frequency of oscillation.

Since at the onset of oscillation the total resistance in the circuit of Fig. 5 is zero and the negative capacitance has the phase relationship (I–V) of an inductance, the latter in conjunction with C makes the circuit oscillatory.

Measurements are made under two conditions. In the first condition (referring to Fig. 4), C is made very large, C_1, to give a low frequency of oscillation, f_1, such that the impedance of C_- is very large compared with R_-. This value, C_1, must be large but is not critical.

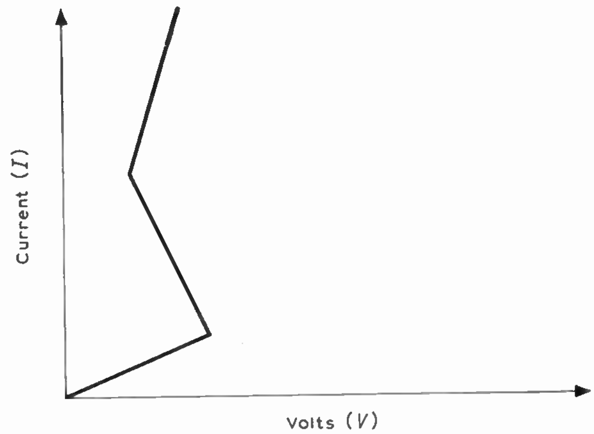

Fig. 3. The d.c. characteristic of a negative impedance booster.

Then, by applying the parallel to series transformation equation to the parallel negative components the equivalent series resistance R_{-s_1} is given by

$$R_{-s_1} = R_- \cdot \frac{X_{f_1}^2}{X_{f_1}^2 + R_-^2}$$

where

$$X_{f_1} = \frac{1}{2\pi f_1 C_-}$$

Provided that f_1 is sufficiently small,

$$X_{f_1} \gg R_-$$

whence

$$R_{-s_1} = R_-$$

R is now adjusted to a value R_1 so that oscillations are

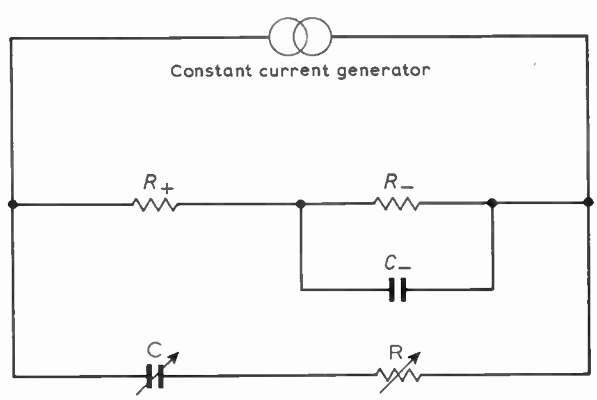

Fig. 4. Circuit showing the low impedance path to make the negative impedance device oscillate.

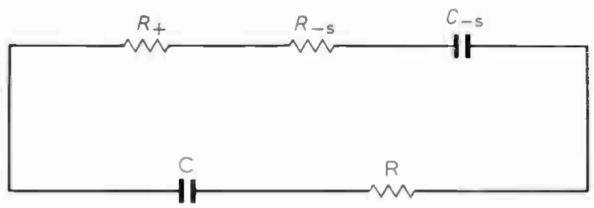

Fig. 5. Equivalent circuit of Fig. 4 at the onset of oscillation.

just sustained and we then have

$$R_1+R_++R_{-s_1}=0$$

Therefore

$$R_1+R_++R_-=0 \tag{1}$$

In the second condition (again referring to Fig. 4), R is made zero and C is adjusted to a value C_2 such that oscillations are just sustained. The frequency of the oscillation, f_2, is measured. Because the oscillations are just sustained the total series resistance must be zero. There is no external resistance and therefore the internal positive resistance R_+ must be equal in magnitude to the equivalent series resistance R_{-s_2} of the parallel negative impedance components, at the frequency of oscillation f_2.

Therefore we have,

$$R_{-s_2}=R_+=R_-\cdot\frac{X_{f_2}^2}{R_-^2+X_{f_2}^2}$$

where

$$X_{f_2}=\frac{1}{2\pi f_2 C_-}$$

Hence,

$$R_+=R_-\cdot\frac{X_{f_2}^2}{R_-^2+X_{f_2}^2}$$

Similarly, the reactance of the equivalent series capacitance, X_{-s_2}, of the parallel negative impedance components is given by

$$X_{-s_2}=X_{f_2}\cdot\frac{R_-^2}{R_-^2+X_{f_2}^2}$$

Since, at the frequency of oscillation

$$X_{-s_2}=X_{C_2}$$

where

$$X_{C_2}=\frac{1}{2\pi f C_2}$$

we have

$$X_{C_2}=X_{f_2}\cdot\frac{R_-^2}{R_-^2+X_{f_2}^2}$$

From the expressions above for R_+ and X_{C_2} we can transform them to

$$R_-=R_{-s_2}\cdot\frac{R_{-s_2}^2+X_{C_2}^2}{R_{-s_2}^2} \tag{2}$$

and

$$C_-=C_2\cdot\frac{X_{C_2}^2}{R_{-s_2}^2+X_{C_2}^2} \tag{3}$$

Manipulation of these equations is greatly simplified, by assuming a relation,

$$K=\frac{1}{(2\pi f_2 C_2 R_1)^2}$$

whence

$$X_{C_2}^2=K.R_1^2 \tag{4}$$

Substituting from equation (4) in equations (2) and (3) and manipulating these with equation (1) we obtain the following relations:

$$C_-=\frac{C_2}{K+1} \tag{5}$$

$$R_-=-R_1.(K+1) \tag{6}$$

$$R_+=R_1.K \tag{7}$$

we also have the slope resistance of the negative region (I–V) from equation (1),

$$R_+-R_-=-R_1 \tag{8}$$

3 Measurement Procedure and Evaluation of Errors

Figure 6 shows the arrangement for the measurement of the n.i.b. parameters. The value of the current from the constant current generator is set to enable the n.i.b. to be operated at the quiescent point at which measurements are to be made. An oscilloscope and electronic frequency counter with high input impedances are connected across the terminals of the n.i.b.

C1 is a low-loss capacitor whose value is large enough to give a low frequency of oscillation. R1 is a decade resistance box. C2 is a high-grade capacitance box; a fine control air capacitor may be included for added precision.

The measurements are carried out by selecting the two circuit conditions by means of the switch.

With switch in position 1 the decade resistance box is

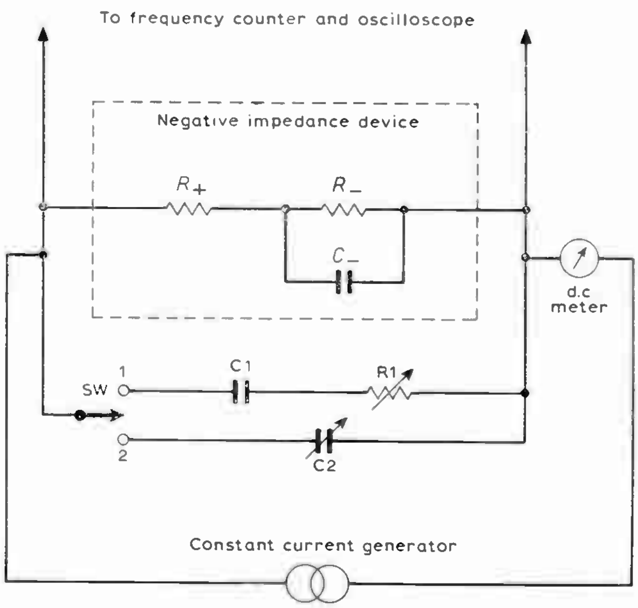

Fig. 6. Arrangement for measuring the negative impedance components.

adjusted until oscillations just commence (as detected on the oscilloscope). The value of R1 gives the value of the slope resistance of the 'negative' portion of the d.c. characteristic of the device. The uncertainty of determination on the onset of oscillation was less than 0·05 Ω with a device which had a slope resistance of about $-100\,\Omega$. Thus, using 0·10% grade decade resistance, it is possible to achieve an overall maximum systematic error of ±0·15% in the determination of the negative slope resistance.

It may be mentioned here that if the negative slope of the n.i.b. is perfectly linear, the amplitude, immediately after the onset of oscillation, will increase until one of the limits of dynamic swing of the n.i.b. is reached. In practice the slope is slightly non-linear and the sensitivity of the amplitude of oscillations to small changes of R1 gives an indication of the degree of linearity which can, if desired, be investigated closely by varying the quiescent point.

In the second circuit condition (switch position 2), the value of C2 is varied until oscillations just start. The value of C2 and the frequency f_2 measured by the counter are noted. The value of K is calculated and the values for C_-, R_- and R_+ can be determined from equations (5), (6) and (7).

In this case the uncertainty of the onset of oscillation was detectable to within ±20 pF on a nominal value of C2 of 0·08 μF. The error in the measurement of the frequency is negligible. Using a 1·0% grade decade capacitor and 0·1% decade resistance maximum systematic errors of the measurements were:

on the estimation of R_-: ±0·75%
on the estimation of C_-: ±1·85%
on the estimation of R_+: ±2·15%
on the estimation of slope resistance: ±0·15%.

4 Discussion

The values of the negative resistance and negative capacitance measured by the above method are very close to the mathematically predicted values for the circuit used for the device. The slope resistance has also been found to agree quite well with estimated performance. (It is virtually impossible to calculate the exact value due to the variability of the transistor parameters at the time of operation.)

Hence the assumptions made in this method of measurement may be assumed to be valid within the errors quoted.

It may be possible to measure the negative impedance components of an N-type device using a technique based on the philosophy of the method described above.

5 Acknowledgments

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On the design of equal ripple delay filters with Chebyshev stopband attenuation

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SUMMARY

This paper is concerned with the design of low-pass filters, without all-pass sections, approximating to a constant group-delay in an equal ripple manner which simultaneously exhibit a Chebyshev type of stopband attenuation. A special type of equal ripple delay approximation, referred to as the constrained Chebyshev approximation, is used to derive the polynomial of odd degree in the denominator of the rational transfer functions of these filters. Then, using the procedure, described by Temes and Gyi, the imaginary axis transmission zeros are determined so that the Chebyshev stopband attenuation of the resulting filter is obtained.

The steady-state and transient responses of these filters are discussed and shown to be superior when compared with those for the filters using Chebyshev delay approximants with the so-called standard error function, especially for larger values of the maximum delay deviation. Tables are also presented enabling direct determination of the odd-ordered approximants for minimum stopband attenuation.

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1 Introduction

The recent progress in communication systems has led to great interest in filtering networks exhibiting both good phase linearity and skirt selectivity of their magnitude response. Unfortunately, the linear phase characteristic can be obtained with polynomial filters only at the expense of a bell-shaped amplitude characteristic so that the maximally flat and equal ripple delay filters have neither pronounced passband nor stopband magnitude response. Hence, when a sharp cut-off filter with small passband magnitude distortion and a linear phase is required in a given application, such as, for example, is the case for filters used in systems for compatible speech and data transmission, it has been customary to design a filter with either a Chebyshev or an elliptic function amplitude by the usual methods and then to equalize the delay with additional all-pass sections. Using this technique it is virtually impossible to obtain a good phase linearity without resorting to a rather complicated phase equalizer which contains a comparable or even greater number of elements than that of the filter. Another method for designing Chebyshev filters having flat group delay characteristics in the main part of the passband has recently been described.[1] The transfer function of Chebyshev filters with complex transmission zeros is derived which are so determined that the group-delay characteristics may be flat in the main part of the passband.

On the other hand, Bennett[2] was first to suggest a method of synthesis in which a transfer function is determined to approximate a constant delay and then augmented by a numerator polynomial, which does not contribute to the delay, in order to improve the passband magnitude response. Allemandou[3] employed similar technique to derive a family of low-pass filters having both the time delay properties of Bessel filters and the attenuation properties of Butterworth filters.

However, the transmission zeros, i.e. the numerator polynomial, can also be used to improve the stopband performance of the linear phase filter. As is well known, the transmission zeros placed on the $j\omega$ axis do not influence the phase but completely suppress the output at the corresponding frequencies. Examples of this technique are the Unbehauen and Feistel methods for determining transfer functions of low-pass filters approximating a constant delay or a linear phase in the maximally flat sense[4] and in the Chebyshev sense[5] and exhibiting Chebyshev stopband attenuation characteristic. The main drawback of this technique is that the passband magnitude response is inadequate so that the time domain behaviour, i.e. the transient response to a unit step input, is rather poor especially if the maximum tolerable delay ripple is not very small.

In this paper a new class of filter functions of odd degree is presented having Chebyshev stopband attenuation and equal ripple delay response. The procedure used to obtain the polynomial in the numerator of the transfer function is similar to that used by Unbehauen[5]. However the polynomial in the denominator is determined by applying a special type of equal ripple delay approximation which will be referred to as constrained Chebyshev approximation. This type of delay approximation has recently been studied[6] and shown to provide better frequency and time domain characteristics than any other equal ripple delay approximant of odd degree. The investigations reported here have revealed that the constrained delay approximants are especially suitable for the construction of transfer functions with finite real frequency zeros since the latter help to reduce the ringing in the transient response to a unit step input.

This paper is organized as follows. First the approximation technique used to obtain a rational minimum phase transfer function which provides a constrained Chebyshev approximation of a constant delay and equal ripple stopband attenuation is briefly outlined. The second part is mainly concerned with the frequency and the time domain characteristics of the resulting filters and the comparison of the results with those obtained by other design methods. Tables are presented which include the zero and pole locations of some selected approximants of 3, 5, 7 and 9 degrees together with the most important frequency and time domain parameters of the filters.

2 Approximation Technique

We consider constructing the minimum phase transfer function of the form

$$F_{2n+1}(s) = \frac{M_{2n}(s)}{P_{2n+1}(A,s)} = \frac{\prod_{v=1}^{n}\left(1+\frac{s^2}{\omega_v^2}\right)}{\sum_{i=0}^{2n+1} A_i s^i} \qquad A_0 = 1 \quad (1)$$

where $P_{2n+1}(s)$ is a strictly Hurwitz polynomial of odd degree approximating a constant delay, $M_{2n}(s)$ an even polynomial with all zeros on the imaginary axis, $s = \sigma + j\omega$ is the complex frequency, and, for simplicity of notation, the set of A_i is denoted by A. Let the transmission coefficient be defined as

$$S_{12}(s) = (4R_1/R_2)^{1/2} K F_{2n+1}(s),$$

where K is a constant which for equal resistance terminations must be chosen such that $|S_{12}(0)| = 1$. For such a realization it is also necessary that $|S_{12}(0)| \geqslant S_{12}(j\omega)|$ from which it follows that $|F_{2n+1}(j\omega)| \leqslant 1$ for all ω.

The group delay of (1) is completely determined by the polynomial $P_{2n+1}(A,s)$ in the denominator since any pair of imaginary axis zeros lying symmetrically to the origin has no influence on the group delay because their contributions cancel. Hence, the phase and group delay responses of (1) are

$$\Phi(A,\omega) = -\arctan\frac{\mathrm{Im}\,(\omega)}{\mathrm{Re}\,(\omega)} \qquad (2)$$

$$D(A,\omega) = -\frac{d\Phi}{d\omega} = \frac{\mathrm{Im}'\,(\omega)\,\mathrm{Re}\,(\omega) - \mathrm{Im}\,(\omega)\,\mathrm{Re}'\,(\omega)}{\mathrm{Re}^2\,(\omega) + \mathrm{Im}^2\,(\omega)} \qquad (3)$$

where $\mathrm{Re}(\omega)$ and $\mathrm{Im}(\omega)$ are the real and imaginary parts of $P_{2n+1}(A,j\omega)$ respectively.

First we restate certain results discussed in Reference 6. If the error of the approximation to a constant group delay is stated in terms of a Chebyshev norm and the coefficients A_i are optimized so that the so-called standard error function results, the well-known equal ripple delay approximants of Abele[7] and Ulbrich and Piloty[8] are obtained. The error function of a given delay approximant of degree $2n+1$ is said to be a standard error function and to have a standard error curve if there are $2n+2$ extremal or critical points including the two endpoints of the approximation interval.

Now we impose one constraint at the origin by requiring that $|\varepsilon(0)| < \varepsilon$ where $\varepsilon(0)$ is the zero frequency delay error and ε is the maximum prescribed delay error in the approximation interval. Let $D_0 = 1$ be the ideal delay characteristic to be approximated, then, since $A_0 = 1$, the group delay of (1) at the origin is equal to A_1 so that we can write

$$A_1 = 1 + \varepsilon(0) \qquad (4)$$

To find the other unknown parameters A_2, A_3, ..., A_{2n+1} of the constrained approximant a set of non-linear equations is formed

$$D(A,\omega_i) = 1 \quad i = 2, 3, \ldots, (2n+1) \qquad (5)$$

where ω_i represent the zeros of the standard error curve (Abele's solution). The first zero of Abele's solution is disregarded because of the constraint at the origin (4). When the non-linear equations (5) are solved using the Newton–Raphson method an oscillatory error function is produced with unequal deviations between the fitting points ω_i. Finally, the perturbation method described by Crane,[9] or any other refinement technique can be used to equalize the maximum absolute errors between the matching frequencies.

Once the polynomial in the denominator of (1) has been found for any prescribed ε and $\varepsilon(0)$, the numerator polynomial can be determined using the procedure developed by Temes and Gyi[10] so that the equal ripple stopband attenuation is obtained.

Suppose the stopband attenuation equal to or higher than α_s(dB) is required for frequencies $\omega > \omega_s$ and consider the mapping

$$z^2 = s^2 + \omega_s^2 \qquad \mathrm{Re}\,(z) > 0 \qquad z = x + jy \qquad (6)$$

applied on the function $S_{12}(s).S_{12}(-s)$ which, for $s = j\omega$ equals the magnitude squared of $S_{12}(j\omega)$. The transformation (6) maps the passband portion of the $j\omega$ axis in the s plane on a portion of the real axis in the z plane. On the other side, it maps the stopband portion in the s plane over the whole imaginary axis in the z plane.

It can be shown that if the z-plane transformation of $S_{12}(s).S_{12}(-s)$ is given the form

$$H(z) = \frac{10^{-\alpha_s/10}}{1 + R(z)R(-z)} \qquad (7)$$

where $R(z) = zG(z)/E(z)$ is a z-plane reactance function, then, by transforming $H(z)$ back to the s plane and properly factoring it into $S_{2n+1}(s).S_{2n+1}(-s)$, the equal ripple stopband transmission coefficient is generated.

The numerical computation is very simple. Given the zeros of the constrained Chebyshev delay approximants

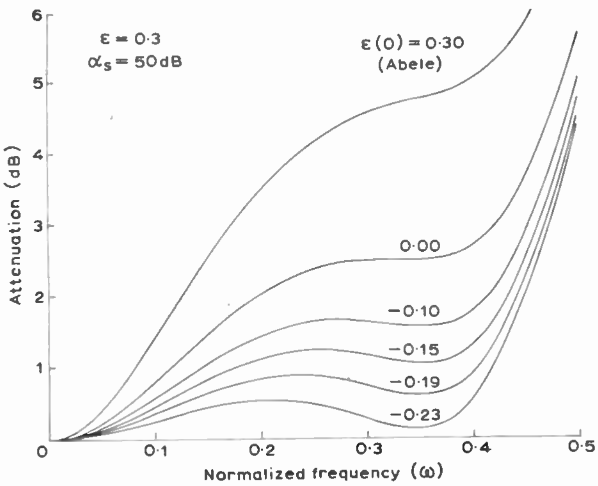

Fig. 1. Passband magnitude responses of the fifth-order filter for $\alpha_s = 50$ dB, $\varepsilon = 0{\cdot}3$ and different values of the zero frequency delay $\varepsilon(0)$.

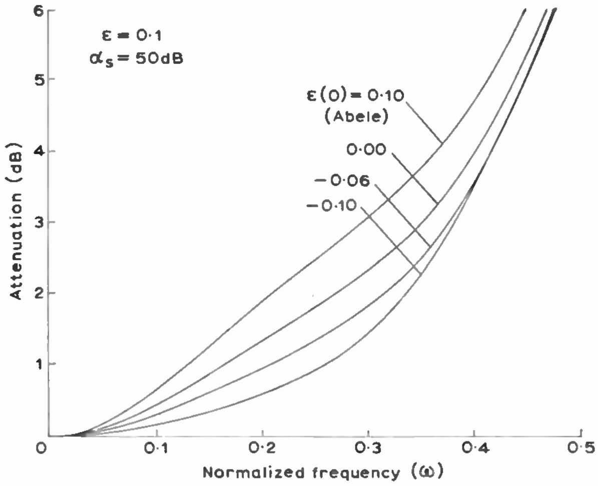

Fig. 2. Passband magnitude responses of the fifth-order filter for $\alpha_s = 50$ dB, $\varepsilon = 0{\cdot}1$ and different values of the zero frequency delay.

$s_i = \sigma_i \pm j\omega_i$ all we need is to find the even part $E(z)$ of

$$E(z)+zG(z) = \prod_{i=1}^{2n+1} (z_i+z) \tag{8}$$

where $z_i^2 = s_i^2+\omega_r^2$. Then

$$S_{2n+1}(s) = \frac{10^{\alpha_s/20} E(z)\big|_{z^2=s^2+\omega s^2}}{\prod_{i=1}^{2n+1} (s+s_i)} \tag{9}$$

For equal resistance terminations $|S_{12}(0)| = 1$ and using the normalization $\omega_s = 1$ we find from (7)

$$\alpha_s = 10 \log_{10} \left|1 - \frac{G^2(1)}{E^2(1)}\right| \tag{10}$$

Alternatively, an iterative technique may be used to find ω_s for a prescribed value of minimum stopband attenuation α_s.

3 Frequency and Time Domain Characteristics

3.1 *Frequency Responses*

As has been shown elsewhere,[6] the Chebyshev delay approximants with one constraint at the origin provide equal or even larger bandwidth of delay approximation than that obtained by the approximation based on the so-called standard error curve. This indicates that the solution obtained by Abele[7] and by Ulbrich and Piloty[8] is neither unique nor the best approximation. However, the increases in delay bandwidth are too small to be of any practical significance and we can assume that the bandwidth of delay approximation is virtually unaffected by the zero frequency delay error $\varepsilon(0)$, of course, provided that $|\varepsilon(0)| \leqslant \varepsilon$.

On the other hand, the value of $\varepsilon(0)$ has a considerable influence on the shape of the magnitude response both in the passband and in the stopband. In the odd-ordered case, for any prescribed maximum tolerated delay ripple ε decreasing $\varepsilon(0)$ from $\varepsilon(0) = \varepsilon$ towards $\varepsilon(0) = -\varepsilon$ decreases the passband magnitude distortion and increases the stopband attenuation. The computer analysis has revealed that even when the imaginary axis transmission zeros are added the same general relationship between the shape of the magnitude response and the value of $\varepsilon(0)$ is retained. This is illustrated in Fig. 1 in which the passband magnitude responses of the fifth-order filters are shown having Chebyshev stopband characteristics with minimum stopband attenuation of $\alpha_s = 50$ dB, the maximum tolerable delay distortion of 30% ($\varepsilon = 0{\cdot}3$) and different values of the zero frequency delay $\varepsilon(0)$. The normalization used is such that the 50 dB attenuation is first reached at $\omega_s = 1$. The characteristic for $\varepsilon(0) = 0{\cdot}3$ corresponds to the standard error curve of delay approximation (Abele's solution). It can be seen that as $\varepsilon(0)$ decreases towards its lower limit $\varepsilon(0) = -\varepsilon$ the passband magnitude response becomes non-monotonic and the $\omega_{3\text{ dB}}$ bandwidth is very much increased. For $\varepsilon(0) < -0{\cdot}25$ the condition $|F_{2n+1}(j\omega)| \leqslant 1$ is not fulfilled meaning that the realization with equal resistance terminations is not possible without the use of an ideal transformer.

As the maximum delay error decreases the hump in the passband magnitude characteristic tends to disappear

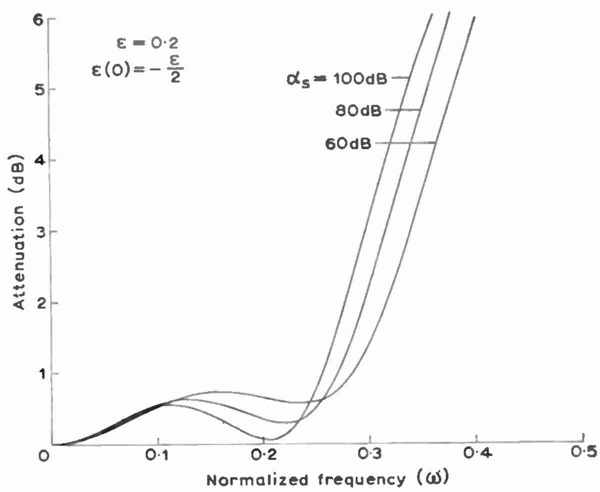

Fig. 3. Passband magnitude responses of the ninth-order filter for $\varepsilon = 0{\cdot}2$, $\varepsilon(0) = -\varepsilon/2$ and different values of the minimum stopband attenuation.

Table 1

2n+1	α_s(dB)	ε	$\varepsilon(0)$	ω_{3dB}	t_r	p_1%	p_2%
3	50	0.05	-0.03	0.206	11.07	1.84	1.04
5	50	0.05	-0.03	0.345	6.56	1.07	0.85
7	50	0.05	-0.03	0.350	6.47	0.93	0.89
9	50	0.05	-0.03	0.350	6.47	0.88	0.90

and the increase in the $\omega_{3\,\mathrm{dB}}$ bandwidth relative to the Abele case is greatly reduced but it still amounts to approximately 30% for $\varepsilon = 0{\cdot}1$. This can be seen from Fig. 2 in which the passband magnitude responses of the fifth-order filters for $\varepsilon = 0{\cdot}1$, $\alpha_s = 50$ dB are shown.

The shape of the magnitude response in the passband and the $\omega_{3\,\mathrm{dB}}$ bandwidth of the filter are also affected by the value of the minimum stopband attenuation. Increasing α_s decreases the $\omega_{3\,\mathrm{dB}}$ bandwidth. As an illustrative example in Fig. 3 the passband magnitude responses of the ninth-order filter are shown for $\varepsilon = 0{\cdot}2$, $\varepsilon(0) = -\varepsilon/2$ and $\alpha_s = 80$, 90 and 100 dB. Again, the normalization used is such that the minimum stopband attenuation is first reached at $\omega_s = 1$.

The bandwidth of delay approximation greatly exceeds the passband of the filter even for smaller values of the maximum delay distortion. As an example in Fig. 4 the magnitude and group delay responses of the fifth-order filter are shown for $\varepsilon = 0{\cdot}05$, $\varepsilon(0) = -\varepsilon$ and $\alpha_s = 40$ dB. Hence, in practical design the order of the filter is determined by the requirements imposed on the magnitude characteristic and/or the transient response. It has been found by computer analysis that for all ε and $\varepsilon(0)$ little can be gained both in respect of the $\omega_{3\,\mathrm{dB}}$ bandwidth and the transient behaviour from increasing the order of the filter above approximately $\alpha_s/10$. To illustrate this point in Table 1 the values of the $\omega_{3\,\mathrm{dB}}$ bandwidth, the 10–90% rise-time to a unit step input (t_r), the first overshoot (p_1%) and the first undershoot (p_2%) are summarized for $\varepsilon = 0{\cdot}05$, $\varepsilon(0) = -0{\cdot}03$, $\alpha_s = 50$ dB for $\omega_s = 1$ and $2n+1 = 3$, 5, 7 and 9. From this the following simple rule can be deduced for selecting the order of the filter: *Given the minimum stopband attenuation α_s in dB, the optimum order of the polynomial $P_{2n+1}(s)$ in the denominator of (1) is equal to the nearest odd integer to $\alpha_s/10$.*

3.2 *Transient Responses*

It has long been recognized[11] that the transient responses to a unit step input of all pole filters with equal ripple delay approximation are quite unsatisfactory unless the delay ripple is very small (1% or less). For filters of odd degree the output voltage oscillates below its final value for a relatively long time and then suddenly jumps to its final value. For example, in the odd-ordered case and $\varepsilon = 0{\cdot}1$, the leading edge of the output pulses is followed by a long lasting undershoot of approximately 5%. The addition of finite real transmission zeros does not improve the shape of the input pulses and the undershoot is even slightly increased. The situation is very much the same if, instead of using equal ripple delay approximant, the Chebyshev approximation of a linear phase characteristic is employed to construct the polynomial in the denominator of (1). As is well known, these two approximants are almost equal only if the maximum delay error is very small.

On the other hand, if the constrained Chebyshev approximants are used for the denominator polynomial $P_{2n+1}(s)$ in (1), the shape of the output pulses can be controlled by the zero frequency delay $\varepsilon(0)$ which acts as a free parameter. For any value of ε, if $\varepsilon(0)$ is decreased towards its lower limit $\varepsilon(0) = -\varepsilon$ the undershoot in the transient response is replaced by an increasing overshoot. Hence, by properly choosing ε and $\varepsilon(0)$ almost any prescribed specifications in respect of the maximum tolerable values of the overshoot and the undershoot can be met in practical design. As shown in the preceding section, decreasing $\varepsilon(0)$ increases the $\omega_{3\,\mathrm{dB}}$ bandwidth so that the 10–90% rise-time of the output pulses is also considerably improved when compared to that for the delay approximants having the standard error curve. In Figs. 5 and 6 the transient responses of the fifth-order

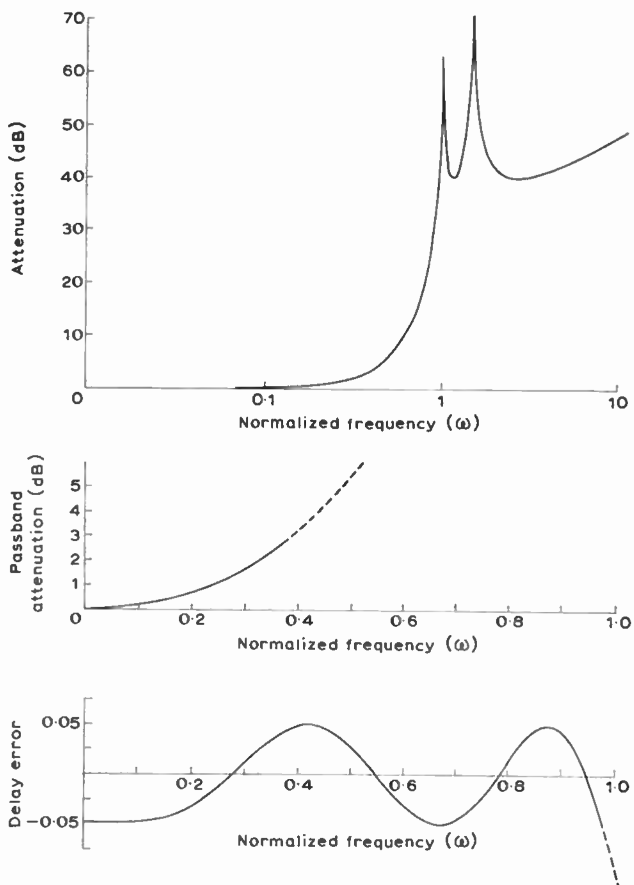

Fig. 4. Frequency domain characteristics of the fifth-order filter for $\varepsilon = 0{\cdot}05$, $\varepsilon(0) = -\varepsilon$ and $\alpha = 40$ dB.

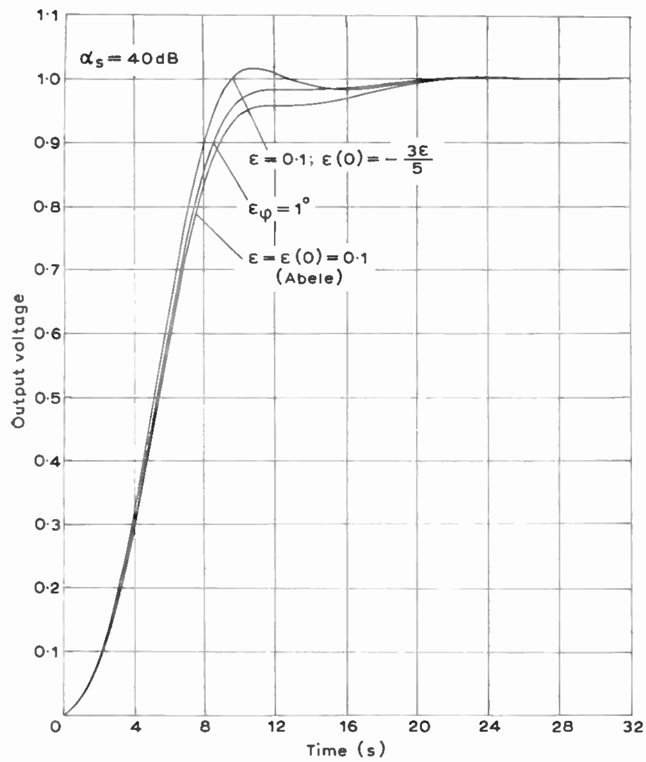

Fig. 5. Transient responses of the fifth-order filters with the minimum stopband attenuation $\alpha_s = 40$ dB for $\omega_s \geqslant 1$.

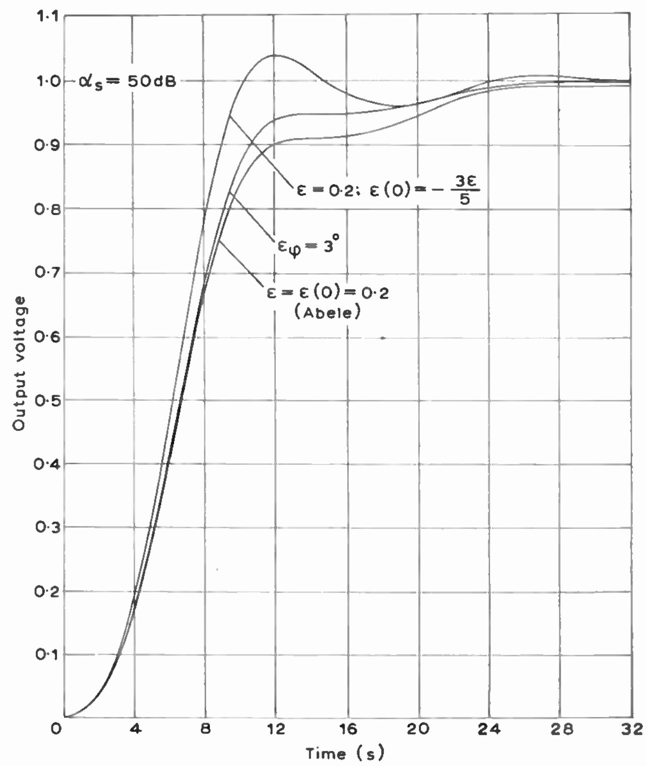

Fig. 6. Transient responses of the fifth-order filters with the minimum stopband attenuation $\alpha_s = 50$ dB for $\omega_s \geqslant 1$.

filters of the present design for $\varepsilon = 0{\cdot}1$ and $0{\cdot}2$, $\varepsilon(0) = -\frac{3}{5}\varepsilon$ and $\alpha_s = 40$ dB and 50 dB respectively are compared with those obtained by substituting Abele's solution or Chebyshev linear phase approximant[12] with the maximum phase error $\varepsilon_\phi = 3°$ for $P_{2n+1}(s)$ in (1).

For all values of ε and $\varepsilon(0)$ the 10–90% rise-time increases with increasing the maximum stopband attenuation α_s. This is due to the fact that the $\omega_{3\text{ dB}}$ bandwidth is reduced for higher stopband attenuation. Also the overshoot and the undershoot increase with increasing α_s. As an illustration of this in Fig. 7, the transient responses of ninth-order filters are shown with $\varepsilon = 0{\cdot}1$, $\varepsilon(0) = -\varepsilon/2$ and $\alpha_s = 40$, 80 and 100 dB. However, this presents no real difficulty in choosing the proper value of $\varepsilon(0)$ since, as shown in the preceding section, for optimum design the order of the filter is equal to the nearest integer to $\alpha_s/10$. It has been found by computer analysis that the best values of $\varepsilon(0)$ for all ε and n lie in the narrow range $-\frac{3}{5}\varepsilon \leqslant \varepsilon(0) \leqslant -\frac{1}{2}\varepsilon$.

The aforementioned results greatly simplify the construction of tables of transfer functions of odd degree that approximate a constant delay to within a given error ε and provide Chebyshev stopband attenuation (Tables 2–5). Included in these tables are the locations of the poles ($s_{pi} = \sigma_{pi} \pm j\omega_{pi}$) and the imaginary axis zeros $s_{0,i} = \pm j\omega_{0,i}$ of the transfer function, the $\omega_{3\text{ dB}}$ bandwidth, the 10–90% rise-time (t_r) of the transient response to a unit step input, the first overshoot (p_1%) and the first undershoot (p_2%). As before, the frequency scale is normalized so that the minimum stopband attenuation is first reached at $\omega_s = 1$. For prescribed values of ε and $\varepsilon(0)$ the pole locations are equal to within a constant multiplier representing the normalization factor. The

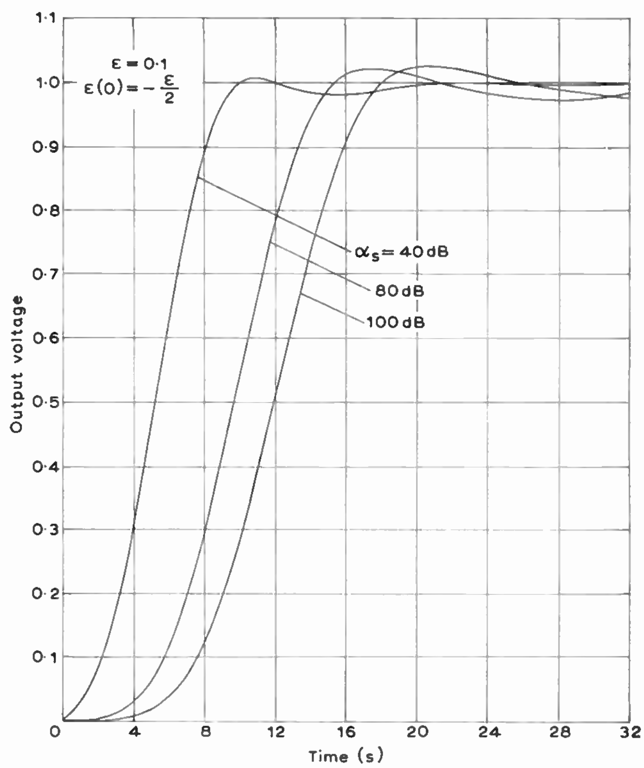

Fig. 7. Transient responses of the ninth-order filter for $\varepsilon = 0{\cdot}1$, $\varepsilon(0) = -\varepsilon/2$ and three different values of the minimum stop-band attenuation.

Table 2

ε	α_s (dB)	$\sigma_{po} \pm j\omega_{po}$	$\sigma_{p1} \pm j\omega_{p1}$	$\pm j\omega_{01}$	ω_{3dB}	t_r	p_1%	p_2%
0.05	20	-0.57131 +j0	-0.43243 ±j0.71522	±j1.16590	0.481	4.83	0.89	0.75
0.05	30	-0.38639 +j0	-0.29246 ±j0.48372	±j1.15193	0.386	5.95	1.29	0.90
0.05	40	-0.15218 +j0	-0.19987 ±j0.33058	±j1.15218	0.289	7.89	1.64	0.99
0.10	20	-0.35124 +j0	-0.37778 ±j0.74719	±j1.15246	0.496	4.72	0.99	1.36
0.10	30	-0.36386 +j0	-0.25875 ±j0.51178	±j1.14582	0.415	5.71	1.64	1.67
0.10	40	-0.25009 +j0	-0.17785 ±j0.35175	±j1.14944	0.317	7.47	2.20	1.85
015.	20	-0.50413 +j0	-0.33679 ±j0.76659	±j1.14221	0.507	4.65	1.04	2.00
0.15	30	-0.34871 +j0	-0.23296 ±j0.53026	±j1.14137	0.442	5.53	1.98	2.48
0.15	40	-0.24069 +j0	-0.16080 ±j0.36600	±j1.14790	0.342	7.16	2.76	2.78
0.20	20	-0.48287 +j0	-0.30307 ±j0.78031	±j1.13363	0.517	4.59	1.05	2.65
0.20	30	-0.33682 +j0	-0.21140 ±j0.54429	±j1.13777	0.471	5.40	2.30	3.36
0.20	40	-0.23330 +j0	-0.14643 ±j0.37700	±j1.14593	0.366	6.92	3.31	3.76

Table 3

ε	α_s (dB)	$\sigma_{po} \pm j\omega_{po}$	$\sigma_{p1} \pm j\omega_{p1}$	$\sigma_{p2} \pm j\omega_{p2}$	$\pm j\omega_{o1}$	$\pm j\omega_{o2}$	ω_{3dB}	t_r	p_1%	p_2%
0.05	40	-0.36517 +j0	-0.30467 ±j0.46787	-0.24905 ±j0.93087	±j1.04326	±j1.53152	0.380	5.92	0.87	0.78
0.05	50	-0.29659 +j0	-0.24744 ±j0.38000	-0.20227 ±j0.75603	±j1.04127	±j1.57305	0.345	6.56	1.07	0.85
0.05	60	-0.24056 +j0	-0.20070 ±j0.30821	-0.16406 ±j0.61322	±j1.04404	±j1.61557	0.303	7.46	1.31	0.91
0.10	40	-0.33138 +j0	-0.25380 ±j0.48200	-0.21513 ±j0.95834	±j1.04018	±j1.48889	0.406	5.77	1.60	1.67
0.10	50	-0.27122 +j0	-0.20772 ±j0.39450	-0.17608 ±j0.78437	±j1.03859	±j1.54462	0.375	6.31	2.07	1.82
0.10	60	-0.22120 +j0	-0.16941 ±j0.32174	-0.14360 ±j0.63970	±j1.04237	±j1.59786	0.332	7.11	2.53	1.94
0.15	40	-0.31106 +j0	-0.19136 ±j0.97172	-0.22026 ±j0.49022	±j1.03792	±j1.46028	0.435	5.63	2.33	2.65
0.15	50	-0.25605 +j0	-0.15751 ±j0.79986	-0.18131 ±j0.40351	±j1.03680	±j1.52641	0.403	6.11	3.09	2.89
0.15	60	-0.20960 +j0	-0.12894 ±j0.65477	-0.14842 ±j0.33032	±j1.04130	±j1.58675	0.356	6.84	3.79	3.07
0.20	40	-0.29629 +j0	-0.194744 ±j0.49615	-0.17242 ±j0.97908	±j1.03599	±j1.53809	0.464	5.52	3.02	3.70
0.20	50	-0.24506 +j0	-0.16106 ±j0.41036	-0.14261 ±j0.80979	±j1.03542	±j1.51298	0.428	5.94	4.08	4.04
0.20	60	-0.20119 +j0	-0.13223 ±j0.33689	-0.11708 ±j0.66481	±j1.04052	±j1.57872	0.375	6.61	5.03	4.29

Table 4

ε	α_s(dB)	$\sigma_{po} \pm j\omega_{po}$	$\sigma_{p1} \pm j\omega_{p1}$	$\sigma_{p2} \pm j\omega_{p2}$	$\sigma_{p3} \pm j\omega_{p3}$	$\pm j\omega_{o1}$	$\pm j\omega_{o2}$	$\pm j\omega_{o3}$	ω_{3dB}	t_r	p_1%	p_2%
0.05	60	-0.26261 +j0	-0.22466 ±j0.34854	-0.21092 ±j0.70694	-0.17518 ±j1.03276	±j1.02049	±j1.18517	±j1.86428	0.324	6.99	0.85	1.03
0.05	70	-0.22815 +j0	-0.19518 ±j0.30281	-0.18324 ±j0.61418	-0.15220 ±j0.89725	±j1.01760	±j1.19024	±j1.95477	0.303	7.45	1.01	1.08
0.05	80	-0.19872 +j0	-0.17001 ±j0.26375	-0.15961 ±j0.53495	-0.13256 ±j0.78151	±j1.01890	±j1.21058	±j2.04439	0.281	8.01	1.15	1.12
0.10	60	-0.23497 +j0	-0.18484 ±j0.35670	-0.17536 ±j0.72239	-0.15139 ±j1.05624	±j1.02027	±j1.17335	±j1.80713	0.350	6.77	1.77	2.15
0.10	70	-0.20476 +j0	-0.16107 ±j0.31084	-0.15281 ±j0.62950	-0.13192 ±j0.92042	±j1.01636	±j1.17721	±j1.90887	0.328	7.18	2.12	2.27
0.10	80	-0.17915 +j0	-0.14093 ±j0.27196	-0.13370 ±j0.55077	-0.11542 ±j0.80531	±j1.01790	±j1.20124	±j2.01156	0.303	7.67	2.44	2.36
0.15	60	-0.21884 +j0	-0.15973 ±j0.3626	-0.15258 ±j0.72995	-0.13506 ±j1.06753	±j1.02019	±j1.16566	±j1.77150	0.374	6.58	2.69	3.35
0.15	70	-0.19109 +j0	-0.13947 ±j0.31615	-0.13323 ±j0.63738	-0.11793 ±j0.93214	±j1.01549	±j1.16877	±j1.88072	0.348	6.95	3.26	3.54
0.15	80	-0.16771 +j0	-0.12241 ±j0.27746	-0.11693 ±j0.55938	-0.10350 ±j0.81808	±j1.01728	±j1.19551	±j1.99193	0.321	7.39	3.76	3.68
0.20	60	-0.20729 +j0	-0.14107 ±j0.36640	-0.12217 ±j1.07364	-0.13548 ±j0.73436	±j1.02014	±j1.15952	±j1.74519	0.394	6.42	3.57	4.62
0.20	70	-0.18131 +j0	-0.12339 ±j0.32048	-0.10686 ±j0.93908	-0.11850 ±j0.64232	±j1.01481	±j1.16240	±j1.86049	0.364	6.76	4.35	4.87
0.20	80	-0.15951 +j0	-0.10855 ±j0.28194	-0.09401 ±j0.82615	-0.10425 ±j0.56508	±j1.01682	±j1.19144	±j1.97821	0.334	7.16	5.05	5.08

Table 5

ε	α_s (dB)	$\sigma_{po}\pm j\omega_{po}$	$\sigma_{p1}\pm j\omega_{p1}$	$\sigma_{p2}\pm j\omega_{p2}$	$\sigma_{p3}\pm j\omega_{p3}$	$\sigma_{p4}\pm j\omega_{p4}$	$\pm j\omega_{o1}$	$\pm j\omega_{02}$	$\pm j\omega_{03}$	$\pm j\omega_{04}$	ω_{3dB}	t_r	p_1%	p_2%
0.05	80	-0.20694 +j0	-0.17740 ±j0.27637	-0.17064 ±j0.56465	-0.16270 ±j0.84078	-0.13536 ±j1.09279	±j1.01248	±j1.10528	±j1.32275	±j2.20595	0.287	7.87	1.10	1.15
0.05	90	-0.18555 +j0	-0.15906 ±j0.24780	-0.15301 ±j0.50630	-0.14588 ±j0.75388	-0.12137 ±j0.97984	±j1.00987	±j1.09335	±j1.34445	±j2.31979	0.271	8.27	1.24	1.19
0.05	100	-0.16760 +j0	-0.14368 ±j0.22383	-0.13821 ±j0.45732	-0.13177 ±j0.68096	-0.10963 ±j0.88506	±j1.00991	±j1.10117	±j1.38446	±j2.44182	0.256	8.68	1.36	1.23
0.10	80	-0.18468 +j0	-0.14527 ±j0.28194	-0.14030 ±j0.57437	-0.13536 ±j0.85532	-0.11710 ±j1.11364	±j1.01249	±j1.10444	±j1.30271	±j2.14333	0.309	7.57	2.37	2.40
0.10	90	-0.16563 +j0	-0.13029 ±j0.25286	-0.12583 ±j0.51513	-0.12140 ±j0.76711	-0.10502 ±j0.99878	±j1.00949	±j1.08728	±j1.32295	±j2.26361	0.290	7.94	2.66	2.49
0.10	100	-0.15005 +j0	-0.11803 ±j0.22907	-0.11399 ±j0.46667	-0.10998 ±j0.69494	-0.09514 ±j0.90482	±j1.00929	±j1.09534	±j1.36740	±j2.39732	0.274	8.30	2.94	2.57
0.15	80	-0.17197 +j0	-0.12536 ±j0.28598	-0.12146 ±j0.57945	-0.11798 ±j0.86203	-0.10466 ±j1.12359	±j1.01244	±j1.10410	±j1.29015	±j2.10609	0.326	7.31	3.66	3.74
0.15	90	-0.15425 +j0	-0.11244 ±j0.25651	-0.10894 ±j0.51974	-0.10583 ±j0.77320	-0.09387 ±j1.00782	±j1.00926	±j1.08333	±j1.30966	±j2.23038	0.305	7.67	4.13	3.87
0.15	100	-0.14002 +j0	-0.10207 ±j0.23285	-0.09890 ±j0.47180	-0.09606 ±j0.70187	-0.08521 ±j0.91484	±j1.00888	±j1.09171	±j1.35724	±j2.37147	0.287	7.98	4.55	3.99
0.20	80	-0.16299 +j0	-0.11070 ±j0.28948	-0.10755 ±j0.58270	-0.10497 ±j0.86554	-0.09486 ±j1.12887	±j1.01238	±j1.10380	±j1.28058	±j2.07971	0.340	7.10	4.93	5.14
0.20	90	-0.14622 +j0	-0.09930 ±j0.25969	-0.09648 ±j0.52273	-0.09417 ±j0.77646	-0.08510 ±j1.01269	±j1.00908	±j1.08019	±j1.29992	±j2.20707	0.316	7.43	5.57	5.32
0.20	100	-0.13294 +j0	-0.09029 ±j0.23611	-0.08772 ±j0.47528	-0.08562 ±j0.70597	-0.07737 ±j0.92076	±j1.00858	±j1.08908	±j1.35018	±j2.35388	0.297	7.72	6.15	5.47

transmission zeros depend on α_s and for each case the pole and zero locations for three different values of α_s are given (the 'optimum' value $\alpha_{s(opt)} = 10(2n+1)$ dB, and the lower and upper practical limits $(\alpha_{s(opt)} \pm 10)$ dB).

When compared with the results obtained by using Chebyshev delay approximants with the standard error curve, the improvements resulting from the use of the constrained Chebyshev approximants are higher for larger values of the maximum delay error. However, even for $\varepsilon = 0{\cdot}03$ a seemingly innocent variation in the zero frequency delay yields a noticeable improvement in both the rise-time and the transient ringing. For example, the transient response of the fifth-order approximant for $\varepsilon = \varepsilon(0) = 0{\cdot}03$ (the standard error curve) $\alpha_s = 50$ dB yields the 10-90% rise-time $t_r = 6{\cdot}96$, and the undershoot of 1·26% which is to be compared with the rise-time $t_r = 6{\cdot}69$ and the overshoot of 0·69% obtained when the constrained approximant $\varepsilon = 0{\cdot}03$ and $\varepsilon(0) = -\frac{3}{5}\varepsilon$ is substituted for $P_{2n+1}(s)$ in (1).

The practical realization of these filter functions is well-documented in the literature and need not be repeated here. Since transmission zeros lie on the finite $j\omega$ axis, zero-shifting technique has to be used and we may expect a simple ladder network realization. However, in some cases this technique demands the use of negative elements. These can always be realized with a set of perfectly coupled coils.[4]

4 Conclusion

The steady-state and transient responses of a new class of low-pass filter functions of odd degree have been studied and it has been shown that they approximate to a constant delay in an equal ripple manner and provide Chebyshev stopband attenuation of the resulting filter. The synthesis procedure is similar to that of Unbehauen.[5] However, a different approach is used to determine the denominator polynomial of the transfer function which forces an equal ripple approximation of a constant group delay with one constraint at the origin. It has been shown that, if the constraint imposed on the delay error function at the origin is such that the zero frequency delay is in the range $-\frac{3}{5}\varepsilon \leqslant \varepsilon(0) \leqslant -\frac{1}{2}\varepsilon$, the steady state and transient responses of the resulting filter are very much improved when compared to those described by Unbehauen, especially for larger values of the maximum delay deviation ε.

A limitation inherent in the procedure of Temes and Gyi, used to determine the numerator polynomial, that requires a special mention is that the odd-ordered transfer functions synthesized have only one zero at infinite frequency. This may be undesirable for some applications since the insertion loss of the filter approaches only 6 dB/octave in the stopband. Also, it is well known that for these transfer functions the amplitude of the impulse response is not zero at $t = 0$ but may amount to a rather large percentage of the peak value leading to a highly unsymmetrical impulse in the time domain. However, if necessary, this difficulty can easily be overcome by decreasing the order of the polynomial in the numerator of (1). Of course, in this case, instead of using the procedure of Temes and Gyi, an iterative technique must be employed to determine the locations of the imaginary axis transmission zeros.

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Failure-to-danger Potentialities of Solid-state Logic Control Systems

Over recent years there has been extremely rapid technical development of solid-state electronic control systems. All forms of these, whether consisting of discrete semiconductor devices, integrated circuits, or hybrids, can offer great advantages over electromagnetic relay control systems in speed of switching, signal voltage, and power consumption requirements. The risk of contact-welding, which must always be present with electromechanical systems, is avoided where static switching is used, and because the active components contain no moving parts, complete encapsulation can provide protection against hostile environments such as moisture, steam, dust, dirt and corrosive fumes, in which industrial control systems must often work.

The attractions of solid-state control systems are thus obvious, but the 'Annual Report of H.M. Inspector of Factories for 1971'* points out that where failure-to-safety is a prime consideration there are disadvantages which are not, perhaps, so immediately obvious, and which are not always appreciated by system designers, for there are two potential modes of failure in solid-state control which do not normally exist in electromagnetic systems.

The failure to-safety requirements in the control of dangerous machines and apparatus are basically that failure must not lead to uncovenanted starting of, or the inability to stop, dangerous motion. Where an electromagnetic relay is used as the active element in the control of a machine, failure of the relay itself cannot result in uncovenanted starting provided the system is arranged so that the coil has to be energized to close the contacts to give a 'go' signal. Further, in the 'run' condition, only failure of the contacts to open (due to contact welding or spring failure) can lead to inability to stop, since other modes of failure, i.e. open- or short-circuit coil fault and high contact resistance, give a 'stop' signal. Thus, a relay can be installed so that there is only one mode of failure of the relay itself which can lead to danger.

Where a semiconductor switch is used as the active element, the situation is very different. 'Go' signals can be produced not only by short-circuit failure of the device (and this is probably the commoner mode) but also by random momentary electrical interference impulses, for the high switching speed and low signal voltage requirement of the semiconductor switch ensure that it can respond to transient pulses which would have no effect whatsoever on the electromagnet of a relay. Thus, with semiconductor switching, there are two failure-to-danger possibilities which can produce uncovenanted starting and which do not exist with electro-magnet relays.

On this simplified basic analysis it is obvious that it is much more difficult to achieve the maximum degree of failure-to-safety with semiconductor control than with electromagnetic control. In practice, of course, the situation is more complicated A single solid-state logic switching system will normally contain a considerable number of semiconductor devices, and signal inversion between stages frequently occurs. This means that an output 'stop' condition may be represented by an active input to some stages, so loss of input due to open-circuit fault or failures of earlier stage outputs will give a 'go' output signal. Because normal design practice (to facilitate replacement of failed components) is to mount the assemblies of diodes, transistors, integrated circuits, etc., on plug-in printed circuit boards or other form of interchangeable module, loss of signal can be anticipated not only from open-circuit faults within the module but also from high resistance contact at the plug-in connexions.

Much can be done to guard against system malfunction from external interference by careful design of the system and the use of common mode rejection and propagation delay methods at the input interface, while majority logic, feedback, and cross-monitoring techniques can provide safeguards against failure in intermediate stages. But analysis by Electrical Inspectors of systems which have had to be considered in detail following accidents and potentially dangerous incidents, or which have been put forward for consideration in proposed safety control schemes, suggests that, as it is virtually impossible to guard against failure of the ultimate control element by the above techniques, it is almost always preferable to employ an overriding back-up electromagnetic, mechanical, hydraulic, or pneumatic safety loop.

The case with which such ultimate protection can be provided can be illustrated by a single example. Positive interlocking of the door of an X-ray enclosure was required with an X-ray apparatus using a solid-state logic switching control system. The designer proposed that the door interlock switch would be connected into one of the logic system inputs. Theoretical analysis postulated many failure-to-danger possibilities in the system which could result in failure to shut off the X-rays on opening the door, or the sudden switching on of the X-rays with the door open. It was pointed out to the user that a complete back-up against failure, without any alteration or elaboration of the solid-state logic system, could easily be provided by the simple addition of a second limit switch connected directly into the wiring to the primary winding of the high voltage X-ray transformer. The switch would be mounted at the enclosure door so that the opening of the door operated first the logic control switch to initiate the correct shut-down sequence (which occupied only a fraction of a second) and then the positively-operated back-up switch. On closing the door the sequence was, of course, reversed. Thus the back-up safety switch would not normally be called upon either to make or to break load current, and would be called into play only on a failure of the logic system which would lead to X-rays being generated with the door open without this simple back-up.

In his general comments presenting the Report to the Secretary of State for Employment, the Chief Inspector of Factories, Mr. Bryan Harvey, points out that time and money must be lavished by industry on protecting employees and the community in the same way that it is lavished upon improving industrial techniques and developing new ways of manufacturing. If industry as a whole were to accept the need for this kind of forward planning, it would soon find that the sort of techniques used to anticipate production problems were equally valid to anticipate the hygiene and safety problems which a new process might create. Computers should increasingly be used on long-range safety and health control problems. If it was not possible to develop adequate measures of controlling hazards then industry might have to take a decision not to develop a particular plant or process until the way ahead for both workers and the environment was clear.

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An experimental adaptive echo canceller for long distance telephone circuits

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SUMMARY

A new concept of eliminating spurious echoes in long distance telephone circuits by adaptive echo cancellation has been experimentally implemented. In this paper the principles underlying the experimental device and the techniques used in their implementation are presented. Experimental results obtained from measurements in the laboratory as well as in a practical telephone system environment are reported.

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1 Introduction

Telephone connexions extending beyond the local distribution network employ two essentially different transmission techniques. In the local network a single wire pair serves as a bi-directional transmission medium ('two-wire network'), whereas beyond, voice signals are carried in various ways on a separate channel in each direction ('four-wire network'), (Fig. 1(a)). On every transition between the four-wire and two-wire part of the network a 'hybrid' transformer is inserted to serve as an interface. This can be conceived as a bridge network whereby the two four-wire terminals are connected across the two bridge diagonals; the two-wire terminal is connected to one arm of the bridge and the bridge is balanced by a 'balancing network' in the arm opposite to the two-wire terminal. Thus, if the bridge is in balance, the two directions of the four-wire network are de-coupled.

In practice, however, the input impedance of the two-wire circuit is found to vary due to a number of reasons. These variations may occur from connexion to connexion as well as a function of time after a connexion has been made. Thus with a fixed balancing network the bridge may (as a rule) be more or less unbalanced and voice signals arriving at the hybrid through the 'receive' channel of the four-wire network may be coupled into the 'send' channel and return to the signal source as 'echo', with a delay depending on the distance between the source and the four-wire/two-wire junction and on the propagation velocity constant of the transmission medium.

The effects of such echoes may range from instability of the circuit to annoying, intelligible repetitions of the spoken words. The latter condition obtains whenever round trip circuit delays exceed, say, about 50 ms, whereby the annoyance effect depends on echo amplitude and delay and may cause severe disturbances to the flow of conversation between the two speakers. It is the solution of this problem with which the investigation reported in this paper are concerned.

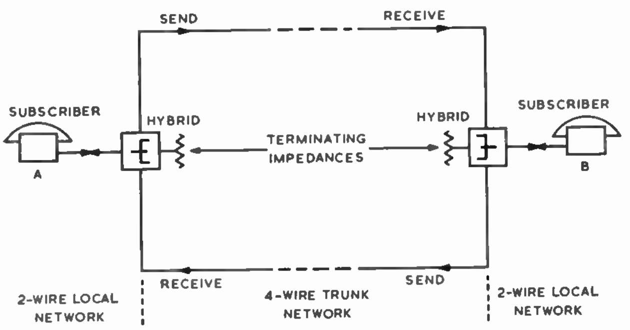
It will become apparent to the reader that the solution of this telecommunication problem was significantly dependent on the solving, in the first instance, of typical control problems related to the identification and control of time variant system variables. Thereby the conditions of residual error and speed of control imposed by the telecommunication system's requirements were governing factors for the design and implementation of the experimental control system.

2 Echo Suppression and Echo Cancellation

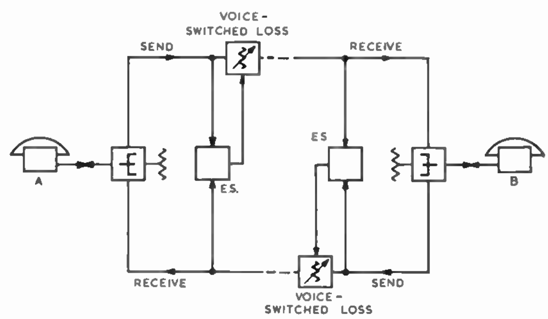
Until now the problem of eliminating disturbing echoes from long distance telephone connexions has been practically solved by installing what are known as 'echo suppressors' in appropriate locations of national and international telephone networks. These devices are based on switching operations by which (primitively described) the echo return path, that is, the send channel, of the four-wire circuit is either interrupted or heavily attenuated, whenever a sensing circuit incorporated in the device detects a signal in the receive channel which has a higher level than that in the send channel. Numerous variations of this principle have been implemented and are in operation all over the globe (Fig. 1(b)).

It is obvious that difficulties can arise under conditions of 'double talk', that is, when both partners in a conversation speak simultaneously. These manifest themselves as speech clipping and loss of syllables. This condition has become especially critical since the introduction of geo-stationary communication satellites for which the round trip signal delay is in the order of half a second. It is therefore not surprising that about the time of the introduction of these satellites in the mid-1960s new thought was given to the solving of the echo problem.

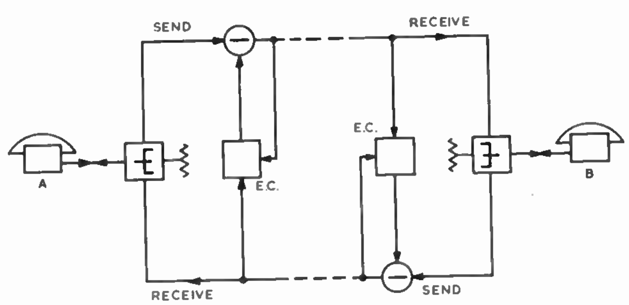
Logically, the most fundamental solutions would be either to operate all telephone circuits in the four-wire mode or at least to devise a truly self-balancing hybrid network. Both these solutions, however, are at present considered economically impracticable and thus a new concept was evolved which became known as 'echo-cancellation' (Fig. 1(c)).



(a) Simplified 2-wire/4-wire long-distance connexion.



(b) Long-distance connexion with echo suppressors.



(c) Long-distance connexion with echo cancellers.

Fig. 1. Long-distance telephone connexions.

In contrast to echo-suppression, this new concept suggested that at the same points of the telephone network, where previously echo-suppressors were installed, a device be provided which would synthesize a signal identical to the echo and inject this synthesized signal into the send channel with opposite polarity to that of the echo and thus the echo would be cancelled. Consequently its operation would not interfere with the useful signals in the send and receive channels and thus the disturbing effects of the echo-suppressor would be eliminated. This echo synthesis would of course require to be adaptive in order to match the variations of the echo signal resulting from the variations of the input impedance to the two-wire circuit connected to the hybrid network. This new concept was evolved almost simultaneously and independently by different people and in a variety of organizations.[1-7]

This paper refers to the development of an experimental echo canceller in the A.P.O. Research Laboratories based on the original conceptual development by R. B. Zmood[1] which, however, differs little from that published by others about the same time.

Any such device will have to perform three basic functions:

(i) identify the echo signal;
(ii) derive control criteria for synthesis;
(iii) synthesize the cancelling signal.

It will be shown that functions (i) and (iii) can be reduced to the identification and synthesis of impulse responses. All reported proposals consider the synthesis function to be performed by means of a transversal filter which is implemented in a tapped delay line configuration; thereby the only differences of approach noted arise from the realization of the cancelling system in either digital or analogue techniques.

More basic differences are apparent in the identification function. Here the alternatives are to choose either the actual voice signals or specific test signals (test pulses or pseudo-random noise (p.r.n.) waveforms) as excitation from which the control information is to be derived.

The choice of the identification signal largely determines the complexity of the control algorithm, that is, the function (ii), and its implementation and consequently, because of its simpler realization, the preferred choice so far tended towards specific test signals. However, as will be shown from practical results obtained with the device described in this paper, there are definite limitations in this method which strongly point to the need to use the actual voice signals, either alone or in combination with specific test signals, for the purpose of systems identification and control.

3 Identification and Control Algorithms

The experimental echo canceller described in this paper has been implemented in analogue circuit techniques and uses bursts of p.r.n. waveforms as test signal to provide the control information for the echo synthesis.[8] Cross-correlation analysis of the error signal after echo cancellation is used to derive this information and to compute the coefficients of the transversal filter, that is, the tap-coefficients of the tapped delay line.

These operations are based on the following algorithms:

If $s(t)$ and $e(t)$ denote the four-wire input and output signals of the hybrid network, respectively, and the trans-hybrid impulse response is $h(t)$, the echo signal $e(t)$ is given by the convolution

$$e(t) = s(t)^{*}h(t). \tag{1a}$$

Similarly the cancelling signal to be synthesized is represented (Fig. 2) by

$$\hat{e}(t) = s(t)^{*}\hat{h}(t) \tag{1b}$$

and thus the error signal, that is, the residual echo, is given by

$$\varepsilon(t) = e(t) - \hat{e}(t). \tag{2}$$

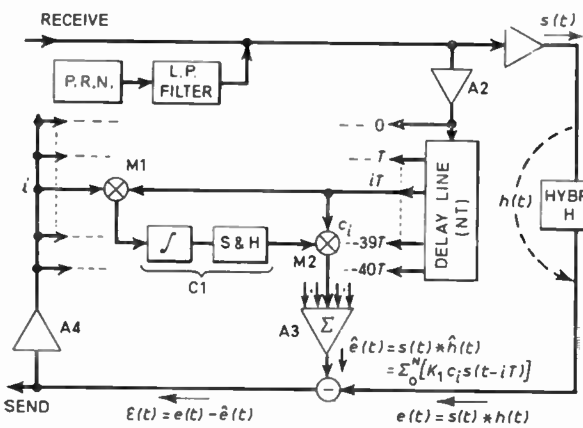


Fig. 2. Schematic circuit of adaptive echo canceller.

As implicit in the realization of the synthesized impulse response, $\hat{h}(t)$, by means of a tapped delay line, the cancelling signal is more realistically represented by the time series

$$\hat{e}(t) = \sum_{i=0}^{N} K_1 c_i s(t-iT) \tag{3a}$$

where T is the delay difference between adjacent taps of the delay line; this elementary delay is made equal to the Nyquist interval appropriate to the signal, $s(t)$, that is, by its usual definition $T = 1/2f_m$ where f_m is the maximum frequency component which may occur in $s(t)$. The time series is limited to $N+1$ members, $\{c_i\}$ denotes the set of coefficients of this time series, and $K_1 = k_{M2}.k_{A2}.k_{A3}$, that is, the product of the fixed gains of the circuit elements M2, A2, A3 as shown in Fig. 2.

Representing $e(t)$ in discrete time series notation also, equation (1a) becomes

$$e(t) = \sum_{i=0}^{N} h_i s(t-iT) \tag{3b}$$

where $\{h_i\}$ now represents the time series coefficients of $h(t)$ and thus

$$\varepsilon(t) = \sum_{i=0}^{N} s(t-iT)\,(h_i - K_1 c_i). \tag{4}$$

The identification and control function of the echo canceller in terms of equation (4) is then to control the coefficients, c_i, in such a way that $(h_i - K_1 c_i)$ converges to zero.

For this purpose the error signal, $\varepsilon(t)$, is cross-correlated with the input signal, $s(t)$, to obtain the cross-correlation coefficients, m_i, associated with each time series coefficient as

$$m_{ij} = \frac{K_2}{T_1}\int_{(j-1)T_1}^{jT_1} \varepsilon(t)s(t-iT)\,dt \tag{5}$$

where T_1 is a limited integration time, $j(=1, 2, 3 \ldots M)$ denotes successive iterations of the cross-correlation operation and $K_2(=k_{A2}.k_{A4}.k_{M1}.k_{C1})$ represents the product of the gains of the corresponding circuit elements.

After each iteration the coefficients, c_i, are updated such that

$$c_{ij} = c_{i,(j-1)} + m_{ij}. \tag{6}$$

It is now necessary to determine the control loop gain as given by the gains K_1 and K_2.

Ideally convergence should be reached after the first iteration, $j = 1$. Then

$$m_{i1} = \frac{h_i}{K_1} - c_{i0} \tag{7}$$

where c_{i0} is assumed to be the initial magnitude of c_i. Also from equation (5), with $\varepsilon(t)$ from equation (4), we obtain

$$m_{i1} = K_2(h_i - K_1 c_{i0})\,\frac{1}{T_1}\int_0^{T_1} s^2(t-iT)\,dt. \tag{8}$$

The term following the bracket on the right-hand side of equation (8) will be recognized as the mean power $\overline{(s^2)}$ of $s(t)$. Then by equating equations (7) and (8) and reordering terms we obtain the following equation:

$$h_i\left[\frac{1}{K_1} - K_2\overline{s^2}\right] = K_1 c_{i0}\left[\frac{1}{K_1} - K_2\overline{s^2}\right]. \tag{9}$$

The first solution of this equation

$$h_i = K_1 c_{i0}$$

is a trivial one simply stating that the initial value of c_i was already the correct one.

The second solution of equation (9) obtains if

$$\frac{1}{K_1} - K_2\overline{s^2} = 0$$

from which follows

$$K_1 K_2 \overline{s^2} = 1.$$

Since $s(t)$ is for the present design chosen to be a p.r.n. test waveform, $\overline{s^2}$ is constant and thus K_1 and K_2 are determined by design of the hardware. However, it can be asserted that the ideal situation of achieving $\hat{h}(t)$ to be equal to $h(t)$ and thus $\varepsilon(t)$ to be equal to zero in one iteration, cannot be realized in practice. Moreover, an analytical analysis of the real conditions seems rather difficult and even fruitless. It must therefore be accepted that, operationally, a certain number of iterations of the control process will be necessary and that the residual echo cannot be made exactly zero.

An essential task of the implementation was therefore to optimize the hardware design such that an acceptably small $\varepsilon(t)$ was obtained with a minimum number of iterations, that is, within a minimum convergence time. The following sections of this paper will deal in detail with this task and with the results obtained experimentally.

4 System Implementation

From the outline of the system functions of the echo canceller presented in the previous section, it is obvious that the degree of cancellation attainable is directly governed by the precision with which the transversal filter output signal, $\hat{e}(t)$, matches the trans-hybrid echo, $e(t)$. To achieve a cancellation performance which would render the echo signal after cancellation imperceptible during normal conversation and within a convergence time of the control process of less than 0·5 seconds, will require the hardware components of the system (Fig. 2) to satisfy the following design guide-lines:

(i) the delay line should be as non-dispersive in amplitude and phase versus frequency as possible over the frequency range of normal voice frequency signals, that is, 200 to 3400 Hz;

(ii) the coefficient multipliers (M2) should be highly linear at least with respect to the signal input;

(iii) the correlation multipliers (M1) should be highly linear with respect to both inputs. Their dynamic range should be fully utilized and any drift of their output should be sufficiently small to minimize integration error;

(iv) the integration and sample hold circuit (C1) operating in conjunction with the correlation multipliers to compute and store the cross-correlation coefficients (m_i) and time series coefficients (c_i) should have negligible offset voltage;

(v) the frequency spectrum of the p.r.n. test signal should sharply cut off at the frequency corresponding to the Nyquist interval of the time series representing $\hat{e}(t)$ in equation (3), that is, the delay interval between adjacent taps of the delay line;

(vi) the loop gain should be ideally unity but provision for adjustment should be made to compensate for the non-ideal nature of the system, keeping in mind that its magnitude must be less than 2 to maintain convergence of the control process.

The circuitry used to achieve these requirements is now described in detail.

4.1 *The Delay Line*

A commercially available, passive, analogue delay line with 40 taps at 150 µs spacing was used. It is highly linear but has a non-ideal group delay characteristic at some taps. The delay line tap spacing of 105 µs is slightly more than that required to match the nominal 3·4 kHz bandwidth of voice signals but had to be accepted for reasons of ready availability of a suitable device. These shortcomings do not significantly affect the results obtained.

4.2 *The Multipliers*

The multiplier used for the correlations, M1, and transversal filter controls, M2, is described in references 9 and 10. In fact the coefficient multiplier, M2, could have been less sophisticated but it was convenient in the design of the system to use the one type of multiplier. This multiplier is of novel design and its particular feature is that a greater linear range can be obtained than with any comparable known linear circuit. This property is essential to enable the correlation to be done with sufficient accuracy and the successful multiplier development was a major factor in deciding that the hardware would be based on analogue circuitry. The operation of the multiplier will be briefly described, with reference to Fig. 3.

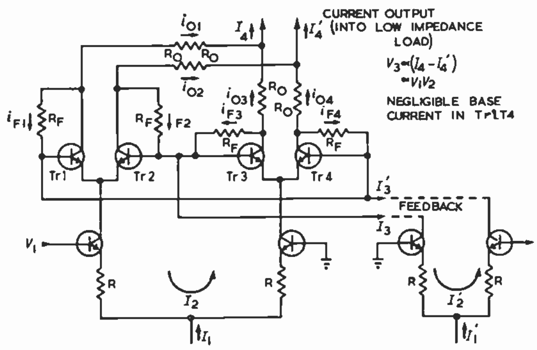


Fig. 3. Analogue multiplier circuit.

The balanced current sources, I_2 and I_2, are controlled by the input voltages, V_1 and V_2, such that

$$\begin{aligned} I_2 &= V_1/2R \\ I_2' &= -V_2/2R. \end{aligned} \tag{10}$$

The current, I_2, is applied between the emitters of two matched transistor pairs, Tr1, Tr2 and Tr3, Tr4. With simple feedback applied via R_F and with the outputs feeding into a signal earth via R_0, the high current gain of the transistors causes the ratio between the currents i_{0n} and i_{Fn} to be approximately

$$\frac{i_{0n}}{i_{Fn}} \simeq \frac{R_F}{R_0}. \tag{11}$$

With the corresponding bases of Tr2, Tr3 and Tr1, Tr4 connected together the voltage between the bases of Tr1 and Tr2 is in practice very small and is equal to that between the bases of Tr3 and Tr4. Therefore with well matched transistors it is a good approximation that

$$\frac{i_{02}}{i_{01}} \simeq \frac{i_{03}}{i_{04}} = \phi. \tag{12}$$

Using equations (11) and (12) the constant current, I_1, is related to the output currents by

$$I_1 \simeq \left(1+\frac{R_0}{R_F}\right).(\phi+1).(i_{01}+i_{04}). \tag{13}$$

The feedback is connected so that the current $(I_3 - I_2')$ is forced to be equal to the second input current, I_2'. Therefore by using equations (10) to (13)

$$I_2 \simeq \sum_{n=3}^{4} (i_{0n}+i_{Fn}) - \sum_{n=1}^{2} (i_{0n}+i_{Fn})$$
$$\simeq I_1 \left[\frac{i_{04}-i_{01}}{i_{04}+i_{01}}\right] \quad (14)$$

and

$$I'_2 = I_3 - I'_3$$
$$= i_{F2}+i_{F3}-i_{F1}-i_{F4}$$
$$\simeq \frac{R_0}{R_F}(\phi-1).(i_{04}+i_{01}). \quad (15)$$

The output current from the circuit is taken as $(I_4 - I_4)$ and

$$I_4 - I'_4 = i_{01}+i_{03}-i_{02}-i_{04}$$
$$\simeq (i_{04}-i_{01})(\phi-1). \quad (16)$$

From equations (10) and (14) to (16) it is easily shown that

$$(I_4 - I'_4) \simeq I_2 . I'_2 \frac{R_F}{R_0 I_1}$$
$$\simeq \frac{-R_F}{4R_0 . R^2 I_1} . V_1 V_2$$
$$= B . V_1 V_2. \quad (17)$$

The multiplication of the two input variables is therefore achieved.

The static linearity of the practical multiplier is shown in Fig. 4. With careful selection of transistor pairs the deviation from linearity over the dynamic range is better than 0·5%. Output drift was measured with both inputs grounded and was of the order of 0·1% of the maximum linear output. The bandwidth of the practical multiplier was found to be above 100 kHz for both inputs; however this could be increased by reducing the resistance values R_0 and R_F and by using transistors with higher cut-off frequencies. The r.m.s. noise measured over a bandwidth of 100 kHz at the output was found to be about 90 dB below the value of the maximum linear output.

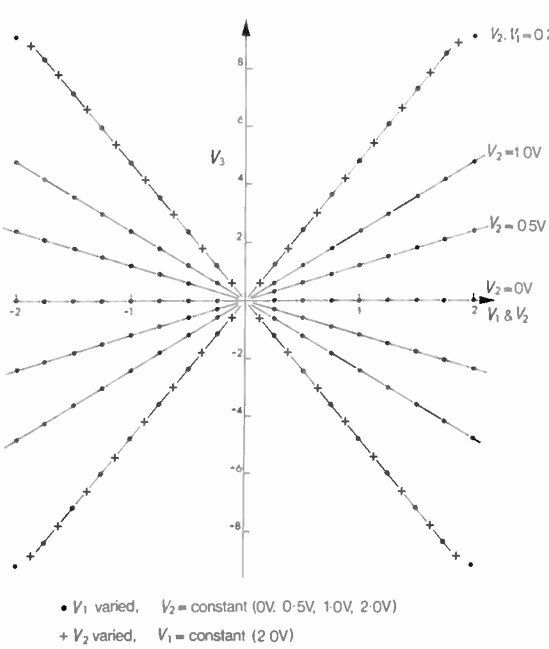


Fig. 4. Static linearity of analogue multiplier.

4.3 *The Integrator and Sample Hold Circuits*

The functions of this circuit, one of which is associated with each tap of the delay line, are described by equations (5) and (6) and its actual configuration is depicted schematically by Fig. 5.

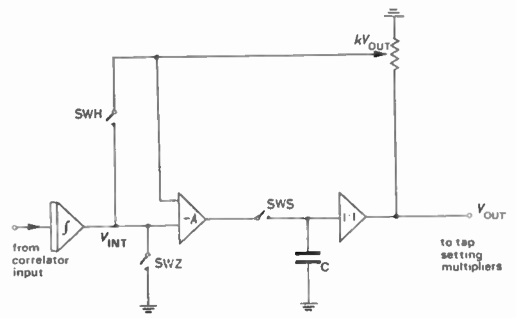


Fig. 5. Block schematic of integrator and sample-and-hold circuits.

Initially the integrator output, V_{INT}, is set to zero by closing the switch, SWZ. When the sampling switch, SWS, is closed this value is transferred to, and stored in, capacitor C and thus to the output, V_{OUT}. This forces the feedback voltage, kV_{OUT}, to be equal to V_{INT} via the high gain amplifier. The initial value of the delay line coefficients, $c_{i0} = k_{M2} V_{OUT}$, is thus set to zero. The operating cycle is then completed by opening SWS and SWZ and clamping V_{INT} to kV_{OUT} by closing SWH.

This sequence of switching operations is repeated for each iteration interval except that SWZ is not operated and hence, at the beginning of the operation, V_{INT} remains at the previous value. The operation then commences with the opening of SWH to permit V_{INT} to change to a new value during the integration interval, T_I (eqn. (5)). By closing SWS this new value is transferred to C and V_{OUT} and, after opening SWS and closing SWH, the new output value and thus c_{ij} is held until the beginning of the next iteration. Thus equation (6) is satisfied and the iterations are repeated until there is no longer any measurable difference between V_{INT} and the previous value of V_{OUT}, that is, $m_{ij} = 0$.

The adjustable factor k in the feedback loop of the sample-hold circuit has been incorporated to permit the gain product, K_2 in equation (5), to be made equal for all 41 tap control circuits.

4.4 *The Test Signal Generator*

The p.r.n. test waveform is generated in the usual way by a shift register with feedback[8] and the clock rate (100 kHz) is chosen so that the spectrum of the waveform is essentially flat over the nominal frequency band of voice signals.

The system can only cancel frequencies up to 3·33 kHz as defined by the delay line tap spacing of 150 μs. Any components in the p.r.n. signal above this frequency cause spurious components to be superimposed on the correct correlation value. In practice even small spurious signals are sufficient to cause considerable deterioration in the echo canceller performance and it is necessary to filter them out. An ideal low-pass filter cutting off at 3·33 kHz and having a linear phase characteristic (that is, $(\sin t)/t$ impulse response) would be required to maintain the mutual independence of the time series coefficients of the test signal. This, of course, is not attainable in practice but it was found experimentally that the sharpness of the cut-off rather than the phase-linearity was more important to the overall system performance; that is, a considerable amount of phase dispersion of the test signal could be tolerated as was the case with the phase linearity of the delay line. The effect of this is to increase the convergence time which within limits is not as critical a parameter in the system realization as the cancellation error.

Nevertheless, to limit the convergence time the minimum allowable value of T_I (integration time) should be used. For short integration times T_I must be an integral multiple of the p.r.n. sequence length to maintain the validity of equations (5) and (6).

If the bandwidth over which cancellation is required is 300 Hz to 3·33 kHz, a large number of p.r.n. spectral lines must fall within this band. Since the p.r.n. sequence length is the inverse of the spectral line spacing, the minimum sequence length is determined by the line spacing required to achieve satisfactory cancellation over the continuous frequency band. Experimental findings in this regard are outlined in Section 6.

4.5 *Control Logic*

The iterative control strategy (eqns. (5) and (6) and Sect. 4.3) by which the coefficients of the tapped delay line are set requires the generation of a correct sequence of sample, hold and reset 'instructions'. More specifically, these instructions perform the following operations:

(i) Reset: reset integrator and hold circuit output voltage to zero, at the beginning of the first iteration only;

(ii) Sample (pulse): transfer the integrator output voltage to the hold circuit;

(iii) Hold: constrain the integrator output voltage to be equal to the hold circuit output voltage.

The correlation algorithm requires an integral number of p.r.n. test sequences per iteration and thus a close coupling of the timing of these signals with the start and end of the p.r.n. test sequence is necessary. Timing pulses for the generation of the sample, hold and reset signals are therefore derived from the p.r.n. sequence itself.

The sequential logic module which generates the correct sequence of instructions has been designed according to the synthesis technique described in Chapter 12 of the Reference 11. Due to the complexity of the design, details have been omitted in this paper. Because of the need for experimentation provision has been made for the following operating modes:

(i) single control sequences;

(ii) continuous iterative control sequences;

(iii) a preselected number of iterative control sequences.

Furthermore, the number of p.r.n. sequences per iteration may be preselected as 1, 2, 4 or 8.

5 Data Acquisition Facilities

Since the echo canceller was designed and constructed as an experimental device, data acquisition and recording facilities for subsequent computer processing have been provided. According to its principle of operation the echo canceller is required to synthesize trans-hybrid impulse responses but, on the other hand, very little is known about these responses as they actually occur in an installed telephone network. It was therefore decided to use the canceller also as a measuring and recording device with which to collect this information, rather than to take separate action and build special equipment for this purpose.

For this purpose a multiplexing electronic switch with 46 inputs was provided, the output of which is fed into a 12-bit a/d converter. This facilitates the recording on punched tape of the 41 delay line coefficients and of five other data, such as the mean square error signal. Whereas for tape punching of data the multiplexer is driven synchronously with the punch, a faster mode is also provided to obtain a steady oscilloscope display of the tap settings for visual observation or photographic recording. This is an instructive mode for the observation of time-variant systems behaviour, effects of noise and other time-dependent phenomena.

As part of the data acquisition facility, manual control of the p.r.n. generator was provided which permits single sequence, multiple sequence and continuous injection of the test signal. With these facilities and associated computer programs for data evaluation and plotting, a sufficient number of experimental results have been compiled rapidly to determine the device performance and limitations under various network conditions.

6 System Optimization

The initial laboratory experiments were aimed at determining what type of p.r.n. filter would produce satisfactory overall system performance and also at determining the minimum p.r.n. sequence length which would provide satisfactory cancellation over the continuous voice bandwidth.

To resolve the first problem two types of low-pass filters were tested:

(i) A filter with an extremely sharp cut-off and high stop-band attenuation (seventh-order Cauer filter).

The stop-band limit of this filter was varied to determine the optimum value.

(ii) A filter with an impulse response very accurately approximating a $(\sin t)/t$ form and with high stop-band attenuation.

The optimum stop-band limit in the first case was found to be very close to the Nyquist frequency corresponding

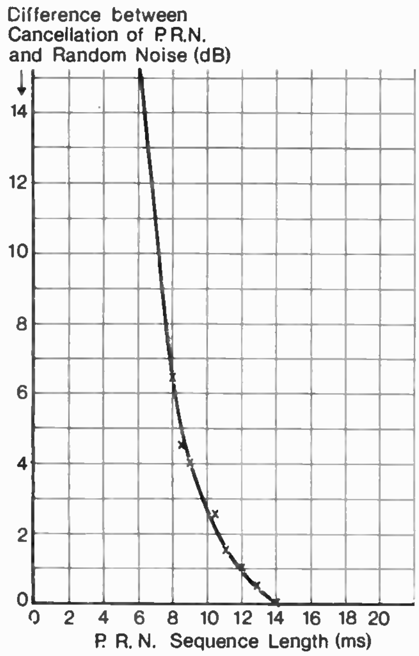


Fig. 6. Effect of p.r.n. sequence length on cancellation.

to the delay line tap spacing (3·33 kHz). When comparing the optimum sharp cut-off filter with the (sin t)/t filter, negligible difference was observed in the cancellation achieved; however, despite its effect on the orthogonality of the system, the sharp cut-off filter did not significantly reduce convergence speed. Thus this simpler sharp cut-off filter offers the most efficient solution.

Because the p.r.n. has spectral components at discrete frequencies only it is possible that, even though cancellation is good at these frequencies, it may be poor at points in between them. It was therefore necessary to experimentally determine how close these lines needed to be spaced to produce uniform cancellation over the voice bandwidth. This was most readily measured by comparing the cancellation achieved using the p.r.n. test signal with that when filtered white noise generated by a noise diode was used. The results of this test are shown in Fig. 6, from which it is evident that at least a 14 ms sequence length is required.

A 20 ms sequence length and the optimum low-pass filter were used in subsequent performance trials of the echo canceller.

Another parameter which was experimentally adjusted was the loop gain of the system. The loop gain, as defined from equation (9), should be unity for the fastest possible convergence in an ideal system. Because of the necessity of utilizing the full dynamic range of the correlator multipliers, a high gain amplifier (A4 in Fig. 2) was placed in the feedback path to drive the input to the cross-correlators and was designed to produce approximately unity loop gain. With the performance of the other system components optimized individually the resultant correlator drive amplifier gain (k_{A4} in equation (5)) is of the order of 100. This causes A4 to overload at the output during the initial stages of convergence because of the relatively large error signal at this time. However, it has been found in practice that this temporary non-linearity does not prevent the system from converging or seriously limit the convergence rate. The ultimate cancellation which is achieved is also unaffected because in the latter part of the convergence process the operation is linear.

In practice the loop gain is adjusted more exactly by first introducing a temporary additional gain factor of 2 into the system by mismatching a nominal 600 Ω termination in the feedback path. The optimum operating condition then corresponds to a loop gain of 2 and by adjusting the test signal level, which is one of the loop gain parameters, to the point where the system is on the verge of oscillation this condition is easily obtained. The mismatch is then removed and the theoretically optimum condition of unity loop gain is attained. However, the unavoidable non-orthogonalities in the system and the initial non-linearity mentioned above cause interaction between delay line settings. The effect of this interaction is minimized by increasing the loop gain as far as possible, while maintaining stability and thus the fastest possible convergence rate is achieved. The lining-up process is therefore modified to produce the optimum stable condition rather than unity loop gain. The effect of loop gain on the obtainable convergence time is shown in Fig. 7, for the loop gain ($L = K_1 K_2 s^2$ from equation (9)) being 1 and 1·7 respectively. For this experiment the integration interval, T_I, was equal to one p.r.n. sequence length, that is, 20 ms. It was found that a value of $L = 1{\cdot}7$ represented the optimum for the existing hardware configuration leaving sufficient stability margin for slight variations of component values.

It was then of interest to determine the effect of the length of the integration interval, T_I, on the convergence rate, in terms of the attained echo balance return loss (defined in Sect. 7) and the required number of iterations. The results of this experiment are shown in Fig. 8.

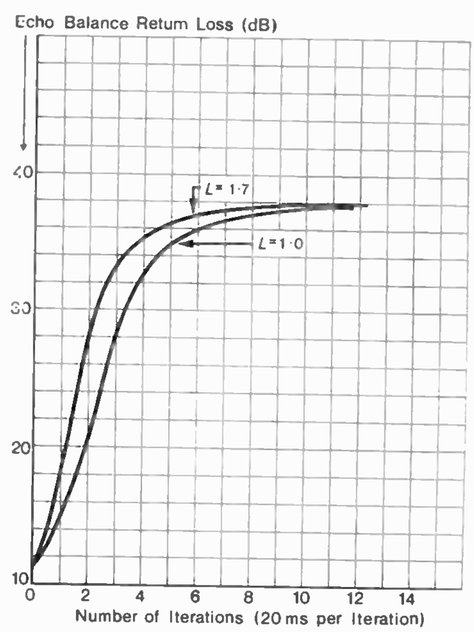


Fig. 7. Effect of loop gain on convergence rate.

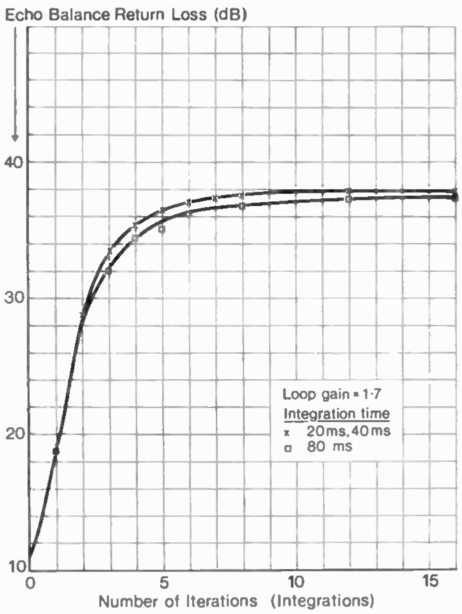


Fig. 8. Effect of integration time on convergence.

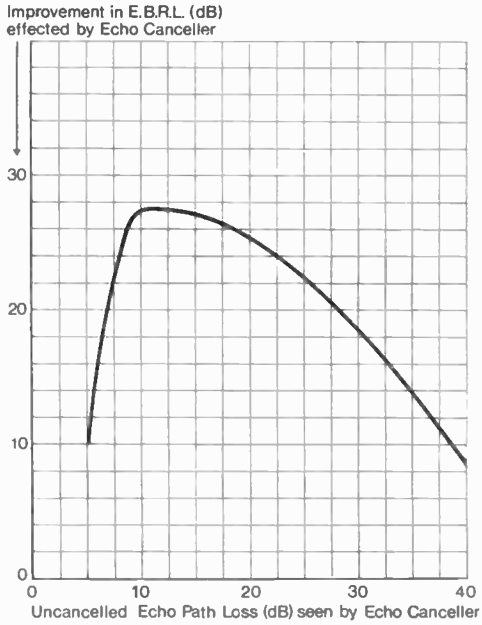


Fig. 10. Effect of input error signal level on echo canceller performance.

Maximum cancellation is achieved only after a considerable number of iterations but a value within 2 dB of the final value is obtained after five iterations. If essentially complete cancellation is assumed to have occurred after 8 iterations using a 20 ms integration time, then the canceller preparation time for each new circuit connexion would be 160 ms. It can be seen that the integration time has negligible effect on the end result and the 20 ms value is therefore used. It may, however, be necessary to increase T_I if the echo canceller was operating in an unusually high noise environment when a compromise could result between the attainable cancellation and T_I and thus convergence time (this was, however, not investigated during this series of experiments).

Since the experimental system must be expected to contain non-linear components, as, for example, the amplifier A4, it was important to know the cancellation performance, in terms of the improvement in echo balance return loss, obtained as function of the magnitude of the uncancelled echo 'seen' by the echo canceller.

By inserting two amplifiers as shown in Fig. 9, such that their cascaded gain equals unity, the uncancelled return loss, may be varied without affecting the levels within the telephone network. By varying the gain, G, the uncancelled return loss against the improvements in return

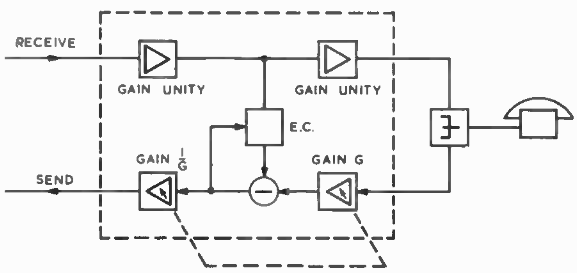


Fig. 9. Echo canceller inserted in four-wire circuit.

loss effected by the echo canceller was measured and the results are shown in Fig. 10. At low uncancelled levels the performance is limited by circuit noise while at high levels non-linearities limit the performance. The cancellation achieved by the device varies by less than 2 dB over a range of about 11 dB of the input level and, because of the low standard deviation of trans-hybrid return losses, (see Sect. 7), the echo canceller could be set up to operate close to its optimum input level wherever it is installed in the network. The gain, G, would then be adjusted such that an average uncancelled echo of about 15 dB would be seen by the echo canceller.

Having optimized the system with respect to the above parameters actual echo cancellation measurements were performed in laboratory and field experiments, the results of which are described in the following section.

7 Results

For practical operational reasons the performance of the echo-canceller is expressed in terms of echo balance return loss (e.b.r.l.), which is a measure of the mismatch between the hybrid terminating impedance and the two-wire line impedance over the voice bandwidth. Accordingly it gives a measure of the amount of signal crossing the hybrid, independently of any four-wire losses between the echo canceller and its associated hybrid. The trans-hybrid loss, on the other hand, includes all losses in the four-wire circuit between the hybrid and the echo canceller (for example, level-adjustment pads), and e.b.r.l.s are derived from measured trans-hybrid losses by subtracting these circuit losses.

The uncancelled trans-hybrid loss, P_u is measured as

$$P_u = 20 \log_{10} \frac{s_{rms}}{e_{rms}} \text{ dB}$$

where s_{rms} and e_{rms} are r.m.s. values of $s(t)$ and $e(t)$,

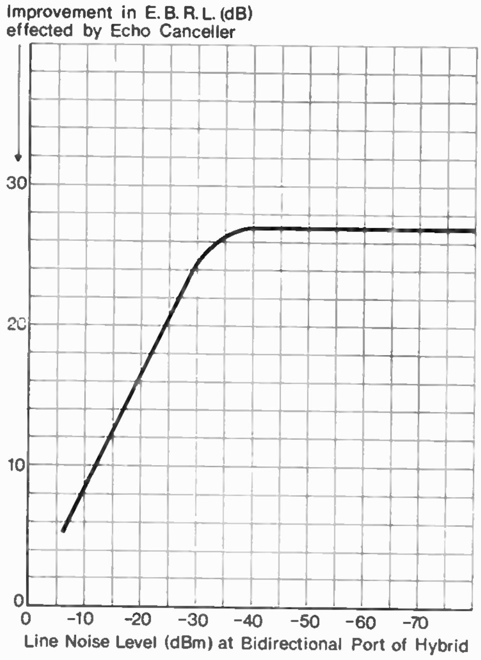


Fig. 11. Effect of line noise on echo canceller performance.

respectively (defined in Fig. 2). After cancellation, the trans-hybrid loss, P_c is given by

$$P_c = 20 \log_{10} \frac{S_{rms}}{\varepsilon_{rms}} \text{dB}$$

where ε_{rms} is the r.m.s. value of the residual error signal, $\varepsilon(t)$, as defined in Fig. 2 and by equation (2). Measurements of P_c were made after the echo canceller had converged but with the control process continuing, using filtered p.r.n. as the test signal, $s(t)$, as described earlier.

The corresponding e.b.r.l.s, E_u and E_c, were derived from the uncancelled and cancelled trans-hybrid losses P_u and P_c, respectively, by subtracting the appropriate losses (for the configuration of Fig. 12(a), for example, these losses were approximately 17 dB).

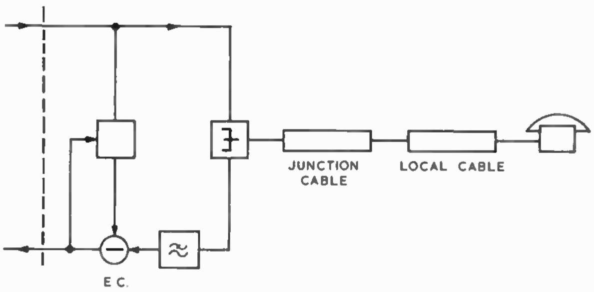
7.1 *Laboratory Measurements*

In the laboratory the echo canceller was used with a hybrid terminated on the line side in various (electrical) lengths of simulated loaded cable, which were themselves terminated resistively. The hybrid terminating network used was a 600 Ω resistor. For these conditions the uncancelled e.b.r.l. was quite low (typically 11 dB); after cancellation the e.b.r.l. was measured as generally greater than 36 dB and the greatest improvement between uncancelled and cancelled e.b.r.l. was 27 dB.

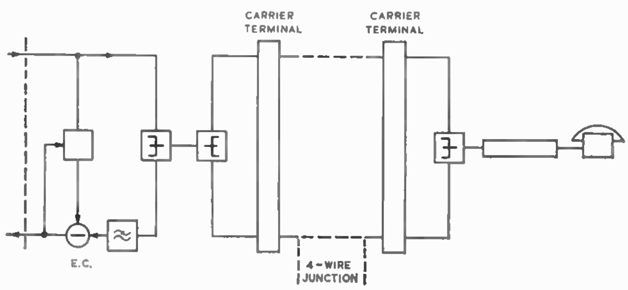
Noise was introduced into the echo path at the bidirectional port of the hybrid to observe its effect on the cancellation process. During these tests the test signal was injected into the receive path at 0 dBm. The improvement in e.b.r.l. after cancellation as a function of line noise level is shown in Fig. 11 and from this it can be observed that noise levels below about −30 dBm have little effect on the cancellation performance.

7.2 *Measurements in the Telephone Network*†

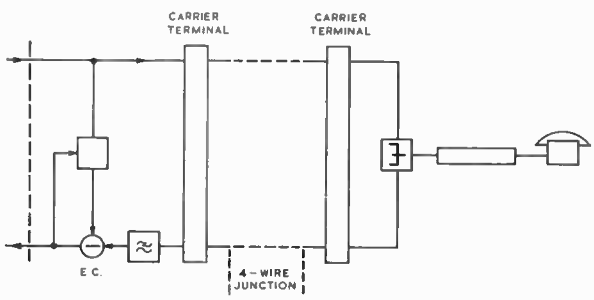
Measurements were made to determine the degree of echo cancellation that could be obtained with the experimental echo canceller when operated with circuits of the actual telephone network and, also, to record the trans-hybrid impulse responses of these circuits. The circuit configurations used are shown in Fig. 12 and were selected as being fairly representative. Figure 12(a) shows a typical metropolitan configuration in which the canceller is associated with a hybrid terminated in a two-wire line which may comprise loaded junction and local cable. Figure 12(b) represents a two-wire switched configuration as used for example to Extended Local Service Areas, in which the terminal exchange is connected to the trunk exchange via a junction containing a carrier system. The third configuration, Fig. 12(c), shows a four-wire switched connexion where the echo canceller is separated from the hybrid by a relatively long junction or trunk circuit containing a carrier system.



(a) Two-wire tail.



(b) Four-wire system in two-wire tail.



(c) Extended four-wire working.

Fig. 12. Circuit configurations for echo canceller tests.

† Preliminary results were first published in reference 12.

Table 1

Echo balance return losses measured with adaptive echo canceller

Line configuration		Uncancelled e.b.r.l. (E_u)		Improvement ($E_c - E_u$)		Number of lines
Figure	Type	mean	σ	mean	σ	n
12(a)	Local (unloaded)	15·3	2·7	20·1	1·9	81
12(a)	Short loaded junction (< 0·7 mile)+local	12·2	2·7	21·5	1·6	21
12(a)	Loaded junction (> 1·0 mile)+local	10·8	0·8	19·9	0·9	127
12(b)	Four-wire carrier in two-wire tail	10·1	3·1	9·4	3·2	20
12(c)	Extended four-wire tail	10·2	1·3	10·5	1·4	18

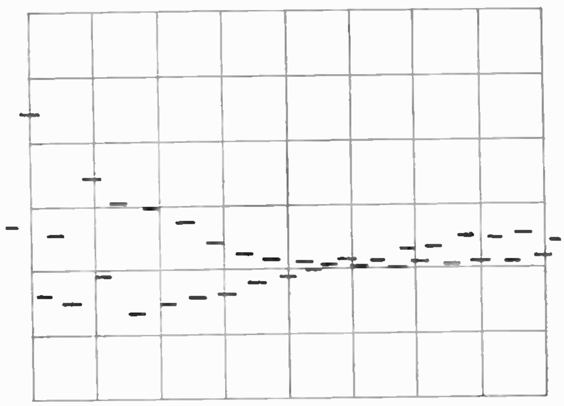
Note: Hybrid terminating impedance was 600 Ω+1 μF.

Figures 13(a), (b) and (c) show typical synthesized trans-hybrid impulse responses for the circuit configurations of Figs 12(a), (b) and (c), respectively. The displays show the time multiplexed coefficient-multiplier settings.

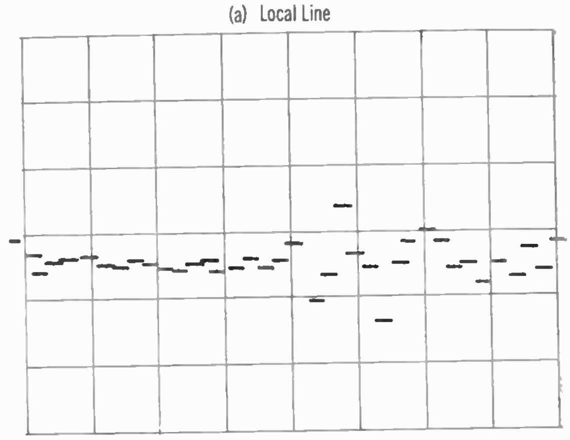
Mean uncancelled and cancelled echo balance return losses, E_u and E_c, for the line configurations of Figs 12(a), (b) and (c) are given in Table 1.

It is observed that for purely local lines the uncancelled e.b.r.l. is better by a few decibels than for lines containing loaded junction cable, since the latter tend to have higher line impedances. The mean improvement in cancellation and the standard deviation, σ, is also given in Table 1. For the configuration of Fig. 12(a) the mean cancellation improvement for all types of line is seen to be at least 20 dB but, for the configurations of Figs 12(b) and (c), only about 10 dB. This low improvement measured in the actual operating environment is due to the fact that the trans-hybrid impulse responses had a time spread which extended over a larger time interval than the total delay of the delay line. Thus the number of coefficients available to synthesize $\hat{h}(t)$ was insufficient. More specifically, in the configuration of Fig. 12(a) this problem was largely absent. It occurred, however, in some cases when the cable length between the subscriber and hybrid was short and the effect of the d.c. feed bridge, a network through which d.c. power is supplied from the exchange to the subscriber, became the dominant component in the two-wire impedance faced by the hybrid. These feed bridges have typically a resonance at about 100 Hz which in the measurements caused a large echo spectrum component in the region of this frequency. Since this frequency is normally well below the nominal voice frequency spectrum it was realistic and appropriate to change the low-pass filter shown in Fig. 12 into a band-pass filter having a low-frequency cut-off at about 300 Hz. The results shown in Table 1 were obtained with this filter in circuit.

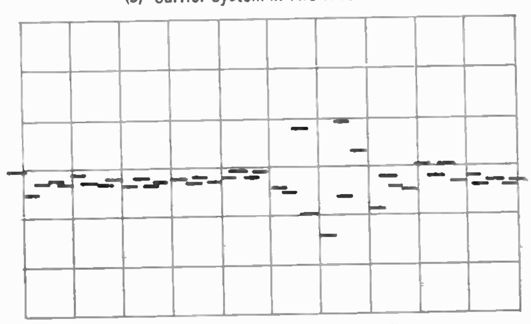
In the configuration of Fig. 12(b) the extended impulse response is due to the group delay of channel filters in the carrier systems. The echo component passing directly through the left-hand hybrid does not experience this delay whereas the component traversing the long path



(a) Local Line



(b) Carrier System in Two-Wire Tail



(c) Carrier System in Four-Wire Circuit

Fig. 13. Time-multiplexed coefficients of delay line after cancellation.

through the three hybrids does. Therefore the impulse response (Fig. 13(b)) is spread over a longer time, determined mainly by the channel filters which delay the second echo by typically 2 ms. This problem can only be overcome by a longer delay line because the echo components from both paths have been found to be of the same order of magnitude.

For the configuration of Fig. 12(c) the channel filter delay again causes an extedned impulse response but, since in this case the entire echo is delayed, the initial portion of the response corresponding to the filter delay time is substantially zero (Fig. 13(c)) and need not be cancelled. A fixed delay in front of the delay line would essentially eliminate this region from the cancellation process. The magnitude of the fixed delay would have to be determined automatically but a simple test at the beginning of the call would be adequate since it would not change during a call.

The fact that the trans-hybrid impulse responses are extended by the carrier filter delays is the reason for the relatively poor cancellation achieved for the configurations of Fig. 12(b) and (c). However, this is not a limitation of the principle but merely of the specific hardware implementation described here.

Another 'active' effect of the presence of carrier systems in the operational networks is that for some of these systems the frequencies of the oscillators in the modulating and demodulating terminals are not exactly in synchronism. This causes the trans-hybrid impulse response seen by the echo canceller to vary cyclically at a rate determined by the difference frequency of the two oscillators. The effect can be observed by maintaining the control process and is then apparent as a cyclic variation of the coefficients of the synthesized impulse response, $\hat{h}(t)$, as shown in the time exposures Figs. 14(a) and (b), corresponding to the 'frozen' responses of Figs. 13(b) and (c). Figure 14(c) depicts the variation of a single delay line coefficient, in this particular case with a frequency of about 1 Hz.

Thus, as long as this type of carrier system remains in the telephone network, this echo canceller and any others whose control process is non-continuous and/or not fast enough to follow these cyclic variations (say up to 5 Hz) will not be acceptable as a valid operational solution. However, longer term drifts and variations of the trans-hybrid impulse response can be dealt with in a discontinuous mode by occasional repetition of the test and control process. During operational trials of the described echo canceller, it was found that single iterations (20 ms) repeated at intervals of 20 s maintained the desired cancellation. These short bursts of p.r.n. were not perceived by the user although they were injected only during gaps in the speech signal.

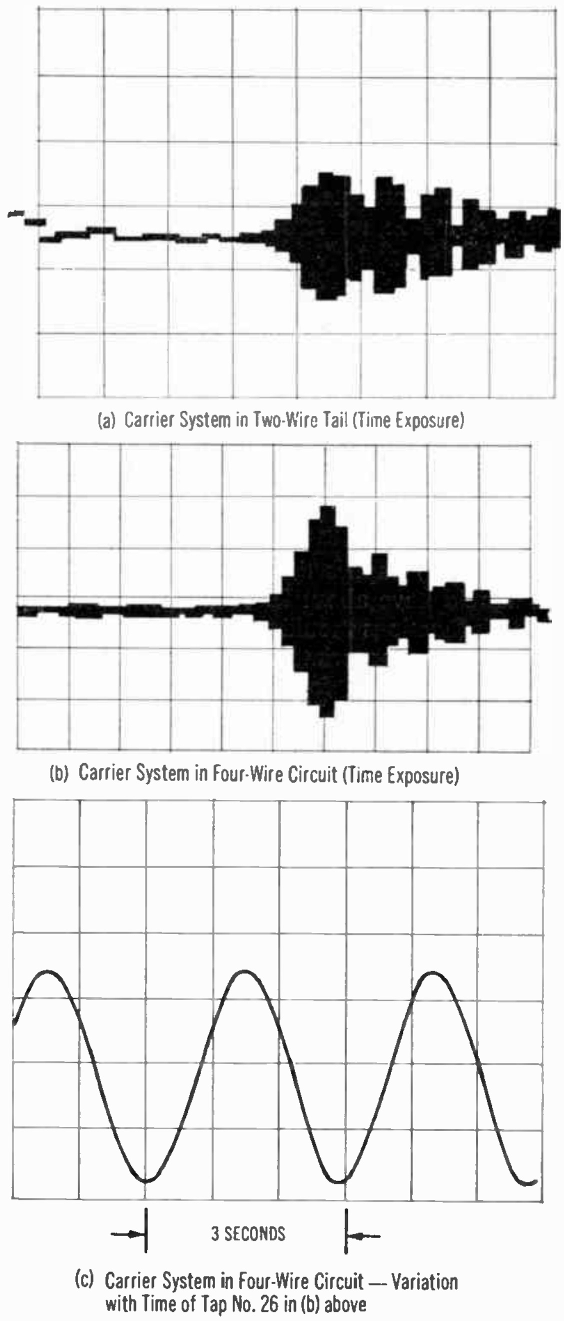


Fig. 14. Effects of unsynchronized carrier terminals on transversal filter tap settings.

7.3 *Addendum (January 1973)*

The echo canceller described in this paper underwent further field trials on a satellite link between Sydney, New South Wales, and Perth, Western Australia, in mid-1970. The performance obtained was consistent with that recorded in Section 7.2 of the paper.

The control philosophy used in this canceller, based on the injection of a p.r.n. test signal, is unsatisfactory when the device is confronted with an unsynchronized carrier system (see Sect. 7.2). This problem occurs on a relatively small number of circuit configurations but must certainly be accommodated if the device is to be versatile. To this end an echo canceller,[15] capable of following the time-varying trans-hybrid impulse responses which result from the unsynchronized circuits, is under development and a paper reporting on this development is being submitted to *Australian Telecommunications Research*.

8 Conclusion

By constructing and experimenting with the described experimental echo canceller it was demonstrated that the concept of adaptive echo cancellation is operationally valid and physically and economically implementable in terms of present technology.

Operationally acceptable performance criteria, as for example stated in references 13 and 14, can be met provided:

(i) the time spread of the trans-hybrid impulse response of about 8 to 10 ms occurring in practical telephone networks is matched by the synthesized time series;

(ii) the identification and control process is continuous and convergence is rapid enough to follow fast variations of the trans-hybrid impulse response caused by the presence of 'nonsynchronous' carrier systems in the telephone network.

Condition (i) implies that delay line type transversal filters should have about 80 coefficients. Alternatively other basis functions should be investigated with which to synthesize the appropriate impulse responses aiming at a smaller number of coefficients required to meet the operational error criteria. Such studies are now more meaningful than before the reported experiments since representative samples of real trans-hybrid responses have been obtained.

Condition (ii) implies that the actual voice signals transmitted over the network are used for systems identification and control and that rapidly converging control strategies can be implemented.[15]

9 Acknowledgments

The authors are indebted to a number of colleagues in the A.P.O. Research Laboratories who during the time of preliminary work and in the implementation phase have contributed to the success of this investigation. Particular mention is due to Messrs. R. B. Zmood, R. Buring, G. Jacoby, M. Subocz and Dr. W. H. Otto. The permission for publication of this paper of the Senior Assistant Director-General, Australian Post Office Research Laboratories, is acknowledged.

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Correction

The following alteration should be made in the Letter on 'Designing asynchronous counters' published in the March 1973 issue of *The Radio and Electronic Engineer*.
Page 227, col. 1, Table 1:

The symbols R and S in the first column of this table should be interchanged.

UDC 621.3.049.77: 621.3.015.5: 621.391.823

An investigation of integrated circuit destruction by noise pulses

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SUMMARY

Conventional integrated circuits have a high failure rate when operated in an environment including electromechanical switching systems. This paper discusses the destruction modes for the circuits and the equivalent circuit for the noise generation system. Experimental results which support the concept of the equivalent circuit are discussed. Finally, the transient thermal characteristics of typical packaging systems are briefly discussed.

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1 Introduction

Military applications of electronic systems are frequently thought of as the most demanding in respect of environmental operation. Although this is the case in terms of absolute technical limits, it is not the case in terms of total system environment in markets which are controlled by the conventional economic constraints. This lesson is rapidly being learned by those companies which have for many years depended to a large extent on the military and space market but are presently expanding into civilian applications of electronics technology. Engineers are finding the difficulty of fitting electronic systems to the motor car and automation industries is greater than that of applying electronics to parallel military applications.

There are two reasons for this: in the first place the overall cost differential, and in the second place technical difficulties which are too costly to avoid within a civilian cost environment.

One area in which both these factors come into play is that of the use of electronics in automation systems. The use of conventional logic systems, making use of t.t.l. integrated circuits, has of course made inroads into this field of application. However their use has been very limited by the industrial environment, not in terms of temperature and vibration etc., but in terms of electrical noise. A conventional computing system, in which a reasonably complex set of logic can be installed within one screened metal box, through which a comparatively small number of inputs from the external noisy world can be carefully introduced and filtered, can only be applied to a fairly complex automation project.

However there are numerous industrial areas in which automation must be applied at a fairly low level of complexity. Such applications call for electrical components which can withstand the high level of electrical noise within factories but without expensive filtering at each input. Such components are likely to be utilized in small quantities at a given location and interconnected to other groups and the control centre by means of long wires. The noise coupled into these interconnecting wires is likely to have two effects on the system operation, the first being temporary disturbance of the circuitry outputs and the second being total destruction in respect of the system performance parameters.

Therefore it is clear that components intended for use in these areas should be capable of overcoming these likely causes of trouble. Unfortunately it is true to say that up to the present time, integrated circuits have not been used in these environments to any extent because they are not sufficiently reliable or cost too much to be ruggedized. Although there are several logic systems for industrial use on the market, none of them meets the challenge of the industrial environment. This is because there is a basic lack of knowledge in respect of the industrial environment within the electronics industry as a whole and the integrated circuit manufacturers in particular. An additional reason is that for many years the integrated circuit manufacturers have had an enormous and easy to supply computer market, plus military research backing. It has thus been something of a waste of money for integrated circuit manufacturers to develop a difficult market.

However all the signs are that because of changing industrial pressures, the integrated circuit manufacturers are attempting to move into the industrial automation area. This paper discusses some of the design features to be taken into consideration when designing integrated circuits for industrial applications and shows why some present day integrated circuits are unreliable.

2 The Electrical Environment

Electrical components operating within a factory are likely to be connected to long wires, running through the same trunking as cables used to control heavy duty contactors, motors, electrohydraulic valves and various like devices. Thus the electrical environment of the integrated circuit is basically that of the voltages and currents at the terminals of the piece of wire.

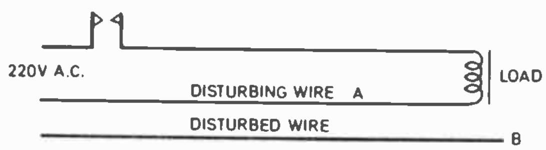

Fig. 1. Coupling of disturbances.

Figure 1 shows a piece of wire parallel to a wire having a load which is being switched by means of an electromechanical contact. It is clear that as wire A is switched, noise will appear in the wire B because of coupling; and this will be dependent upon the rate of change of current in the voltage in wire A, as well as duration of the switching period.

Since the rate of change and duration are very much a function of the switching mechanism, we should begin by looking at the make and break periods. Perusal of manufacturers' literature shows that relays have typical changeover times varying from 200 μs in the case of reed relays to milliseconds in the case of contactor size relays. For a portion of this changeover time, the relay contacts will be close enough to form a spark gap. Measurement of this period and comparison with available sales literature suggests a time of less than 1 ms as the period during which the contacts are likely to produce sparking.

This sparking will be in the form of high frequency oscillations which are associated with the inductance and capacitance in the vicinity of the contacts. Measurement of the frequency range, both on an oscilloscope (by looking at the terminal voltages) and by the use of a radio receiver, indicates that the resulting noise is likely to be within the range of hundreds of kilohertz to tens of megahertz.

Therefore, as a first estimate, we can consider the noise as likely to be less than 1 ms duration pulses at a frequency of hundreds of kilohertz to megahertz. It is rather important to know the power level of these pulses, so that will now be discussed.

If we regard the oscillating system as the prime source of power, we can calculate the energy stored in likely associated values of inductance or capacitance and thus obtain an idea of the power introduced into the disturbed line. Take for example a coil of value 100 μH resonating with a capacitor of approximately 200 pF at 1 MHz. The energy stored by the wire inductance when passing a current of 10 A would be $\frac{1}{2}LI^2$, giving 5 millijoules. Much of this energy would be used up in the spark and lost via coupling. Therefore the energy from that source can be considered as not very significant.

Another source of induced noise would be that caused by the collapse of the magnetic field along the length of the line carrying the large current. This could cause a very high voltage to be induced along the disturbed wire, but the power would be small and the coupling impedance very high—thousands of ohms probably in the case of likely wire dimensions.

However there is a very important source of power, in that as the relay contacts are made and broken by means of the small value local parasitic reactances or low power line reflexions from mismatched wires, the 220 V line voltage is effectively modulated by this switching. Therefore since the 220 V source is likely to have a local storage system (such as a high value capacitor), we can consider the noise source as the mains voltage modulated by a relay operating at high frequency. This voltage will have a possible peak amplitude equal to the peak of the mains voltage (220 V or 110 V times 1·4).

This state of affairs is shown on Fig. 2, in which are represented the capacitances coupling together the disturbed and disturbing wires, as well as the capacitance to earth.

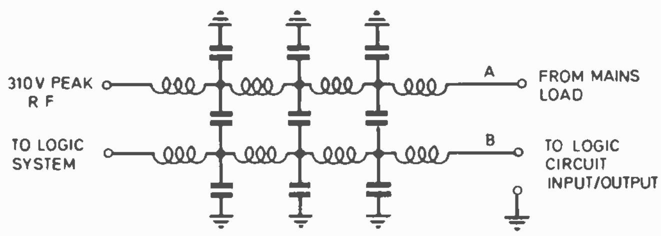

Fig. 2. Capacitances between disturbed and disturbing wires.

3 Evaluation of Power Levels

If the capacitance to ground in Fig. 2 is assumed to be very small, as with roadside telephone wires, the pair of wires A and B is likely to have an impedance of approximately 140 to 320 Ω for wires having a dielectric constant of about 4 and ratios of separation distance to wire diameter within the range 5 to 100. For this range of wire arrangements, which can be taken as typical, the line inductance is in the region of 2 μH/m; and line capacitance in the region of 25 to 70 pF/m.

For the range of frequencies being considered, the wavelength is from tens to hundreds of metres. Thus for given frequencies within that range, the likely lengths of wire can obviously act as quarter-wave transformers as well as producing other effects associated with lines having sub-wavelength dimensions. In the case of short lines, the line can act as a mismatched delay section, thus producing high-frequency pulse effects. However the typical line to be met in the factory might have attenuations in the range of 1 to 10 MHz between 1 and 5 dB/30 m, under ideal conditions. In the case of humidity and lossy surroundings, the attenuation can become much greater.† The line length is likely to have an effect on the radio-frequency rate of commutation at the spark gap. However, the effect on the absolute power level by virtue of the charge stored within the line reactances is likely to be low.

Referring to Fig. 2, one can consider the input to a logic circuit to be connected to a radio frequency source with a peak value of 310 V when sparking coincides with a peak of the 220 V supply. The connexion is made via a line which can have an impedance of 140 to 320 Ω (under the worst case conditions of two closely positioned wires which are sufficiently far from the earth and trunking etc. to be considered in isolation). This state of affairs can be considered as not likely to arise in practice. However, if such a combination did exist, then the logic circuit would have to take a current of up to 320 divided by 140 A, i.e. 2·4 A peak. This of course assumes that the voltage drop at the input to the logic circuit is small by comparison with 310 V—an assumption which is normally justified. Turning away from this rather hypothetical set of conditions to one which is more representative of actual conditions, the disturbing wire is likely to be coupling energy into other wires in the proximity of the wire considered to be the disturbed wire, and also to earth via trunking etc. As a result the power coupled into the disturbed line will not be as high as just calculated. To obtain a reasonable figure we must estimate the relative values of the impedances between the various wires, as well as to earth. Regarding the latter, since the surface area of the trunking is likely to be considerably larger than that of the disturbed wire, the coupling to ground is likely to be of larger value than the coupling between wires. This capacitance to earth cannot be considered as a lumped element but is of a distributed nature, coupling with the inter-wire capacitance at an infinite number of points.

In fact we can consider the practical equivalent circuit to be similar to that in Fig. 3, in which the wires are coupled by an impedance equal to their inter-wire impedance when viewed separately, but also coupled to

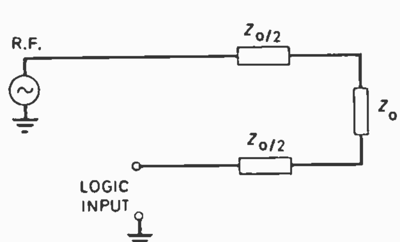

Fig. 3. Practical equivalent circuit of input to logic circuit and a disturbance.

† Hund, A., 'Frequency Modulation', p. 322 (McGraw-Hill, New York, 1942).

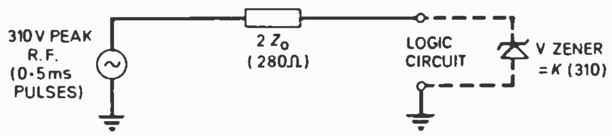

Fig. 4. Equivalent circuit for high power noise system.

earth, as with a single wire above the earth. Assuming both of the wires to be of the same dimensions, the inductance will be the same in both cases. However the capacitance to earth will be greater than the capacitance between wires under normal conditions, so resulting in a potential dividing effect, as shown in Fig. 3.

Considering the ratio of the impedance between two parallel wires to the impedance between one wire with a coaxial surrounding; this does not depend upon the wire dimensions but only on the ratio of larger diameter to inside diameter to wire spacing to wire diameter. This ratio is approximately constant at a value of 2 for anything other than very small ratios. Therefore we can consider, as an approximation, that the noise generator drives through an internal impedance of value equal to half of the inter-wire impedance, then through the inter-wire impedance before being applied to the logic circuit input through the disturbed line impedance which is equal to that of the disturbing line. Thus the effective inter-line impedance is twice that of Fig. 2. Therefore we can consider the noise source at the input to the logic circuit as being a 310 V peak generator over the frequency range already mentioned and with an internal impedance of 280 to 640 Ω. The value of 280 Ω assumed that the lines were lossless. In fact there are losses which will effectively increase this lowest value of impedance by an unknown factor. This worst case value cannot be considered as a representative case because of this unknown loss factor. The loss factor will become greater at higher frequencies since the voltage driving the current from one line to the other will decrease more rapidly with progression along the line. It can be seen from this that the loss is best translated into a reduction in the generator voltage, as compared with an increase in line impedances. Since this loss factor can be only roughly estimated, and could even be insignificant, it must be ignored, so that finally we have the equivalent circuit of the noise generator as shown in Fig. 4.

4 Effect of Noise on System Inputs

In general the input impedance of electrical devices used in typical automation systems will not be linear. As an approximation, we can consider the input impedance to be a resistor in series with a diode, because integrated circuits incorporate a diode at all of their inputs by virtue of the mode of manufacture. The resistor can vary in value between ohms and kilohms, with diode breakdown voltages varying between zero and (say) 100 V. Obviously the maximum current flows when the diode voltage approaches zero. However, although that sets the limit for the peak current handling capacity of the input connexion bond and the diode itself, the power dissipated in the input will be low. As with the transmission line matching system, maximum power will be dissipated in the input when the voltage breakdown of the diode and the resistance of the input, combine to given maximum power transfer.

However the resistance will be small if Zener protection is used at the input. Therefore we will assume that the diode has zero internal impedance. As a result the current is limited by the line impedance itself, as shown in Fig. 4. Figure 5 shows the variation of peak power dissipated in the diode, as a function of breakdown voltage. The current variation is also plotted.

The K factor in Fig. 5 is the ratio of the input diode breakdown voltage to the noise source voltage (in the case of 220 V mains, this being the peak value 310 V). The horizontal axis indicates voltages in the particular case of 220 V mains on the lower scale, and any voltage on the upper % scale. The left-hand vertical axis refers to the variation of power dissipated as a function of % of P_0, where P_0 = open-circuit disturbing system voltage × short-circuit current at terminals. The right-hand vertical scale indicates the value of the current flowing as a percentage of maximum current.

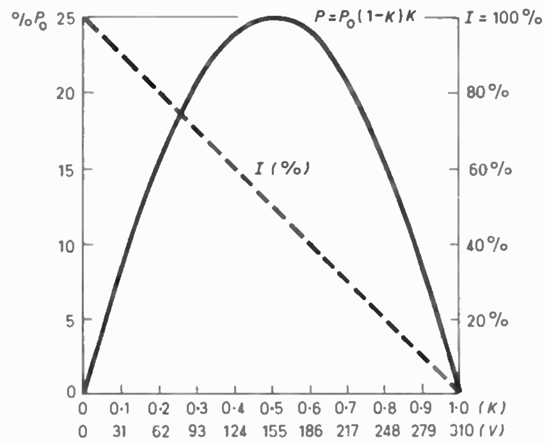

Fig. 5. Variation of current and power in load for a 310 V d.c. source.

It can be seen that the greatest power transfer efficiency occurs when the diode breakdown voltage is equal to half the peak applied voltage. The efficiency of 25%, and input level of 310 V into 280 Ω, gives a basic power level of 340 W and actual peak power dissipated in the input of 85 W. The peak current to flow at the low breakdown voltage end of the scale is approximately 1·1 A.

In fact the power dissipated will be a function of the period during which the diode is conducting. In the case of a square wave, since the diode will be conducting for half the time, the power transfer curve will be as in Fig. 6 where the verticals are equal in value to half those of Fig. 5. In the case of a sine wave, the power transfer efficiency is nothing like as high because the diode is now conducting for only part of the cycle, as opposed to a complete half-cycle. By the method shown in the Appendix, the lower curve on Fig. 6 can be calculated. This shows that the power transfer to the diode is much less efficient (note that the efficiencies are with respect to the peak power,

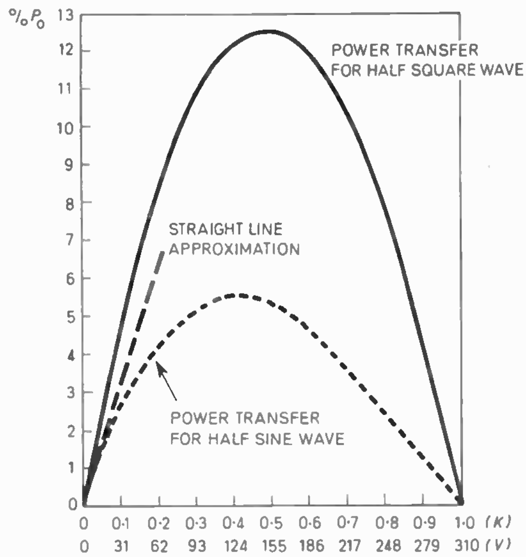

Fig. 6. Power dissipated in logic input for half sine and square waves.

not the basic r.m.s. power) in the case of a sinusoidal waveform.

One can see from these curves that although there is a large difference in the maximum power transfer, as a result of waveshape, the difference is much smaller in the range of K from zero to 0·2 (K = ratio of diode breakdown voltage to 310 V). Since we do not know the waveshape, it seems reasonable to take a representative straight line approximation as shown on Fig. 6. Thus eventually we find the curve of Fig. 7, which can be taken as a curve showing the average power which must be dissipated by a protection diode during the time that the disturbing pulse is present. The average power can be calculated from a knowledge of the pulse width and repetition frequency.

Figure 7 also takes into account the fact that during the negative half cycles the diode acts as an ordinary diode, thus allowing the passage of the maximum current. Typical diodes in integrated circuit inputs (so-called clamping diodes) produce a voltage drop of around 1·5 V when pulsed with short 1 A current pulses. Thus it seems reasonable to add 1·5 W to the totals. In addition, the diodes are not resistanceless when operating in the

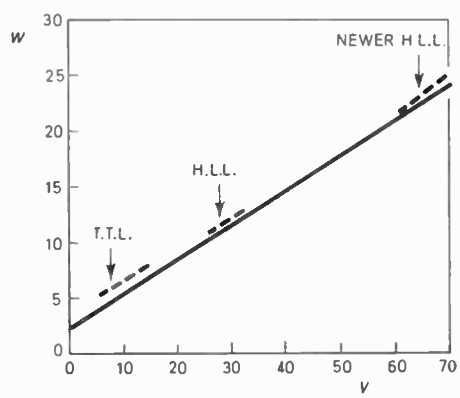

Fig. 7. Power in input diode versus input breakdown voltage.

Zener breakdown mode. The power generated in the internal resistance will increase with the square of the current, but also of course decrease the current. Since a correctly designed Zener should have a very low resistance it will be ignored.

5 Effect of Noise on Well-known Circuit Types

Best known of the available logic circuits is the t.t.l. family. These circuits have a diode breakdown voltage of the order of 8 to 15 V built into the circuitry. As a result the power dissipated within the circuit inputs, for perhaps 0·5 ms, will be in the range 4 to 7 W. This is a not inconsiderable power and explains why t.t.l. type circuits are easily destroyed when operated in an industrial environment. Unfortunately for would-be users, the problem of noise immunity has led to the use of high-input-threshold integrated circuits, described as high immunity circuits. These circuits do have a high immunity against operation by pulses which do not exceed the normal operating threshold. However they also, rather unfortunately, have breakdowns in the range 30 to 40 V and as a result, the power dissipated in the inputs is greater than with t.t.l., being 10 to 15 W. These circuits are thus even more prone to damage than the t.t.l. family when used in an industrial environment.

Yet another family of high immunity circuits is now available, having even higher breakdowns; and as a result even less protection against destruction. The associated breakdown voltage is in the region of 60 V, which causes over 20 W to be dissipated by the input during the pulse.

5.1 *Supporting Experimental Evidence*

Various logic circuits were first connected to their appropriate supply voltages and then connected to a wire coming from a noisy environment. A resistor was placed in series with the input, so as to limit the input current coming from the 'noise wire'. It was found that t.t.l. circuits were less prone to damage than high immunity circuits, and that high immunity circuits were less prone to damage than 'super-high' immunity circuits. This tended to confirm the ideas already developed. One point that should be noted is that not all circuits of the t.t.l. type have the same 'robustness', this apparently depending upon the manufacturer. Also it should be noted that the t.t.l. circuits, although less prone to damage than the circuits ostensibly designed for an industrial environment, are in fact useless in a noisy environment, their unprotected lifetime being measured in seconds or less.

Since all of the failures investigated were caused by excess current passing through the input of a device, the type of failure met was drawn from a small set of modes associated with this type of failure, thus this discussion should in no way be taken as anything other than a look at input failure modes. It is in no way a general review.

As previously remarked, when the negative-going noise pulses reach the input, the current is at a maximum of perhaps an ampere. This occasionally causes the gold bonding wire to explode, either at the bond to the chip or in mid-air. This seems to be associated with poor bonding, as seen by several attempts at bonding to the chip bonding pad, or by an unnecessarily long wire, sometimes with kinks. This seems to occur only when the quality of workmanship is poor, for the commonly used bonding wire diameters.

The greater part of the damage occurred during the positive going cycles. Occasionally the input current flowed through, for example, an internal Zener diode, which provided a low resistance path at a breakdown voltage less than that of larger area diodes in the chip. Since the through hole in the oxide was usually designed to carry perhaps a milliampere, it would burn out at the high current level of scores of milliamperes. This would leave the aluminium contact with the characteristic sintered appearance, and form an open circuit.

In the case of a particular logic family, the internal diode controlling the threshold appeared to develop parallel resistive paths within the Zener. As a result of the particular circuit design, the effect of this was to gradually shift the threshold, until at a given voltage it would become locked into a given state. In fact it was possible to observe the change of V_{in} high with time, after the input noise pulse generator had been adjusted to cause gradual, rather than catastrophic failure. A gradual shift in this voltage from say 7 to 4 V was observed, catastrophic failure occurring at the low end. Possibly this is the only example of a semiconductor device on which preventive maintenance could be used! In the case of another family, the collector to emitter diodes in the transistors would break down in the region 12 to 15 V. Some kind of secondary breakdown would then occur, with a holding voltage of less than 6 V in some cases. This would, if allowed, cause immediate destruction by passage of current from the main i.c. power supply.

Initially, since the high currents involved would cause unwanted currents to flow in various parts of the chip, the possibility of spurious triggering of p-n-p-n etc. type devices seemed highly probable. In fact, the investigation, whilst not exhaustive, indicated the modes of failure to be nothing more interesting than simple burning-out of some part of the input circuit. This was occasionally checked by applying or not applying the V_{cc} supply whilst under test. It appeared to have little or no effect on the failure. However once failure of certain types, which apparently caused internal low resistance static paths, had occurred, connexion of the power supply would cause a great deal of heat to be generated in a short period. The result would be the redistribution of parts of the chip and bonding wires in the package.

Since this work was in connexion with designing a robust logic family, at the outset it seemed best to look for the circuits of greatest robustness, seek out the reasons why they were better, and then develop the desirable characteristics. Although a difference was observed in robustness, within a particular logic family and related to manufacturing source, the difference could not be easily explained. When this was coupled with the fact that the difference was of degree, none of the circuits incorporating a suitable protective system, that avenue of attack was discontinued.

In order to obtain as positive as possible confirmation of the previously developed ideas, additional tests were performed. These were suggested by the fact that since all of the inputs had the same kind of protection input clamping diode, and that this diode was thus performing in the same way on circuits having differing failure rates, the negative noise half-cycles were not having an effect on failures. In order to confirm this point, various combinations of Zeners were connected between the logic input under test and the integrated circuit substrate. Such an arrangement is shown in Fig. 8.

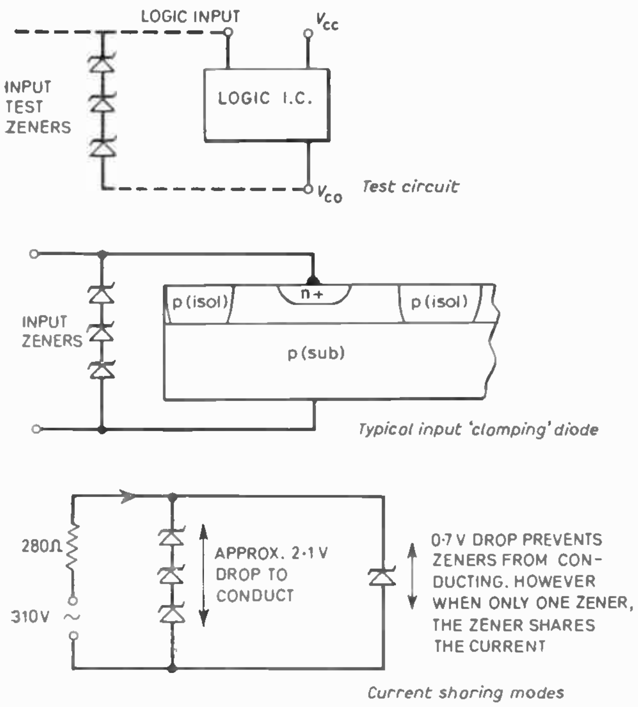

Fig. 8. Confirmation of the cause of integrated circuit destruction.

The Zener voltage was arranged to be less than the positive (Zener) breakdown voltages of the given logic circuit. This protection Zener diode was formed from either a single Zener or several in series. When only one diode was used, the Zener conducted during the negative half-cycle thus sharing the current with the clamping diode at the integrated circuit input. This was verified experimentally using a current probe. However, when the external protection diode was formed from say three Zeners in series, the combined voltage drop was approximately 2·1 V in the forward mode (negative-going noise pulses), as a result of which virtually all of the negative-going current was sunk by the clamping diode in the integrated circuits. It was found that even the most sensitive of the circuits could operate for hours without destruction when 'positive only' protected. The distribution of the currents within the various diodes was confirmed as correct by the use of current probes. During these tests, current pulses of approximately 1 A were observed. However, regarding these observations, it should be mentioned that although a reasonably accurate idea of current ratios could be obtained, the measurement of absolute values was difficult.

The Zener diode protection chain, connected in parallel with the logic input, was modified so as to increase the Zener voltage, the sharing of current between itself and the logic inputs being simultaneously monitored. It was found that the current entering the logic inputs increased very rapidly over a small increase in Zener chain voltage, in the region of the aforementioned integrated circuit input breakdown voltages. The circuits were then destroyed.

As a result of these tests, the equivalent circuit for the noise production system was assumed to be reasonably valid. However additional tests were made and further confirmed the power levels involved. These tests will now be described.

6 Heating of Diodes by Transients

At the outset of the investigation into the reasons why integrated circuits which are currently available are prone to destruction when used in logic systems in industrial environments, it was thought that the circuits were normally destroyed by parasitic devices being triggered into a self-destroying mode by the noise pulses. This was the point of view of one major vendor of high immunity circuits. However a brief calculation indicated that the noise itself was sufficiently powerful to destroy the input diodes. So as to confirm this point, various diodes were tested on both a noise generating system and on a generator delivering known current pulses of a given pulse width and at a given pulse repetition frequency. It was found that the smaller diodes were destroyed by low power pulses, and that the larger area devices required much larger powers for destruction. It was also found that the small devices were destroyed by the noise generator but that the larger ones were not. This allowed correlation between the known energy of the square waves and the energy in the noise pulses, which confirmed the model of the noise generating system as proposed earlier.

Figure 9 shows the voltage waveforms across a typical diode when passing a given current for a given time. The current in (a) is too small to destroy the device, the voltage remaining constant with time. In (b) the current has been maintained at the same level but the pulse length increased. It can be seen that the diode resistance has begun to increase, thus generating more heat and beginning to run away. In (c) the diode voltage drop is shown immediately before destruction.

The time required to produce thermal runaway was monitored on an oscilloscope for devices of various dimensions passing the same current. It was confirmed that runaway was associated with the destruction mode. It was also observed that a diode situated as close as 100 μm from the diode being destroyed was unaffected by the act of destruction. This points to a very local effect.

As the aim of the work was towards producing some logic circuits which could operate satisfactorily in an

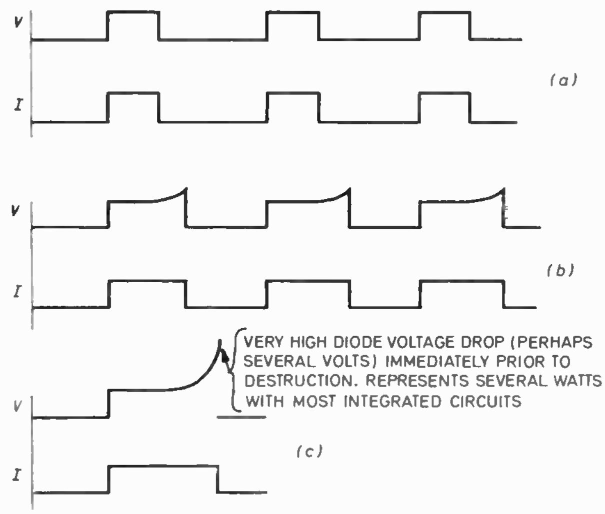

Fig. 9. Variation of diode voltage drop with length of constant current pulse.

Note: Waveforms are shown for demonstration. In fact the p.r.f. was deliberately kept low to avoid average rise in temperature. However by increasing p.r.f. it was possible to go through entire range of above waveforms.

industrial environment, it was important to establish the dimensions of a diode which could be built into an integrated circuit but still provide protection against noise.

6.1 *Simple Thermal Model of Diode*

Figure 10 shows a drawing of a diode within an integrated circuit. When heat is generated within the diode it will result in the leakage of heat to surrounding areas by conduction and the local absorption of heat by virtue of the thermal capacity of the diode itself.

It can be seen that when the diode is operating in the Zener breakdown mode, the majority of the heat will be generated near to the upper surface of the diode, in the region of the epitaxial layer. This heat will in the long term have to drain away via the chip, to the header, and finally to the outside world, which is assumed to be at a reasonably low ambient temperature. This heat path can be represented by a resistance under steady state conditions. However in the short term there is another heat absorbing system, which is the thermal capacity of the silicon itself. When the small part of the silicon near to the heat source has absorbed some heat, it will begin to pass heat onto the next part of the silicon by virtue of the temperature rise it has acquired. Each of these thermal capacities can be thought of as capacitors attached to earth, and the thermal resistance between them as series resistance.

In order to calculate the conversion factors so as to allow a simulated time-constant using resistance and capacitance:

For a resistor:

$$V(\text{volts}) = I(\text{amps}) . R(\text{ohms}).$$

For a thermal resistance:

$$\phi(\deg \text{C}) = W(\text{watts}) . R_{\text{th}}(\deg \text{C/watt}).$$

And in the case of specific heat,

$$\phi(\text{C}^\circ) = \text{heat (watt seconds)/thermal capacity.}$$

This corresponds to

$$V = Q(\text{ampere seconds})/C(\text{farads}).$$

Therefore we can look on the thermal problem as one in which the heat source becomes a current source, and the impedance to earth becomes a distributed resistance-capacitance network, terminated in a short circuit to earth and having line constants of R, being the steady state thermal resistance between the heat source and earth; and line capacitance equal to the thermal capacity of the volume considered (in watt second/deg C). In fact, as the heat travels away from the source, the area available to it for transfer will increase with the square of distance travelled and volume with the cube of distance travelled. Thus it should be represented by a transmission line having tapered characteristics. However as a simple approximation, we shall represent it by a resistor in parallel with a capacitor.

The resistance to earth will be:

$$R = (\text{thermal resistivity of silicon}) \times L/A.$$

The thermal capacity will become:

$$C_{\text{th}} = (\text{specific heat of silicon}) \times (\text{density}) \times (L) \times (A).$$

Thus the time-constant becomes

$$t = (\text{specific heat}) \times (\text{thermal resistivity}) \times (L^2) \times (\text{density})$$
$$= 1{\cdot}1\ L^2/10^5 \text{ milliseconds (for silicon),}$$

where L is in micrometres. This is equal to 0·44 ms for a distance of 200 μm (typical chip thickness).

It is interesting to see how silicon compares with some other materials which commonly go into the packaging of a semiconductor. In Table 1 the appropriate

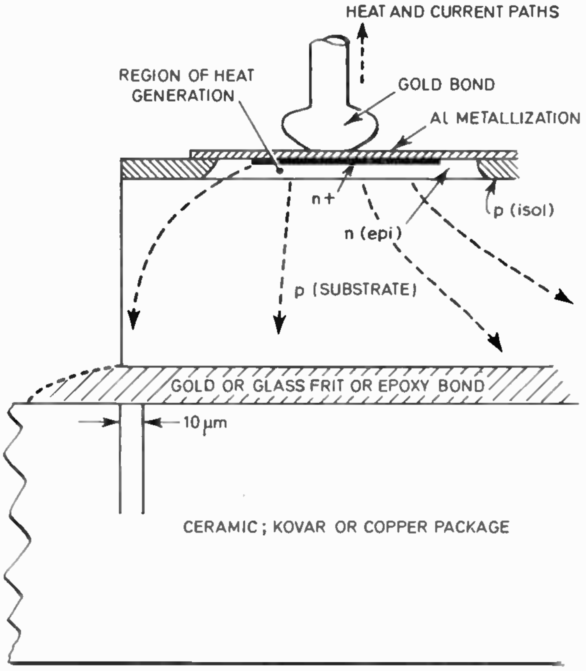

Fig. 10. Typical construction of integrated circuit input clamping diode showing heat and current paths.

variables are tabulated and multiplied together in order to give a figure of merit (the temperature rise at the end of a given period being inversely proportional to the thermal capacity and the thermal conductivity).

Table 1

Transient thermal properties of packaging materials

Material	Density (gcm^{-3})	Specific heat ($Jg^{-1}degC^{-1}$)	Thermal conductivity ($Wcm^{-1}degC^{-1}$)	Figure of merit
Silicon	2·33	0·7	1·5	2·4
Alumina	3·6	0·93	0·17	0·6
Aluminium	2·7	0·9	2·1	5·2
Copper	8·8	0·4	3·9	14·0
Gold	19·3	0·13	3·2	8·0
Iron	7·8	0·5	0·9	3·5
Silicon dioxide	2·2	1·0	0·014	0·03

7 Concluding Discussion

From Table 1 one can see that both aluminium and gold are considerably better than the silicon at keeping down the short term temperature rise. Unfortunately those materials cannot be connected into the heat path in sufficient bulk to have much effect in the normal case. However a relatively large gold bond could be placed on top of an input clamping diode and thus improve the ability to handle transients.

Some of the proposed methods of improving thermal transient response have been used. For example with the gold eutectic chip to header bond, which is technologically sound, in fact the amount of gold involved is usually very small; since gold is expensive a gold-covered copper system which should be more easily formed and less expensive was proposed.

In fact this system has been in use for a long time but mainly, it is believed, for assembly convenience and not for its thermal response. IBM have used a tin plated copper ball system extensively for their 'flip chip' system of computer sub-assembly manufacture. It would be interesting to learn if operational experience of a system using this technology in an industrial environment has provided sufficient data on failures to be able to compare it with a system using conventionally packaged t.t.l.

Another current system affording some protection is the pillar flip-chip (as opposed to the aluminium film flip-chip). A rather ancient technique was the method of transistor assembly used by STC in the UK some ten years ago. The chip was inverted and placed on the three riser wires which had been turned inwards at the top. The solder-plated 'bonding pads' of the chip melted in a furnace and surface tension rotated the chip into the correct position. This should cause the chip to have a better transient response over a certain range of time-energy than the conventionally-bonded chip. Incidentally, this was reported by a well known electronics magazine recently as a novel new assembly technique for the production of power transistors (chips on copper preforms were passed through a furnace).

For reasons of completeness, conductive epoxies are shown in Fig. 10, but data were not included because one could expect the characteristics to be very poor compared with gold, copper etc. and to be naturally excluded from any system improving thermal performance. However, as a rough estimate, one could expect a 3 to 1 decrease, at least, in thermal conductivity and a similar decrease of thermal capacity compared with gold. Thus an order of magnitude deterioration in thermal transient response seems to be applicable. Additionally, the materials are very difficult to control and could probably produce rogue devices without external indication.

In terms of production costs, assembly costs and thermal response, the 'Minibond' system developed by the now defunct G-E Integrated Circuits Department at Syracuse seems to be the best approach. This system has apparently been sold to two of the largest USA semiconductor manufacturers, and could thus be available in the future. The minibond system uses chips bonded to a continuous reel of copper strip, dimensioned to allow the use of the entire range of amateur cine film handling equipment. When fully developed this system would give a good transient response coupled with low thermal resistance.

Finally, the influence of the bonding pad can be important only when the noise power is dissipated at the input. This does not occur with present-day logic systems, where failure is rapid because the power is not dissipated at the input bonding pad. All of the above technological considerations are then pertinent only when the integrated circuit geometry, doping levels and other circuitry variables cause the noise power to be dissipated at a spot which can quickly absorb a lot of heat.

The temperature reached after the time CR would be two-thirds of the final value for a step input. Taking the case of a typical integrated circuit having a $100 \times 100\,\mu m$ diode and a chip thickness of $200\,\mu m$, $R = 130\,\Omega$, giving a 'final' temperature of between approximately 1300°C for t.t.l. type circuits and 5200°C for high immunity types. Clearly these figures are very approximate and on the high side because spreading effects are ignored. However, this explains why the high immunity circuits are so prone to damage. It can be seen that as the time-constant is proportional to the chip thickness squared, a reduction in thickness would cause stability to be reached sooner but at a lower temperature. At least, that would be the case if the header surface joined to the chip was at constant temperature by virtue of being a high capacity heat sink. In fact as we can see from the above Table, iron is no better than silicon. A thin sheet of gold, or better still gold-plated copper, between the base of chip and the header would seem to be a good way of improving immunity to high power noise pulses, when associated with a decrease in the chip thickness. If this could be associated with a relatively bulky gold bond to the diode upper surface (using say $50\,\mu m$ wire), the diode area could be minimized. It should be noted that a ceramic pack worsens matters when compared with a metal header (not plastic).

Another way of looking at the thermal transient response problem is from the point of view of the temperature rise induced by the total quantity of heat supplied. For example if the volume of material is v and density d, with C_0 = specific heat in joules per gram cm^3: temperature rise = joules/(vdC_0).

Since we are more interested in millijoules and micrometres, conversion of the terms and substitution of the silicon data gives (where $v = x^3$)

$x = 400\,(J)^{1/3}$ micrometres (with J in millijoules) for a temperature rise of 50 deg C.

In the case of the t.t.l. input circuits, with pulses of approximately 3 mJ, this implies that anything placed at a distance greater than approximately 580 μm will have negligible effect on the transient response. Thus in the case of t.t.l. circuits manufactured from slices having a thickness of 200 μm, the header material is not likely to have much effect on the transient response. However, in the case of a logic system operating at higher input voltages, with a breakdown voltage of say 50 V, where the energy is likely to be closer to 10 mJ giving x = 860 μm, the header could possibly become important. In the case of the bond to the top of the diode, as this is very close to the source of heat, its characteristics are very important. Since aluminium is much more efficient than the silicon in taking up transient heat, a thick pad of aluminium supporting a large gold bond would seem to be as near to the ideal as possible using conventional assembly techniques. The beam lead type of construction would give added benefits, if the beams were sufficiently thick and plated to the entire diode area.

When the heat begins to flow away from the heat source, it initially flows equally in all directions since it does not know that there are any limits. However as soon as the heat approaches the chip edge, a temperature differential will build up, so diverting the heat flow in another direction. Input diodes are by their nature situated along the chip edge, therefore having one side very close to the chip edge. Thus the heat will be diverted from a path away from the edge, running sideways and downwards along the chip edge. This will clearly bring about a tendency for a higher temperature and thus likelihood of failure along the outside edge of the input diodes. As the diodes are usually rectangular, the worst voltage stress and temperature conditions will be met on the outside corners of the diodes. As a result these corners should be rounded off to reduce stress and the chances of destruction for a given device area.

8 Appendix: Power dissipated in input Zener diode with sine wave input

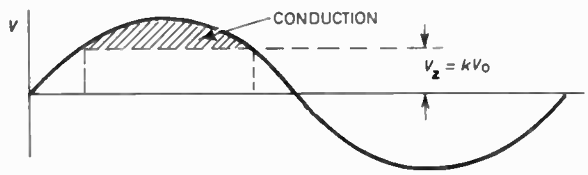

Fig. 11.

Since Zener current.

$$I = \frac{(V_0 \sin \alpha - V_z)}{R},$$

$$\text{power in Zener} = \frac{V_0 \sin \alpha - kV_0)kV_0}{R}.$$

Conduction angles are

$$\alpha_{1,2} = \sin^{-1} k \quad \text{and} \quad \pi - \sin^{-1} k.$$

Therefore

$$\text{average power} = \frac{V_0^2 k}{2\pi R} \int_{\alpha_1}^{\alpha_2} [\sin \alpha - k]\, d\alpha$$

$$= \frac{V_0^2 k}{2\pi R} [2 \cos \alpha_{1,2} + k(2\alpha_1 - \pi)].$$

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IERE News and Commentary

CEI Annual Report, 1972

The Annual Report of the Council of Engineering Institutions just published gives details of the CEI's continuing work in the fields of Industrial Relations, Nomenclature, Metrication, Terotechnology and its affiliation and co-operation with Government bodies. The Report also give details of the work of the Engineers' Registration Board, including details of membership, qualifications, the Register, and the discussion of the mutual recognition and harmonization of engineering qualifications within the EEC.

In his Introduction, Mr. M. W. Leonard, the Secretary of CEI, writes: 'The engineering profession continues to be numerically the largest in Britain. In 1972, the overall membership of the constituent institutions of CEI rose by about 6,000 to just over 312,000.

'A revised Code of Conduct was approved by the Board in July 1972, and the constituent institutions agreed to review their own rules and procedures in order to make the Code effective. The Board of CEI affirmed the individual's freedom to join the Trade Union of his choice, provided that any action arising from his membership of a union did not contravene his professional obligations.

'1972 has also seen the successful launching of the Engineers' Registration Board. At 30th September, 1972, 29 bodies had been admitted to membership of the Technician Engineer Section of the ERB, and 17 to the Technician Section, and more applications are now being considered.

'CEI continues to maintain its links with the principal organizations in the science and engineering community, such as the Royal Society, the British Standards Institution, the Design Council and the Council of Science and Technology Institutes. In conjunction with the CSTI, a Council for Environmental Science and Engineering is to be set up to promote and co-ordinate this interdisciplinary activity among the constituent members of the two founding bodies and other established scientific and engineering organizations.'

There are still twelve Local Committees although there has been discussion in Bristol about the forming of a thirteenth; there are also five CEI Joint Committees of Graduates and Students. Most of these Local Committees have promoted two or three events during the year, aiming at subjects having interest to the majority of Institutions, and several Committees have promoted or co-operated with careers conferences.

The Report does not comment specifically on the financial position of CEI. However, operations during 1972 have resulted in a slightly increased surplus of income over expenditure (£23,000 compared with £21,000 for the previous year) and the adverse balance of the Examinations Fund has been considerably reduced. This is in spite of the appearance under Expenditure of a grant of nearly £11,500 to the ERB.

Copies of the Annual Report for 1972 may be obtained from CEI, 2 Little Smith Street, London SW1P 3DL, price 10p.

Professional Societies' Publishing Association

Twenty-one professional and learned societies that produce journals and other publications, including 11 of the members of CEI, have joined forces to form the Association of Learned and Professional Society Publishers (ALPSP).

The aims of the new Association are to promote and develop the publishing activities of the member societies in disseminating information to their members and to others; to provide facilities for the exchange of publishing knowhow; and to represent the collective views of member societies in discussions with other publishing organizations, with Government Departments, the Post Office, etc.

The Association has already set up five working parties to study copyright (particularly in relation to the illegal photocopying of copyright material), Value Added Tax as it affects publishing by societies, overseas promotion and marketing, postage and distribution costs, and paper and printing costs.

The founder societies of the new Association are:

British Dental Association
British Medical Association
Institute of Fuel
Institute of Marine Engineers
Institute of Physics
Institute of Structural Engineers
Institution of Chemical Engineers
Institution of Civil Engineers
Institution of Electrical Engineers
Institution of Electronic and Radio Engineers
Institution of Mechanical Engineers
Institution of Mining and Metallurgy
Institution of Production Engineers
Iron and Steel Institute
Law Society
Pharmaceutical Society of Great Britain
Royal Institute of British Architects
Royal Institution of Chartered Surveyors
Royal Institution of Naval Architects
Royal Society
Royal Society of Medicine

Observant readers will have noticed that the ALPSP emblem appears on the contents page of the *Journal* to denote the Institution's membership of this Association.

National Electronics Review

The Annual Report of the National Electronics Council for 1971–72 is reviewed in the January-February 1973 issue of *National Electronics Review*, and among the points of interest recorded are the setting-up of a Working Party on Integrated Circuits and the Council's support of the introduction of electronics as an A level subject and of the Link Scheme for the Electronics Industry. Further progress in these and other matters which NEC is considering is given in reports of the Council's quarterly meetings.

The second part of a four-part RAE survey of electronic aids for surface transportation systems—rail systems—is reprinted, while the main addresses at a meeting of the Parliamentary and Scientific Committee on Developments in the Use of Computers, by Sir Robert Cockburn and Mr. Peter Haines are summarized.

Regular features of the *Review* are 'Company Profile' and 'Electronics Research in British Commonwealth Universities'. The subjects in this issue are respectively Hewlett Packard Ltd. and the University of New South Wales.

The annual subscription to *National Electronics Review* is £2.75 (£2.00 to members of the IERE); single copies cost 50p. Orders should be sent to Publication Department, IERE, 8–9 Bedford Square, London WC1B 3RG.

Conference on
'Video and Data Recording'

Organized by THE INSTITUTION OF ELECTRONIC AND RADIO ENGINEERS with the association of The Institution of Electrical Engineers, The Royal Television Society, The Institute of Electrical and Electronics Engineers and The Society of Motion Picture and Television Engineers.

University of Birmingham, 10th to 12th July 1973

PROVISIONAL PROGRAMME

Tuesday, 10th July

Opening Address by Mr. Charles P. Ginsburg, Vice-President, Advanced Development, Ampex Corporation, California, USA.

VIDEO RECORDING: (a) Current Systems

'Video Head Acceptance'.
C. J. C. Laid, R. Vitty and F. Godleman, *Television Centre, London.*

'Track Straightness in Slant Track Video Tape Recorder'.
Yuji Wada, *Sony Corp., Japan.*

'A Wide Band Magnetic Tape Recorder for Broadcast Colour TV Applications'.
R. Matchell, *International Video Corp.*

'Video Tape Editing'.
D. L. Kershaw, D. J. M. Kitson and R. J. Taylor, *BBC.*

'The Digital Timing Correction of Video Tape Recorded Signals'.
S. M. Edwardson, *BBC Research Dept.*

VIDEO RECORDING: (b) Digital Systems

'The Need for New Television Recording Techniques'.
P. J. Darby and P. J. Marshall, *IBA.*

'The Use of Coding Techniques to Reduce the Tape Consumption of Digital Television Recording'.
J. P. Chambers, *BBC Research Dept.*

'Digital Video Recording'.
J. L. E. Baldwin, *IBA.*

CODING AND MODULATION

'Data Compression Techniques as a Means of Reducing the Storage Requirements for Satellite Data—A Quantitative Comparison'.
Dr. L. F. Turner, *Imperial College, London.*

'The Use of Digital Expanded Capacity Delta Modulation for Recording and Delaying Analogue Signals'.
D. W. H. Hampshire, *Portsmouth Polytechnic.*

'A Survey of Retiming and Decoding Methods in Digital Magnetic Recording'.
D. M. Chick and K. F. Walker, *Data Recording Instrument Co. Ltd.*

'High Density Digital Magnetic Tape Recording Using Enhanced-NRZ Coding'.
Jon B. Wells, *Bell & Howell Co., Pasadena, USA.*

Wednesday, 11th July

DATA ACQUISITION AND RECORDING SYSTEMS

'A Simplified Model of the Writing Process in Saturation Magnetic Recording'.
V. A. J. Maller and B. K. Middleton, *ICL.*

'Video-Tape Nuclear Physics Data Recording'.
Dr. R. Nierhaus and Dr. B. G. Taylor, *CERN, Geneva.*

'Tape Transports for Severe Environmental Conditions'.
M. A. Perry, *S.E. Laboratories (Engineering) Ltd.*

'Noise Limitations in Digital Recording'.
J. C. Mallinson, *Ampex Corp., USA.*

'Detection of Error-Causing Defects in Magnetic Recording Disks'.
Jan M. Engel, *IBM Corp., USA.*

'Recent Advances in Wideband Recording Systems'.
J. S. Griffin, *RCA, USA.*

MEDIA

'Low-Temperature Multiplication of Video Tapes'.
Dr. J. Roos, *Philips Research Labs., Netherlands.*

'Wear of Magnetic Heads'.
J. A. L. Potgiesser and J. Koorneef, *Philips Research Labs., Netherlands.*

'Disc Pack Testing—A New Approach'.
C. J. Taylor, *formerly Data Recording Instrument Co. Ltd.*

'Temperature Dependence of Magnetic Properties of Chromium-Dioxide and Cobalt-Doped Gamma-Ferric-Oxide Particles'.
Dr. E. Koster, *Badische Anilin-und Soda-Fabrik, Germany.*

'Chromium Dioxide Video Tape for Third Generation Helical Recorders'.
R. E. Young, *Memorex Corp., USA.*

'Environmental Conditions Effecting Headwear of Magnetic Tape and Heads on Video Recorders'.
R. E. Young, *Memorex Corp., USA.*

'Limitations of Practical Magnetic Masters for Digital Recording Systems'.
H. D. Hughes, *Data Recording Instrument Co. Ltd.*

'Metal Plated Tapes'.
A. G. M. Last, *ICI Ltd.*

'Magnetic Tapes for Contact Duplication'.
Dr. H. Sugaya, *Matsushita Electric Industrial Co. Ltd.*

Thursday, 12th July

OTHER APPLICATIONS

'An Inexpensive "Programmable" Video Tape System for Automated Medical History-Taking'.
Dr. C. R. Evans and P. B. Whittle, *National Physical Laboratory.*

'Instrumenting a Multi-Mode Radar System'.
H. D. Greif, *Hughes Aircraft Co., USA.*

NEW RECORDING TECHNIQUES

'Magneto-Optic Digital and Analog Recording'.
O. N. Tufte, D. Chen, R. L. Aagard and F. M. Schmit, *Honeywell Inc., USA.*

'A Simple Algorithm to Compute the Field-Distribution and Gaploss-Function of an Integrated Magnetic Head'.
J. C. van Lier, *Philips Research Labs., Netherlands.*

'The Production of Nickel Zinc Ferrites by Plasma Spraying'.
C. S. D. Andrews and Dr. I. Preece, *UWIST, Wolfson Centre.*

'Factors Affecting the Design and Construction of Batch Fabricated Recording Heads'.
A. J. Collins and Dr. I. Preece, *UWIST, Wolfson Centre.*

'An Economical Data Store'.
M. E. Whatton, C. Henocq and R. E. Fletcher, *BBC.*

'Magnetic Bubble Memory for Video Display'.
R. S. Quayle, *University of Manchester.*

'Design of Magnetic Disc Store for Mobile Operation'.
A. E. Freeman, A. T. Keefe and L. E. Parker, *Process Peripherals Ltd.*

'Video Frame Storage Using Magnetic Disc Memories'.
A. T. Keefe and J. A. Phillips, *Process Peripherals Ltd.*

'Magnetic Bubble Domain Devices'.
Dr. M. E. Jones, *PO Research Dept.*

'Recording of Video-Signals by One Dimensional Holograms'.
H. Rull, *Siemens AG, Munich.*

'Electron Beam Video and Data Recording'.
Dr. G. S. Plows, *PA Technology and Science Centre.*

For further information and registration forms for the conference, use form on page (xviii) of this issue.

EUROPEAN COMPONENT SURVEY

International Technical Seminars 23rd, 24th and 25th May 1973

in conjunction with

London Electronic Component Show

Sponsors: Institution of Electronic and Radio Engineers and Radio and Electronic Component Manufacturers' Federation

The London Electronic Component Show being held at Olympia from 22nd to 25th May 1973 is a major international event. It is considered therefore most important that exhibitors and visitors are given the opportunity to take part in discussions on technical developments.

A 'European Component Survey' is the theme for the Seminar programme which will be divided into three morning sessions.

Wednesday, 23rd May

'European Component Research'

Chairman: Mr. J. Herbert (*Allen Clark Research Laboratories*)

Speakers: Professor David S. Campbell, Ph.D., C.Eng., F.I.E.R.E. (*University of Technology, Loughborough*)
Mr. Deroy (*Thomson-CSF, France*)
Dr. J. C. van Vessem (*N.V. Philips*)

Thursday, 24th May

'Component Testing and Evaluation'

Chairman: Mr. G. W. A. Dummer, M.B.E., C.Eng., F.I.E.E., F.I.E.R.E.

Speakers: Mr. R. T. Lovelock (*Belling and Lee Limited*)
Mr. A. Lindell (*Mullard Limited*)
Mr. B. Mair (*Plessey Company Limited*)
Mr. J. Dain, M.A., C.Eng., F.I.E.E. (*English Electric Valve Company*)
Mr. David Sawyer (*Erie Electronics Limited*)

Friday, 25th May

'The Evolution of Harmonised Components in Europe'

Chairman: Mr. H. E. Drew, C.B., C.Eng., F.I.E.R.E.

Speakers: Dr. David Reynolds (*British Standards Institution*)
Mr. A. G. Manson (*Plessey Company Limited*)
Mr. M. E. Pulvermacher, M.A., C.Eng., M.I.E.E.

The Seminars will be held at the Royal Westminster Hotel, London, and will commence at 9.30 and end at 12.45. At the end of each Seminar delegates will be taken by coach to lunch in a reserved room at Olympia.

The fee for the three Sessions of the Seminar is £30; this includes attendance, morning coffee, transport to Olympia, admittance to the Exhibition and lunch. Attendance on individual days will cost £12 per day. A reduced charge for IERE members of £25 for attendance at the three Sessions has been arranged with the organizers and members are asked to notify the Institution if they wish to take advantage of this facility; the special registration forms will then be sent.

Further details may be obtained from the organizers: Business Missions International Limited, Conference Department, 24a Trevor Place, London SW7 1LB.

STANDARD FREQUENCY TRANSMISSIONS—February 1973

(Communication from the National Physical Laboratory)

Feb 1973	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
1	+0·2	+0·1	−0·1	728	651·1
2	+0·2	0	−0·1	726	650·7
3	−0·2	+0·1	0	728	649·9
4	−0·1	+0·1	0	729	649·3
5	+0·1	0	0	728	649·6
6	+0·1	0	0	727	–
7	0	0	−0·1	727	–
8	+0·1	+0·1	−0·1	726	649·1
9	0	0	−0·1	726	648·9
10	−0·1	0	−0·1	727	649·2
11	0	−0·1	0	727	650·3
12	−0·1	−0·1	−0·1	728	651·2
13	0	−0·1	−0·2	728	651·5
14	0	0	−0·1	728	651·1
15	−0·1	0	−0·2	729	651·5
16	+0·1	+0·1	−0·1	728	650·5
17	+0·5	+0·1	0	723	650·0
18	−0·1	0	−0·1	724	649·7
19	−0·2	+0·1	0	726	649·0
20	+0·1	+0·1	0	725	647·8
21	+0·2	+0·1	0	723	646·8
22	0	0	0	723	646·4
23	0	−0·1	0	723	647·4
24	0	+0.2	−0·1	723	645·6
25	+0·1	0	0	722	646·3
26	0	0	−0·1	722	646·6
27	0	+0·1	0	722	645·9
28	0	0	0	722	644·5

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; (NTC_{NPL} − Station) = + 500 at 1500 UT 31st December 1968.

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

Other Conferences

Conference on Precision Electromagnetic Measurements

The 1974 Conference on Precision Electromagnetic Measurements will be held from 1st–5th July 1974, at the Institution of Electrical Engineers, London, England.

This will be the 9th Conference in the biennial series which began in 1958 and is the first to take place outside the United States. The change of venue on this occasion is a recognition of the increasingly international character of the Conference, reflected in the theme of the 1974 meeting which is 'the advancement and application of precision electromagnetic measurements against a background of international scientific co-operation'.

The Conference is being organized by the Royal Society and the Institution of Electrical Engineers in conjunction with the following co-operating sponsors: Union Radio-Scientifique Internationale (URSI), the Institution of Electronic and Radio Engineers, the National Physical Laboratory and the Scientific Instrument Manufacturers' Association. The Conference also has the benefit of the advice and support of the permanent sponsors of the CPEM in the United States. The programme will continue to cover the fields traditional to CPEM of d.c. and a.c. electrical measurements, time and frequency and r.f. and microwave measurements, although in the light of the growing diversification in applications of electrical measurement techniques a number of additional topics have been proposed for the 1974 Conference.

Contributions of original papers in the following subjects are invited: d.c. and i.f. measurements; r.f. and microwave measurements; time and frequency; time domain measurements; measurements at millimetre and sub-millimetre wavelengths; laser measurements; and automated and computer-controlled measurements.

Contributions on the following areas of topical interest are also invited: the speed of light; electromagnetic measurements of length and position; measurements on dielectric materials; health and safety; r.f. interference and the electromagnetic environment; precision industrial measurements; management and economic factors in measurement; and international co-operation in the field of precision electromagnetic measurements.

Papers should be submitted to the 1974 CPEM Secretariat, c/o the Conference Department, Institution of Electrical Engineers, Savoy Place, London, WC2R OBL. Authors should contact the IEE 1974 CPEM Secretariat concerning the submission of abstracts for consideration. Special manuscript paper and guide notes will then be supplied, sufficient for an abstract not exceeding 3 A4 pages (approximately 1,500 words of text if diagrams are not included) which should be submitted by 31st October 1973.

SERT Symposium on Maintenance Management

The increasing sophistication and wider use of electronic equipment has made the provision of maintenance an important factor in the thinking of every manufacturing organization. It also means that the service department or organization occupies an important position as the interface between the customer and the manufacturer.

Organization of maintenance requires a combination of managerial and technical skills which the 1973 Symposium of the Society of Electronic and Radio Technicians is aiming to discuss. The main areas into which the subject has been divided are the management function, maintenance policies and philosophies, maintenance aids, and further education and training for maintenance. This whole subject raises many interesting topics for debate, from pay and career prospects to the value of algorithms and fault-finding trees. The whole subject is aimed at electronics staff occupying supervisory and managerial positions in servicing operations.

The Symposium will be held at the University of Nottingham from 9th to 12th July. It will be fully residential and accommodation has been arranged in Cripps Hall. Further details and registration forms can be obtained from the Symposium Secretary, SERT, Faraday House, 8–10 Charing Cross Road, London WC2H 0HP (Telephone 01–240 1152).

Special Rates for IEE Publications

Conference Publications

Members of the IERE may purchase from the Institution of Electrical Engineers the Publications for those Conferences of which the IERE was a co-sponsor.

The following IEE Conference Publications have recently become available:

No. 87: Advances in Marine Navigational Aids.
Special members' rate: £6.80 (Public price £10.40).
No. 88: 1972 International Broadcasting Convention.
Special members' rate: £6.60 (Public price £9.80).
No. 90: Gas Discharges.
Special members' rate: £7.60 (Public price £11.50).
No. 91: The Automation of Testing.
Special members' rate: £5.30 (Public price £8.00).
No. 92: Metering, Apparatus and Tariffs for Electricity Supply.
Special members' rate: £7.70 (Public price £11.70).
No. 95: Satellite Systems for Mobile Communications and Surveillance.
Special members' rate: £5.00 (Public price £7.70).

Periodical Publications

Members of the IERE may subscribe to IEE publications at reduced rates under a mutual agreement. The following are the current (1973) rates (non-member rates in brackets).

Periodical	Reciprocal society members rate
Electronics & Power (Vol. 19)	£11·25 (£15·00)
Proceedings IEE (Vol. 120) paper or microfiche	£34·50 (£46·00)
Combined paper and microfiche	£51·75 (£69·00)
Electronics Record	£12·40 (£16·50)
Power Record	£12·40 (£16·50)
Control and Science Record	£12·40 (£16·50)
Electronics Letters (Vol. 9) paper or microfiche	£20·60 (£27·50)
Combined paper and microfiche	£30·90 (£41·25)

Orders from members wishing to avail themselves of these special rates should be placed through the IERE Publicity Department.

Members' Appointments

CORPORATE MEMBERS

Mr. Sydney Allchurch, O.B.E. (Companion 1964), Director of the British Radio Equipment Manufacturers' Association and Chairman of the Executive Council of the Association, retired at the end of February on attaining the age of 65.

Mr. Allchurch has had a long and distinguished career in the radio and television industry: during the War he served in the Ministry of Aircraft Production, dealing with the supply and installation of special radio and radar equipment for the Royal Air Force and was awarded the O.B.E. He was appointed Secretary of the newly-formed BREMA in 1946 and in 1960 he became Director. Mr. Allchurch was Director and Secretary of the Radio Industry Council from 1967 to this year, Honorary Treasurer of the Radio, Television and Electronics Examination Board, and a member of the boards of the ICETT and BEAB.

Mr. H. N. Holford, B.Sc. (Eng.) (Fellow 1969, Member 1961), who has been at Harlow Technical College since 1962 as Head of the Department of Electrical Engineering and subsequently of the Department of Engineers, has been appointed Vice-Principal of Chesterfield College of Technology and takes up his post on 1st May 1973. Mr. Holford served as a member of the East Anglian Section Committee for a number of years.

Mr. L. A. Smulian, B.Sc. (Fellow 1963, Member 1953), previously General Manager, Transmission and Electronic Exchanges Division, Plessey Telecommunications, has been appointed Managing Director of the newly formed Plessey Microsystems Division. Mr. Smulian has been with Plessey since 1966 when he joined Plessey Radar as Divisional Manager, Display and Data Division. He was previously Industrial Products Officer with British Aircraft Coporation from 1961 to 1966, Managing Director of Gate Electronics, which he founded in 1955, and from 1949 to 1955 Assistant Chief Engineer and later Chief Engineer of the radio and television activities of Ultra Electric Ltd. Mr. Smulian has recently joined the Institution's Papers Committee.

Mr. E. A. W. Spreadbury (Fellow 1967, Member 1937, Associate 1934) and **Mr. K. Tempest** (Member 1963, Associate 1962) have been awarded Honorary Fellowships by the Society of Electronic and Radio Technicians. Both have taken a leading part in the formation of SERT: Mr. Spreadbury was elected the first chairman of the Society in 1965 and on his retirement in 1969 Mr. Tempest succeeded him, serving until last year.

Mr. Spreadbury was for most of his professional life a technical journalist on the staff of what is now the *Electrical and Electronic Trader* and from 1966 to 1969 he was its Editor. Before his retirement he took an active part in Institution affairs, as its representative of the Radio Trades Examination Board, and he also served on the Programme and Papers Committee from 1941 to 1952 and from 1964 to 1969 he was a Trustee of the Institution's Benevolent Fund.

Mr. Tempest has been in technical education for over twenty years; he was head of the Electrical Engineering Department of Carshalton College of Further Education from 1960 until the end of 1972, when he was appointed Vice Principal of the College.

Mr. T. Clements (Member 1966) has been appointed Sales Engineer for the Southern and South-western area of England by Feedback Instruments Ltd.

Mr. D. F. Cook (Member 1969), who was a Senior Project Engineer with Marconi Co. at Chelmsford before he emigrated to Australia in 1970, has been appointed Assistant Marketing Engineer, Microwave, by GTE Australia Pty Ltd., Gordon, New South Wales.

Mr. T. K. Das, B.Sc., M.Sc., Ph.D. (Member 1968, Graduate 1964, Student 1961), who was recently awarded his doctorate by the University of Bradford, has joined Plessey Semiconductors, Swindon, as a Development Engineer for linear integrated circuits.

Mr. R. M. Dare (Member 1968, Graduate 1962, Student 1961) has been awarded the Petrie Memorial Award of the Institution of Works Managers for his Diploma project entitled 'A Feasibility Study into the Profitable Uses of China-Clay Sand'. The Award consists of a silver medal and a cheque for £50.

Mr. Dare is in charge of the Instrument Section of English Clays Lovering Pochin & Co. Ltd., St. Austell, and was formerly with Bristol Siddeley Engines Ltd. as a senior development engineer responsible for the development of electronic equipment used in ground testing Olympus engines for the *Concorde*.

Mr. P. C. Dunstan (Member 1969) has been appointed Senior Electrical Engineer in the Roads Division of the Ministry of Technology, Cardiff.

Mr. G. Galliver (Member 1968) has joined Data Technology Corporation of California to set up a UK and European sales and marketing network, Data Components Division at Crowthorne. Mr. Galliver was previously with Dana Electronics, which he joined after four years with the Solartron Group.

Mr. A. F. Harrison (Member 1959), previously a Principal Engineer with Marconi Space Defence, Frimley, Surrey, has been appointed Head of Traffic Control, Radio, Data and Telephone Groups, Greater London Council (Electrical Services Division).

Wing Cdr. C. Henderson, B.Sc., RAF (Member 1964) has been appointed Officer Commanding Basic Studies Squadron and Senior Education Officer at No. 9 School of Technical Training, RAF Newton, Nottinghamshire. He was previously OC Education Training Squadron, Air Electronics and Air Engineers Training School, RAF Topcliffe, Yorkshire.

Mr. J. V. Lindley (Member 1966, Graduate 1961) who joined the Hydro Electric Power Commission of Ontario in 1968 as a Design Engineer, is now Section Head of Control Services, Atomic Energy of Canada, Clarkson, Ontario.

Wing Cdr. I. Lokanatha, B.E. (Member 1971), formerly Commanding Officer, No. 6 G.T.S., Indian Air Force, Bangalore, has been appointed Command Signals Officer, Headquarters Training Command, Hebbal, Bangalore.

Mr. M. Moore (Member 1968, Graduate 1963) has been appointed General Manager of Semra-Benney (Electronics) Ltd. Mr. Moore previously held the position of Technical Manager with Benney Electronics Ltd., after spending six years with the European Organisation for Nuclear Research (CERN).

Mr. J. W. Parsons (Member 1971, Student 1960) who joined Cable & Wireless in 1954 as a trainee in the Radio and Telegraph Cable Training Department, and has served in various capacities both abroad and in the UK, has now taken up the post of Private Systems Business Executive at their London Office.

Mr. G. J. Pridham (Member 1963), Lecturer in Radio Communication and Electronics at Enfield Technical College, (now Middlesex Polytechnic) has recently been appointed Principal Lecturer.

Mr. A. J. Rickards (Member 1961) has been appointed Deputy Head of Engineering with Thames Television. Mr. Rickards, who joined ABC TV in 1960, became Head of Engineering, Planning and Installation for Thames in 1968, and subsequently Head of Development and Maintenance Department.

Mr. H. S. Sanders (Member 1971, Graduate 1965, Student 1964), has been promoted to the post of Group Manager, Test Equipment Engineering, with Xerox Data Systems, El Segundo, California. Mr. Sanders was formerly Senior Member of the Technical Staff in the Sherman Oaks Branch of the same company.

Mr. I. A. M. Shafik, B.Sc. (Member 1970, Graduate 1966) has taken up an appointment with the newly established Telecommunication and Broadcast Training Center of the Ministry of Communication of Saudi Arabia, in Riyadh.

Mr. R. C. Scrivens (Member 1972, Graduate 1967) who for the past two years has been working with the European Space Research Organization, Noordwijk, as Payload Engineer in the Sounding Rockets Division, has been appointed Assistant Signals Officer with the Civil Aviation Authority, Gatwick Airport.

Mr. P. B. Stott (Member 1971), who was formerly with Elliott Automation Radar Systems Ltd., Mobile Radio Division, has joined H.M. Communications Centre at Wolverton, Bucks, as a Higher Scientific Officer.

Wing Cdr. A. T. Vacquier, RAF (Member 1953) has now been posted as Wing Commander Training 2 (Electrical Engineering), to HQ No. 24 Group, RAF Rudloe Manor. Wing Cdr. Vacquier was previously Group Electrical Engineer at HQ 11 (F) Group, RAF Strike Command, Stanmore.

Mr. P. B. Walker (Member 1968, Graduate 1966) has been appointed Senior Technical Sales Representative with Smiths Industries Ltd. at Cheltenham. Mr. Walker has for the last few years represented the company in America and returned to the UK last year.

NON-CORPORATE MEMBERS

Mr. I. J. Browning (Graduate 1971) who joined British Railways Divisional Signal and Telecommunications Department in 1967, has now been promoted Senior Technical Officer in the Chief Signal and Telecommunications Engineer's Department.

Mr. D. Chadwick (Graduate 1970) has taken up the post of Pormotion Superintendent with Connolly's (Blackley) Ltd., Manchester. He was previously with British Insulated Callender's Cables Ltd., latterly as a Process Development Engineer.

Mr. E. V. Fingall (Graduate 1970) is now a Telecommunication Engineer with the Ministry of Communication and Works, Pine St. Michael, Barbados.

Mr. V. K. Fashola, M.Sc. (Graduate 1970, Student 1963) has joined the Nigerian Ministry of Post and Telecommunications as an Engineer in the Exchange Planning Section. He was previously with the British Post Office as an Executive Engineer in the Operational Programming and Planning Division.

Mr. A. J. Fisher (Graduate 1967), formerly in Technical Sales Department of RSM Transformers Ltd., Redditch, Worcs., has been promoted Chief Projects Engineer.

Mr. M. R. Franklin, B.Sc. (Graduate 1972, Student 1968) is now a Digital Systems Engineer in the Space Division of Messerschmitt-Boelkow-Blohm at Munich. For the past four years he has been with Plessey Communications Research, Ilford, as a Development Engineer.

Mr. T. M. A. Gilly (Graduate 1970), who was formerly in the Communication Department, Getty Oil Co., Kuwait, has joined VBB-Vattenbyggnadsbyran (Consultant Engineers), Kuwait, as Control and Communication Engineer working on the Kuwait Water Distribution Project.

Mr. D. Hewson (Graduate 1932) has been appointed Lecturer in Telecommunications in the Department of Technical Education and Vocational Training, of the Zambia Institute of Technology, Luanshya.

Mr. S. D. Jetha (Graduate 1968) who was a Radio and Electronics Engineer with Philco-Ford Corporation in Tehran, Iran, has been transferred to the Corporation's plant in Pennsylvania, USA.

Mr. R. J. Jones (Graduate 1970, Student 1967) has been appointed Assistant Chief Test Engineer with Devices Instruments Ltd. (Medical), Welwyn Garden City, Herts. He was previously Test Engineer (NATO), with Hawker Siddeley Dynamics, Hatfield.

Mr. M. Leeston-Smith (Associate 1950) has joined the South African Broadcasting Corporation, Johannesburg, as an Instructor in various production and engineering aspects of television.

Mr. G. J. McAllister (Graduate 1966), formerly Senior Product Engineer with Honeywell Controls Ltd., Motherwell, Lanarkshire, has now joined the Central Electricity Generating Board in Cheltenham as Assistant Engineer (Control and Instrumentation).

Mr. R. McBride (Graduate 1970) has been appointed Production Engineer with Hewlett Packard Ltd., West Lothian; he was previously with Plessey Telecommunications Ltd.

Mr. O. Okuwa (Graduate 1971), who was previously a Senior Engineer with Comsac Nigeria Ltd., in Lagos, has been promoted to Area Engineer/Manager, Midwest State, at Warri, Nigeria.

Major S. R. A. Stopford, MBE, Royal Scots Greys (Graduate 1970) has been promoted to Lieutenant-Colonel and posted to the Ministry of Defence (Army), London.

Mr. B. B. Streater (Graduate 1971) has joined Diablo Systems SA as Sales Manager, UK and Scandinavia. For the past four years Mr. Streater has been with Data Recording Instrument Co. Ltd. Staines, heading a group in the Sales Services branch responsible for technical publicity and handbooks.

Mr. A. Tinkler (Graduate 1967) who joined the G.P.O. Engineering Department in 1954 as a trainee and was subsequently promoted to Temporary Assistant Executive Engineer, responsible for maintenance and installation of telecommunication plant has now been promoted to Executive Engineer employed on internal planning duties.

Mr. J. C. Walker (Graduate 1971) has been appointed Instrument Development Engineer with Engelhard Industries Ltd., Cinderford, Gloucestershire. He was a Project Engineer (Electronics) with Arcon Instruments Ltd.

OBITUARY

The Council has learned with regret of the death of the following member, advice of which has only just been received.

Douglas Victor Carroll, M.B.E. (Fellow 1970) served in the Royal Canadian Navy from 1925 to 1950, rising from Wireless Operator to Lieutenant Commander. Senior posts included those of officer-in-charge of Halifax Radio and of the Radio Engineering Division at RCN HQ, where he served as chairman of an Inter-Services committee on the introduction of tape relay systems. He was awarded the M.B.E. for this work.

On retirement from the Navy Mr. Carroll was appointed Vice-President and Managing Director of Technical Materials Corporation (Canada), becoming President in 1952. He was active in Canadian industrial affairs, being a Director of the Electronic Industries Association of Canada, and chairman of its Government Action Committee. His contributions to professional life were considerable: he was an active member of the Ottawa Section of the IEEE and was its chairman in 1968, while in 1906 he was Chartered President of the Armed Forces Communications and Electronics Association in Ottawa. Mr. Carroll held patents for circuit developments and magnetic compasses and his publications included a well-received primer on single-sideband techniques.

Mr. Carroll, who died in Manotick, Ontario, in September last, aged 63 years, left a widow and two young children.

Letters to the Editor

The Institution's Council does not necessarily agree with views expressed by correspondents.

Correspondence of a technical nature, or on any matter of interest to electronic and radio engineers, is welcomed.

From: **J. B. Butcher**, M.A., M.Inst.P., C.Eng., M.I.E.E. and **K. J. Dean**, M.Sc., Ph.D., F.Inst.P., C.Eng., F.I.E.E., F.I.E.R.E.

Education in Semiconductor Technology

While enjoying greatly the excellent paper by Dr. K. J. Dean* in the issue of January/February 1973, I cannot let pass without comment two points in his paper.

Dr. Dean states that teaching facilities for semiconductor technology are set up with the primary purpose of 'training' (*sic*!) people for the semiconductor manufacturing industry. He therefore implies that teaching in this field is an optional subject relevant only to a small minority of students. Further, he quotes one reference as saying that the technology is too difficult for electrical engineers!

Both of these points are 'old chestnuts' in electronics education. At Enfield we believe that *every* student of electronics must have at least a working knowledge of semiconductor and other microelectronic technologies if he is to be capable of communicating with manufacturers and able to make decisions when faced with the bewildering array of devices currently available and constantly increasing.

We have, for over three years, successfully taught these techniques to students on H.N.C. and undergraduate courses (to the great benefit of their final-year projects) and we have had over 250 students from eight countries attending short courses. Many of these had qualifications other than in electronics and none claimed to find the techniques difficult to comprehend.

* 'The influence of semiconductors on the teaching of electronics', *The Radio and Electronic Engineer*, 43 No. 1/2, pp. 140–8, January/February 1973.

I maintain firstly, that microelectronic technology must be taught to all students of the subject in this day and age, and secondly, that the supposed difficulty of the subject is a myth put about either by those who have not studied it or by those who have, but see some personal or commercial advantage in maintaining the aura of mystery and exclusiveness often associated with new technologies.

J. B. Butcher

Head, Microelectronics Centre,
Middlesex Polytechnic,
Queensway,
Enfield EN3, 4SF
13th February 1973

What Mr. Butcher does not say is perhaps as important as what he does. The comment that semiconductor technology is too difficult for electrical engineers was made in 1968 at the Cambridge Conference on 'Electronics in the 1970s' by D. F. Dunster, who is in charge of a similar Microelectronics Centre to that which Mr. Butcher manages. Thus the comment comes from his side of the fence, so to speak.

Whilst I agree that every student of electronics must have some knowledge of semiconductor and other microelectronic technologies, I do not believe this extends to experiences gained in a microelectronics centre where diodes, transistors and simple integrated circuits are being manufactured. The capital and running costs of such an establishment do not, I believe, warrant this except for those students who are going to be employed in manufacture of semiconductor products. One can learn all one needs to know from lectures and text-books to appreciate the limitations and advantages of these devices.

It is not primarily a matter of the difficulty of the subject or of any aura of mystery, but simply that this in detail is something which the average electronics engineer just does not *need* to know.

K. J. Dean

Principal,
South East London Technical College,
Lewisham Way,
London SE4 1UT.
19th February 1973

Streamlined Job Service for 1973

A fast and efficient job service for managerial and professional people, has just been started by the Department of Employment in the United Kingdom. Based on the Department's Professional and Executive Register, it is being restyled to link employers' needs with qualified applicants quicker than ever before.

From March 1st, the service will be known as Professional and Executive Recruitment and it operates on a fee-paying basis for employers seeking staff. But there are no charges to people looking for jobs. Fees for the improved service start at £30 for annual salaries below £1,000 and then range from 5% to a maximum of 8% for salaries over £4,000.

The new service is using computer techniques to store, match and print-out details of vacancies and candidates. The information will be subject to careful checks by trained staff at local offices. These modern methods will enable a sophisticated service to be offered at some 40 offices in England, Wales and Scotland. A full promotional campaign will be professionally directed to attract first rate vacancies and candidates. A feature will be the individual and personal service available to employers for candidates.

Professional and Executive Recruitment is headed by Mr. Dewi Rees, who was appointed from industry to direct this restyled service which, although independently organized, is still part of the Department's employment services.

With no salary-level bars, the specially trained staff can undertake to fill and find jobs through the complete spectrum of professional and executive appointments, including those for newly qualified people. Staff's efforts will be particularly engaged in dealing with the middle salary jobs up to about £6,000 a year. Candidates will find there are no formalities. Provided they have the relevant qualifications or experience, they can simply ask for a form, fill it in, and it is then computer matched against vacancies until a suitable job comes up.

To support this new service and to publicize the many other activities of the Department of Employment, such as training and safety, a free monthly tabloid, 'DE News' has been launched. Copies may be ordered from Department of Employment (Inf 3), 162/8 Regent Street, London W1R 5TB.

INSTITUTION OF ELECTRONIC AND RADIO ENGINEERS

Applicants for Election and Transfer

THE MEMBERSHIP COMMITTEE at its meeting on 22nd February 1973 recommended to the Council the election and transfer of 37 candidates to Corporate Membership of the Institution and the election and transfer of 12 candidates to Graduateship and Associateship. In accordance with Bye-law 21, the Council has directed that the names of the following candidates shall be published under the grade of membership to which election or transfer is proposed by the Council. Any communications from Corporate Members concerning these proposed elections must be addressed by letter to the Secretary within twenty-eight days after the publication of these details.

Meeting : 22nd February 1973 (Membership Approval List No. 155)

GREAT BRITAIN AND IRELAND

CORPORATE MEMBERS

Direct Election to Fellow

DAVIES, Eric Howard, Colonel *Aldershot, Hampshire.*

Transfer from Graduate to Member

BARNES, Colin *Streetly, Staffordshire.*
BETTS, Alan John *Romsey, Hampshire.*
BROCKLEBANK, John Anthony *Aylesbury, Buckinghamshire*
BROCKLEHURST, Arthur *Marlow, Buckinghamshire.*
BROWN, Robert William. *Porthcawl, Glamorgan.*
CATANIA, Victor George *Winkfield Row, Berkshire.*
DUTTON, Robin Howell *Moseley, Birmingham.*
FIELDING, Maurice Edward *South Shields, County Durham.*
GALL, Alan Martyn *Harlow, Essex.*
GEORGE, Leslie James *South Ockendon, Essex.*
GIRLING, Keith Ivor *Reading, Berkshire.*
HAMBLEY, David Thomas *Wetherby, Yorkshire.*
HEASMAN, Brian Reginald *Eastbourne, Sussex.*
HILLS, Keith Cyril *Romford, Essex.*
IBBOTT, Christopher James *Frimley, Surrey.*
McCRAE, Colin *Hamilton, Lanarkshire.*
METCALFE, Eric Antony *Garston, Hertfordshire.*
MULTANI, Shamoo Gordhandas *Poole, Dorset.*
ROBBINS, David *Bushey, Hertfordshire.*
ROBSON, Dennis James *Stevenage, Hertfordshire.*
ROMM, Stuart Malcolm *Gatley, Cheshire.*
SALISBURY, Raymond James *Burntwood, Staffordshire.*
SCOTT, John Rhead *Motherwell, Lanarkshire.*
SHEEKEY, Bartholomew *Chesham, Buckinghamshire.*
SNOWBALL, Malcolm Robert *South Shields, County Durham.*
STONE, David Edward *Great Yarmouth, Norfolk.*

Direct Election to Member

APPLETON, Roger Frank *Hillingdon, Middlesex.*
DIBBLE, Trevor John *Glasgow*
HAGUE, Brian John, Captain REME *Crowthorne, Berkshire.*
WALTERS, William Peter Royston, Instr, Lt. Cdr., RN, B.Sc., M.Sc. *Hayling Island, Hampshire.*

NON-CORPORATE MEMBERS

Transfer from Student to Graduate

KIMBERLEY, Robin *Bourton-on-Dunsmore, Rugby, Warwickshire.*

Direct Election to Graduate

CROSBY, Paul Russell *Cookridge, Leeds.*
HUME, Daniel Joseph Hunter, B.Sc. *Duntocher, Dunbartonshire.*
HUTCHERSON, Peter, B.Sc. *Isleworth, Middlesex.*
McKAY, Graham *Biggleswade, Bedfordshire.*
RISEBOROUGH, Michael Keith *Biggleswade, Bedfordshire.*
SINGH, Sarwan *Wolverhampton, Staffordshire.*
WHICHER, Christopher John *Mitcham, Surrey.*
LEUNG, Hung-Chee *Manchester.*

Direct Election to Associate

GIDDINGS, Stanley Fredrick *Orpington, Kent.*

STUDENTS REGISTERED

BERRY, Michael Richard *Warley, Worcestershire.*
BUTLER, Graham *Nottingham.*
CARVER, John Ernest Llewellyn *Rhiwbina, Cardiff.*
CRANE, Roger Alan *Nottingham.*
GAUNT, Christopher John *Nottingham.*
GIBBS, Christopher Charles *Nottingham.*
HARRIS, Michael John *Nottingham.*
LIEW, Thomas Nyuk Pen *Swansea.*
METCALFE, Michael *Nottingham.*
NATION, Christopher Frederick *Taunton, Somerset.*
NORLING, Richard Arthur *Llandaff North, Cardiff.*
PARRY, Jonathan Robert *Nottingham.*
PEPPER, Adrian Charles John *Worcester.*
ROTSETIS, Dimitri *Northolt, Middlesex.*
SAALABI, Ali Akbar *Surbiton, Surrey.*
SHANKS, Peter Graham *Nottingham.*
SHOA, Hormuz *London, N.4.*
SLATER, Edward John *Walsall, Staffordshire.*

OVERSEAS

CORPORATE MEMBERS

Transfer from Member to Fellow

RAJKUMAR, Gnanapragasam Manuel Pillai *Colombo.*

Transfer from Graduate to Member

EHIAGWINA, Pius Osoba *Lagos, Nigeria.*
GRIFFITH, Peter Sherwin Leslie *San Fernando, Trinidad.*
GUPTA, Dilip Kumar *Calcutta.*
ODUMBONI, Musibau Adeyanju *Lagos.*
PIKE, William Frederick Herbert Edmund *BFPO 34*

NON-CORPORATE MEMBERS

Direct Election to Associate

IYALLA, Winful Vincent *Lagos.*

Transfer from Student to Associate

EDIRISINGHE, Mudiyanselage Siripala, Lieut. *Trincomalee, Sri Lanka.*

Student Registration

AHMED, Hassan Eltom *Khartoum, Sudan.*
CHIN, Yook Kong *Kuala Lumpur, Malaysia.*
OLUGBODI, Timothy Oladejo *Ile-Ife, Nigeria.*
ROUSSOS, Photis Artemis *Khartoum, Sudan.*
WONG, Chi Fun *Kowloon, Hong Kong.*

British Telecommunications Systems in Bolivia

One of the highest microwave-radio telecommunications transmission systems in the world is to be installed in the Bolivian High Andes by a British company, GEC Telecommunications Limited. The system will modernize the Bolivian trunk telephone communications network and provide facilities for television to be relayed from La Paz, the capital, to other cities, and for the transmission of telex and data services. It will also allow international connexions to similar systems in the neighbouring countries of Peru, Argentina, Chile and Brazil. Reference was made in *The Radio and Electronic Engineer* in August 1971 (p. S.10) to similar GEC Telecommunications installations ordered for Chile and other microwave radio telephone networks have been provided for Costa Rica, El Salvador and Peru. These together represent a total bearer-channel length of 21 600 km (13 500 miles) in Central and South America.

This latest turnkey contract, worth £3·25M, has been placed by Empresa Nacional de Telecomunicaciones of Bolivia and covers surveying, planning, installation, commissioning, the training of local staff and responsibility for all associated civil engineering works. The company is also responsible for operating and maintaining the system during the first 12 months operation.

The broadband microwave-radio system, which is designed to meet CCIR recommendations, will cover a route length of 900 km (560 miles) and will provide high-capacity trunk telephone circuits between the largest population centres, La Paz, Oruro, Cochabamba, Santa Cruz and Copacabana. Complete by semiconductored 2 GHz microwave-radio and associated multiplex equipment will provide a bothway channel with a capacity of 960 telephone circuits and a stand-by or protection channel which can also be used to carry television programmes. The network, comprising 5 terminal stations and 10 unattended repeater stations, will be equipped with a comprehensive supervisory indication and remote control system. Equipment reliability is of prime importance particularly as 10 of the 15 radio stations will be sited at altitudes well over 3600 m (12 000 ft) in temperatures varying between −25°C (−13°F) and +40°C (+104°F) and with humidity varying from as low as 2% up to almost 100%.

Forthcoming Institution Meetings

Mid-April—June 1973

London Meetings

Wednesday, 25th April

AEROSPACE, MARITIME AND MILITARY SYSTEMS GROUP MEETING

Electronic Aids to Position Fixing
By D. J. Phipps (*Decca Survey Ltd.*)

MEETING POSTPONED UNTIL AUTUMN. Date to be announced.

Wednesday, 2nd May

JOINT IEE/IERE COMPUTER GROUP AND I.MECH.E. COLLOQUIUM

Electromechanical Problems of Computer Systems

COLLOQUIUM CANCELLED.

Wednesday, 2nd May

COMPONENTS AND CIRCUITS GROUP COLLOQUIUM

Electronic Devices in the Home

COLLOQUIUM POSTPONED until Autumn. Details to be announced.

Wednesday, 9th May

EDUCATION AND TRAINING GROUP MEETING

Preparing Engineers and Managers for Design Specification

By Professor David Pilfold (*Queen's University, Ontario*)

IERE Lecture Room, 6 p.m. (Tea 5.30 p.m.)

The lecture will be applicable to all branches of engineering and will be kept at a general level, but examples will be taken from, and emphasis placed upon, the electrical and telecommunications industries, particularly segments involved in innovative design.

The importance of improved design specification and standardization will be stressed in view of the following trends: adoption of design methodology; increased design responsibilities; more sophisticated contracts. Some of the problems that engineers and managers face in design specification will be discussed, and techniques which can be learnt for overcoming these will be described.

Tuesday, 15th May

COMPONENTS AND CIRCUITS GROUP COLLOQUIUM

Electrical Connectors—Applications and Reliability

IERE Lecture Room, 2 p.m.

Advance Registration necessary. Free to members of the IERE and IEE; £1.50 to non-members.

Chairman's Introduction
By H. G. Manfield (*RRE*)

The Assessment of Connector Reliability
By R. T. Lovelock (*Consultant*)

Plastic Materials for Connectors
By D. Taylor (*Amphenol*)

A User's View of Connectors
By J. Hodson-Smith (*Hawker Siddeley Dynamics*)

Connector Testing and Evaluation for Reliability
By G. M. Matthews and R. Britton (*Plessey*)

Standardization of Connectors and BS 9000
By A. W. Eva (*Consultancy PR Services*)

Wednesday, 16th May

AEROSPACE, MARITIME AND MILITARY SYSTEMS GROUP MEETING

Meteorological Telecommunications Systems Engineering

By C. E. Goodison (*Meteorological Office*)
IERE Lecture Room, 6 p.m. (Tea 5.30 p.m.)

Wednesday, 23rd May

COMMUNICATIONS GROUP MEETING

Optimum Linear and Non-linear Transversal Equalizers
By Dr. A. P. Clark (*Loughborough University of Technology*)

IERE Lecture Room, 6 p.m. (Tea 5.30 p.m.)

The concepts of linear vector spaces are used to derive the optimum linear and non-linear equalizers for a known baseband channel. The equalizers are of the conventional type using transversal filters. The better performance of the non-linear equalizer is explained by considering this as a combination of separate linear and non-linear filters.

Thursday, 14th June

JOINT IERE/IEE MEDICAL AND BIOLOGICAL ELECTRONICS GROUP ONE-DAY COLLOQUIUM

X-Ray Image Intensifiers
Registration necessary. Further details to be announced.

North Western Section

Thursday, 3rd May

ANNUAL GENERAL MEETING at 6.15 p.m.

Followed at 6.45 p.m. by

Facsimile—A Review

By J. Malster and M. J. Bowden (*Rank Xerox*)

Lecture Theatre RG7, Renold Building, UMIST, 6.15 p.m. (Tea 5.45 p.m.)

The paper reviews the growth and potential of facsimile as an alternative communication medium over voice grade public switched and leased telephone lines. The problems confronting facsimile manufacturers in relation to the common carriers, standardization, and economics are discussed. The relative merits of current techniques are considered and future trends are outlined.

Yorkshire Section

Friday, 27th April

ANNUAL GENERAL MEETING
at University of Leeds, 7 p.m.

Followed by Ladies Evening and Buffet Supper

South Midland Section

Tuesday, 17th April

New Radio Receiver Developments

By Professor W. Gosling (*University College of Swansea*)

To be followed by ANNUAL GENERAL MEETING

G.C.H.Q., Oakley, Cheltenham, 7 p.m.

South Western Section

Monday, 7th May

ANNUAL GENERAL MEETING

Royal Hotel, Bristol, 7 p.m.

Southern Section

Saturday, 12th May

Visit to Atomic Energy Establishment, Winfrith. 2–5 p.m.
Applications to attend to Hon. Secretary, Southern Section.

Services Qualifications and T.Eng. Requirements

It has been pointed out that the statement regarding ERB evaluation of Service qualifications in the January-February 1973 issue (p. 172) may be misleading, particularly where the Royal Navy is concerned, in so far as the wording is open to the interpretation that no R.N. qualification met *any* of the requirements for T.Eng. registration. This is not the case. The courses investigated fully met the training and experience requirements; it was only in respect of the academic content that the Qualifications Committee Working Party felt the requirements for T.Eng. registration were not met. No doubt the publication of the relevant DCIs will make the position clear.