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Fibres, Lasers and Communications

ONE of the most interesting potential applications of lasers is to communications and, since 1960, many attempts have been made to use them for this purpose. The potentialities are, of course, considerable—with carrier frequencies many orders of magnitude higher than those currently in use the available bandwidth should also be much larger. The first problem is the medium through which the laser beam is to be transmitted. Within the earth's atmosphere the effects of temperature gradients, turbulence and the weather prevent reliable communication over distances of more than a few kilometres. In outer space where spreading of the optical beam due to diffraction is the only limitation to propagation then signalling over vast distances is possible, but obviously there is going to be no large-scale application in this direction for a little while yet. Even so detailed studies have been made of laser systems for communications with stationary satellites and deep-space probes.

In any terrestrial system some method of protecting the transmitted beam from the vagaries of the atmosphere is essential. One method which has been extensively investigated in the USA is to use a protective pipe, either evacuated or kept accurately at a constant temperature along its length, and periodic refocusing with a sequence of lenses. The transmission loss attainable has been shown to be low, 1 dB/km, and the bandwidth very high, limited only by the modulation techniques, but since the pipe has to be kept almost optically straight the system is inflexible and very costly.

However in 1964 some preliminary experimental and theoretical studies were begun in England at Standard Telecommunications Laboratories on the possibility of using glass fibres as a transmission medium. These, as well as discussions with F. F. Roberts of the Post Office Research Station, led to the publication of the classic paper by Kao and Hockham in 1966 which gave a preliminary materials and systems analysis. As a result work almost immediately started on an investigation of single-mode fibres at the Post Office Research Station and STL and on multimode fibres at the Signals Research and Development Establishment with supporting programmes at the Universities of Sheffield and Southampton. Similar activities followed a year or so later in other European countries and Japan, and subsequently in the USA. The basic idea was to use a cladded, or a graded-index, glass fibre as an optical waveguide and a modulated laser as the carrier source. In principle the fibre diameter need only be a few tens of wavelengths, i.e. a few tens of micrometres, in order to give bandwidths of several gigahertz over several kilometres, compared with coaxial cables of more nearly ten millimetres in diameter giving a bandwidth of about 50 MHz over a kilometre. Thus for the same space occupied by a coaxial cable a change to fibres could give an enormously greater total bandwidth and information-carrying capacity. Glass fibres of such small diameter are flexible and could simply replace coaxial cables in existing cable ducts. The installation cost would therefore be minimal.

At the time the idea seemed a little optimistic since a typical commercial fibre attenuation was in the region of 1000 dB/km and to reduce it by the two orders of magnitude necessary implied reaching a level of some impurities of less than 1 part in 10^8. Since glass technologists had been active for over 2,000 years without getting anywhere near this target it seemed a tall order. However progress in the past seven years has been rapid. Already at least five different types of fibre with losses below 20 dB/km have been produced. The seemingly difficult problem of joining single-mode fibres, requiring a lateral precision of much better than 1 μm, has been solved at the laboratories of AEG-Telefunken in a simple, easily-operated jig. At the same time, whereas multimode fibres were originally thought to be capable of only a small bandwidth due to multi-path dispersion, it has now been shown that when operated with comparable launching conditions, and with the same sources, as single-mode fibres they are capable of bandwidths of as high as 1 GHz over 1 km with a suitably launched Gaussian beam, even with core diameters in the range 30 μm–100 μm. They have the advantage also, of course, that they are easier to make, couple into, join and handle than single-mode fibres. In addition a practical demonstration, by the Southampton University group, of the first transmission of a live colour television programme over 1·25 km of liquid-core fibre, with less than 10 μW of input power from a light-emitting diode, was given on BBC television a year ago and, incidentally, the same link has now also been shown on French and German television networks. The practicability of a fibre link in simple inexpensive form has therefore been shown. One can go further and say that in seven short years a tentative idea has been translated into a nearly practical reality and, in so far as a continuing process of development can be categorized into chronological steps, the first stage of optical fibre communications has nearly been completed. The concept has been all but shown to be technically feasible.

It is pleasing to note that as well as providing what was perhaps a key paper, the United Kingdom has been in the forefront of developments in optical fibre communications. While not wishing to embark on any kind of league table it is worth mentioning a few significant British achievements such as a pulse-modulated semiconductor laser transmitter operating at a bit rate of 1 GHz and l.g.r. (localized gain resonator) semiconductor lasers with threshold current densities as low as 575 A/cm^2 (STL); compound glasses with bulk losses of $\sim$ 20 dB/km (Post Office Research Station, Sheffield University, STL); the lowest loss liquid-core fibres which are also the cheapest and have the largest measured bandwidth (Southampton University); and just recently the impressive announcement of a compound glass (as distinct from silica) with a transmission loss of only 16 dB/km (Post Office Research Station). There have been equally important contributions in other countries of course. I hope that our friends abroad will forgive the above comments which are directed at British readers who only too frequently assume that, almost by definition, all good development ideas must originate elsewhere!

The next stage of progress will be no less interesting and challenging. A laser source would give a larger transmission distance and, if it can be made reasonably monochromatic, a much larger bandwidth than a light-emitting diode. Semiconductor injection lasers would be ideal in that they are very small, efficient and operate at a wavelength where fibre losses are low. Unfortunately they do not yet have a sufficiently long lifetime, nor a narrow-enough linewidth, but recent progress has been both steady and hopeful. The handling of fibres has turned out to be less difficult than had been feared and while suitable methods of jointing, and coupling to sources and detectors, need to be devised they are not likely to be a stumbling block in system development. A more serious problem may be protection and armouring of fibres—an area where much work needs to be done. Again little is known about long-term effects in low-loss fibres. For example will a silica-based fibre take up water from the atmosphere over a period of years? If so the effect could be serious since there is an OH overtone absorption band at 0·95 μm, the wing of which can cause an appreciable transmission loss at the nearby semiconductor laser wavelength of 0·9 μm. There is also a need for a much clearer understanding of propagation in fibres. It is perhaps remarkable that some fibres have been made so accurately that although they are highly-overmoded waveguides, effectively single-mode propagation has been observed in them over distances of hundred of metres. Nevertheless the effect of bends, stress, birefringence and inhomogeneities needs to be studied in detail.

Finally there is the crunch question of system economics. The best research and development in the world is of no avail if it produces something which is technologically first-rate but for which there is no urgent need, or which is too expensive. That there is a need for a line communication system of greatly increased capacity is widely recognized; for example the demand on the trunk telephone networks of most advanced countries is doubling every six years or so. Optical fibres could probably satisfy this need but it is in competition, not only with the existing system with which it must be compatible, but with the overmoded millimetre waveguide and new forms of coaxial cable. So often when a serious challenge appears a 'conventional' device suddenly improves out of all recognition—as Samuel Johnson said, '. . . when a man knows he is to be hanged in a fortnight it concentrates his mind wonderfully'! The final verdict will thus be based on cost and acceptability and will not be known for some time. To those of us enjoying the excitement and the rigours of the chase it may perhaps be just as well.

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Optical fibres for communications

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SUMMARY

This paper presents a review of the state of the art of optical fibres and their potential use as waveguides for long-distance transmission of light. The physical principle of light guidance in fibres is discussed and the advantages and short comings of fibres are pointed out. The properties of single and multimode fibres are reviewed.

1 Introduction

Throughout its history wireless broadcasting and carrier telephony through cables have experienced a continuing trend to utilize higher frequencies. To increase channel capacity communication technology tends to use the highest possible frequencies that the state of the art can provide. With the invention of the laser in 1960 coherent sources of electromagnetic radiation reached the visible region of the spectrum. It is thus only natural that attempts at using lasers for communications purposes were undertaken almost immediately. In the absence of suitable light waveguides the first experiments involved 'radio' systems—light transmission through the atmosphere.[1] However, unlike radio systems at longer wavelength, light 'radio' is severely limited by scattering of light by fog, rain and snow. The only promising applications for light 'radio' systems are special applications in arid regions of the world or in the vacuum of outer space. Other applications for optical 'radio' system include communication links that do not require high reliability, or can utilize repeaters that are spaced so closely that the link functions even during heavy precipitation. For reliable light communication under terrestrial conditions some type of light waveguide is desirable to protect the light beam from atmospheric disturbances.

Early attempts at light guidance over long distances turned to beam waveguides that utilize the focusing properties of lenses or curved mirrors to counteract the natural tendency of light to spread apart by diffraction and that are able to guide the light beam around bends.[2, 3] Such lens waveguides can be developed into workable systems. However, their costs are high since they require careful alignment and possibly electronically controlled positioning of the lenses.[4] In addition, an outer shield is always needed to keep precipitation out of the light path and to avoid temperature gradients transverse to the direction of the light beam.[5, 6] However, beam waveguides may become attractive for extremely high capacity systems utilizing a larger fraction of the available bandwidth.

It has long been known that light can be guided in thin fibres made of transparent dielectric materials.[7] However, most dielectric materials are very lossy even though they may appear clear and perfectly transparent if only a small piece is viewed. Losses of 1000 dB/km are not at all unusual for even high quality optical glasses used to fabricate lenses.[8] Charles Kao at STL first realized that the high loss of most glasses is not an inherent property but is caused by impurity ions that, in principle, can be eliminated from the material.[9] He made measurements proving that fused silica (quartz) may have losses as low as 80 dB/km.[10, 11] The last doubts about the usefulness of fibres for long-distance transmission of light were dispelled in 1970 when Corning Glass Works produced a fibre whose loss was 20 dB/km at the frequency of the helium-neon laser, 0·6328 μm.[12] Since this spectacular breakthrough fibre losses have been reduced steadily. The best fibres at present[13] have loss minima of 4 dB/km and theoretical predictions promise a further reduction to about 2 dB/km. The

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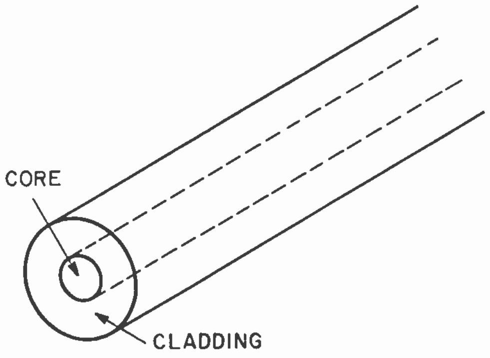

Fig. 1. Schematic of a clad optical fibre.

losses of optical fibres are thus much lower than those of conventional coaxial cables at gigahertz frequencies and approach the losses of the millimetre waveguide operated in the circular electric mode.

Optical communications systems using fibres as waveguides promise to be economical even if each fibre carries only a relatively small number of voice channels. The direction in which optical communications systems develop will depend on the progress that is made in developing cheap, reliable laser sources. The most natural contender appears to be the gallium arsenide injection laser.[14] Its microscopic dimensions (a few hundred micrometres) and its low voltage requirements promise the possibility of a high packing density and the use of integration techniques similar to electronic integrated circuits. Unfortunately, gallium arsenide injection lasers presently suffer from ageing problems. Most diode lasers of this type do not live much longer than tens of hours.[15a] Considerable efforts are being made to increase the lifetime of injection lasers and lifetimes of several thousand hours have recently been reported.[15b,c] Of the many other lasers in existence neodymium doped glasses or crystals or even crystalline neodymium compounds appear promising as possible sources of coherent light.[16, 17]

Even though the optical communication field owes its birth (or perhaps rebirth) to the invention of the laser, a source of coherent light is not absolutely necessary. Relatively low capacity fibres can be envisaged working with incoherent light sources. A close relative of the injection laser, the light emitting diode, l.e.d. for short, provides a cheap, reliable source of a narrow band of incoherent light.[18] Even if the life-time problem of the injection laser, or other lasers, could not be solved (which appears unlikely) optical communications systems using low capacity transmission in multimode fibres offer an interesting alternative to coaxial cable or wire systems. Their small size may alleviate the congestion in underground ducts.

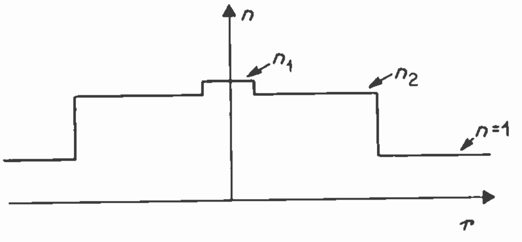

Fig. 2. Refractive index distribution through the cross-section of a clad fibre.

Since this paper is concerned with the transmission medium, the state of the art of optical fibres and their role as light waveguides will be discussed on the following pages.

2 Light Guidance in Fibres

The simplest optical fibre consists of two concentric rods of dielectric materials with different refractive indices. Figure 1 shows the geometry of a fibre with a step index profile. Figure 2 is a plot of the refractive index distribution as a function of distance from the fibre axis. The fibre consists of a small rod of a dielectric material with refractive index n_1 inside a larger hollow cylinder of refractive index n_2. Light guidance occurs if n_1 is larger than n_2.[19] The refractive index of the air outside of the fibre is indicated as $n=1$.

In order to understand the principle of light guidance in an optical fibre we consider a simplified model. Figure 3 shows the cross-section of a dielectric slab whose refractive index profile is also given by Fig. 2. A light ray inside of the core is reflected at the refractive index discontinuity at the core-cladding boundary. The law of reflexion requires that the angle θ is maintained after reflexion and for all subsequent reflexions. If this angle is large enough total internal reflexion occurs and no light escapes from the core. Rays that travel at an angle that is too small for total internal reflexion are only partially reflected at the core-cladding boundary. Their power leaks rapidly into the cladding region so that it cannot be contained in the core and the light may propagate in the cladding. However, this is not desirable since the effective distance of a ray travelling in the cladding is different from that for a ray travelling inside the core. If light travelling in the cladding were allowed to reach the detector, or couple back into the core, severe signal distortion would result. For this reason the fibre may be covered by an absorbing material that prevents light travelling in the cladding from propagating very far. Roughness of the outer cladding surface also serves to scatter light that tries to travel in the cladding.

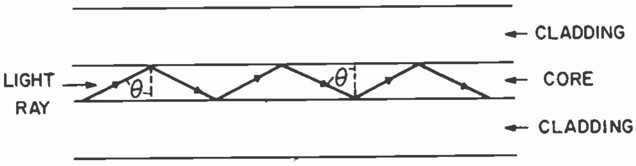

Fig. 3. Ray trajectory in an optical fibre.

Light guidance in the fibre core is restricted not only by the requirement that the angle θ must be larger than the critical angle of total internal reflexion,

$$\theta > \arcsin \frac{n_2}{n_1}. \tag{1}$$

An additional requirement arises from the fact that the

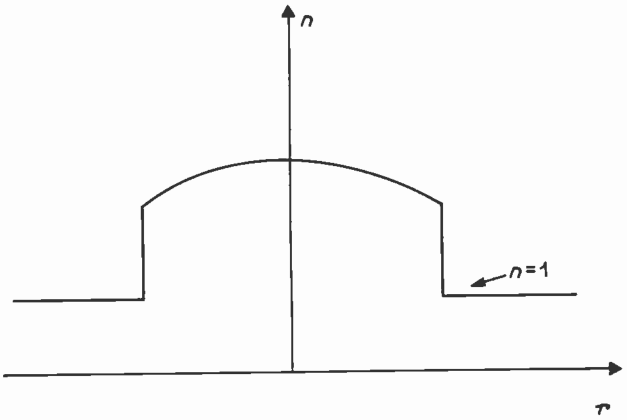

Fig. 4. Refractive index distribution of a graded index fibre.

multitude of rays travelling in the core must superimpose themselves so that constructive interference results. This additional requirement limits the possible angles θ to a discrete set of values. The waves that satisfy the requirements for guided transmission inside of the fibre core are called modes. The optical fibre shown in Fig. 1 can support a finite number of guided modes. The ability to support modes is determined by a characteristic number that is defined as follows[20]

$$V = 2\pi \frac{a}{\lambda} \sqrt{n_1^2 - n_2^2}. \tag{2}$$

The parameter a is the core radius and λ is the vacuum wavelength of the radiation guided by the fibre. The total number of modes that can be supported by the fibre is given approximately by[21]

$$N = \tfrac{1}{2}V^2 \tag{3}$$

These formulae show that more modes can be guided if the core radius is large or if the refractive index difference is large. Equation (3) is not very accurate for small values of N. It is possible to design the fibre so that only one guided mode (of a given polarization) can exist. In this case we have $N = 2$, allowing for both polarizations. From (3) we find $V = 2$ for $N = 2$.

The exact requirement for single mode operation (counting only modes with one polarization) is[19]

$$V < 2{\cdot}405. \tag{4}$$

Thus, single mode fibres can be obtained by either keeping the core radius sufficiently small or by restricting the refractive index difference to small values. If we assume that $n_1 = 1{\cdot}5$ and $n_1/n_2 = 1{\cdot}01$ we need a fibre with a radius smaller than $a = 1{\cdot}82$ µm to ensure single mode operation at a vacuum wavelength of $\lambda = 1$ µm. If we reduce the index ratio to smaller values, for example $n_1/n_2 = 1{\cdot}003$, we can tolerate a larger core, $a = 3{\cdot}3$ µm.

For high capacity systems single-mode operation of the fibre is desirable. To see this, let us consider multimode operation. Each mode travels at a slightly different group velocity. A short pulse, that is shared by many guided modes, thus splits up into a sequence of pulses that arrive at the far end at different times.[22] The difference in arrival time is approximated by the equation[21]

$$\Delta t = \frac{L}{c}(n_1 - n_2). \tag{5}$$

L is the length of the fibre and c the velocity of light in vacuum. For a fibre of 1 km length with $n_1 = 1{\cdot}5$ and $n_1/n_2 = 1{\cdot}01$ we find that the difference in the arrival time of the fastest and the slowest mode amounts to $\Delta t = 50$ ns. With $n_1/n_2 = 1{\cdot}003$ we have $\Delta t = 15$ ns. The values of Δt can be regarded as the width of the detected pulse. For multimode operation the pulse width increases linearly with the fibre length. For a fibre of 1 km length the signalling speed is restricted to 10 Mbit/s for $n_1/n_2 = 1{\cdot}01$ or to 33 Mbit/s for $n_1/n_2 = 1{\cdot}003$. For longer fibres the possible signalling rate would decrease proportionately.

Equation (5) does not apply to single mode fibres. Signal distortion in single-mode fibres is caused by dispersion in the dielectric material of the fibre and by the fact that the group velocity of the mode is somewhat dependent on frequency, even in the absence of material dispersion.[23] However, for small bandwidth these effects are very much smaller than multimode dispersion. It is the bandwidth of the light transmitted through the fibre that determines the signal distortion of single-mode fibres.[22] For light-emitting diodes with a spectral width of approximately 40 nm (400 Å), the width of a narrow pulse spreads to approximately 4 ns/km, for a gallium arsenide injection laser it spreads to approximately 0·1 ns/km and for a neodymium-doped yttrium aluminium garnet laser the pulse width spreads to 0·01 ns/km. Single-mode fibres that are excited with coherent signal sources are thus capable of transmitting signals at the rate of 10^{11} Mbit per second for a guide of 1 km length.

So far we have limited the discussion to fibres with the step index profile shown in Fig. 2. Dielectric optical waveguides can be constructed in different ways. Instead of using the abrupt index distribution of Fig. 2 it is possible to use a graded, continuous index profile.[24, 25, 26] Figure 4 shows the index profile of a typical graded index fibre. This fibre does not have a distinct core and cladding region. Instead, the refractive index changes continuously from its maximum value on-axis to a lower value at the fibre boundary. Outside of the fibre it drops abruptly from the refractive index value of the glass to that of the surrounding air. The trajectory of a ray in a graded fibre is shown in Fig. 5. The index distribution can be approximated by the equation[27]

$$n = n_0\left(1 - \frac{r^2}{a^2}\,\delta\right). \tag{6}$$

The parameter δ determines the rate of change of the refractive index, a is the radius of the fibre.

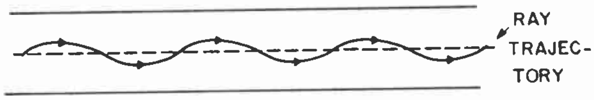

Fig. 5. Ray trajectory in a graded index fibre.

There are several names by which graded index fibres are known. The Nippon Sheet Glass Company uses the trade name Selfoc for self-focusing fibre.[28] The names 'parabolic index fibre' is sometimes used since (6) is the mathematical expression for a parabola. The name GRIN fibre has also been proposed for graded index fibre.[29] Graded index fibres also support a finite number of discrete modes.[19] By making the fibre sufficiently narrow, or letting δ be small enough, a single-mode fibre can be produced. However, the advantage of a multimode parabolic index fibre lies in its ability to transmit signals with less distortion than the simpler multimode step fibre. The pulse width of a very narrow input pulse at the end of a parabolic index fibre of length L is approximately[30]

$$\Delta t = \frac{L}{2c} n_0 \delta^2. \tag{7}$$

For comparison purposes we assume again that $n_0 = 1{\cdot}5$ and use $\delta = 0{\cdot}01$. The parabolic index fibre thus corresponds roughly to the step fibre with $n_1/n_2 = 1{\cdot}01$. However, instead of a pulse width of 50 ns we obtain $\Delta t = 0{\cdot}25$ ns from (7). The signal distortion of the multimode parabolic index fibre is thus considerably less than that of the fibre with the abrupt index profile. From the point of view of signal distortion in multimode fibres the parabolic index fibre is clearly superior.

3 Fibre Losses

It was mentioned in the introduction that fibre losses are the most crucial consideration determining the feasibility of fibres as optical waveguides for long distance transmission of light. There are many different physical processes leading to losses in optical fibres. We will proceed to discuss the most important loss mechanisms.

Absorption losses in glasses are typically caused by the presence of impurity ions. Glass is an insulator with a wide electronic energy band gap making it transparent in the visible region of the spectrum. However, the presence of transition metal ions such as iron, cobalt, chromium, nickel or copper introduces additional electronic energy levels that can absorb light photons. The requirements for the purity of glass from contamination with metal ions are very stringent indeed. For most metal ions[31, 32] we must require that their concentration remains below 10 parts in 10^9. The purity of glasses for optical fibres is thus comparable with the purity requirements of semiconductors for electronic applications. Additional absorption losses are caused by harmonics of the vibration of OH radicals that are associated with the presence of water in the glass.[32, 33, 34] The fundamental frequency of OH vibration corresponds to a vacuum wavelength of 2·8 µm. The third and fourth harmonic of these vibrational transitions are observed at vacuum wavelength of 0·95 and 0·725 µm. These absorption lines are very weak by conventional standards, but they cause significant absorption losses in long fibre waveguides. In fused silica the third harmonic of the OH vibrational absorption at 0·95 µm causes a loss of approximately 1 dB/km per 1 part in 10^6 of OH concentration.[35, 36]

Absorption losses can be measured in bulk samples of a few centimetres length,[8-10] but much more sensitivity and accuracy results by using unclad fibres for such loss measurements.[37, 61] However, even though an advantage is gained by letting the light pass through a considerably longer length of material, the outer surface of the fibre must be meticulously clean to avoid measurement errors that are caused by light scattering from the surface roughness introduced by dust particles, surface flaws and other contamination.

Scattering is another important source of fibre losses. Two types of scattering losses must be considered. Inhomogeneities of the dielectric material cause light scattering. Some inhomogeneities are unavoidable since the molecules in an amorphous material are randomly distributed. Scattering caused by the random distribution of the individual molecules is known as Rayleigh scattering.[38] The Rayleigh scattering loss depends on the inverse fourth power of the light wavelength. Thus it increases very rapidly with decreasing wavelength. Extremely low losses can therefore be obtained only at the long visible wavelength and in the infra-red region. The Rayleigh scattering loss in fused silica at 1 µm wavelength is 0·8 dB/km.[33]

Additional scattering is caused by core-cladding interface roughness.[39, 40, 41] The scattering process depends on the Fourier spectrum of the core-cladding interface deformation function. A given spatial frequency $\Omega = 2\pi/D$ (D is the length of the period of the spatial frequency) causes scattering at an angle[40]

$$\phi = \arccos\left[\frac{\beta - \Omega}{k}\right]. \tag{8}$$

The angle ϕ is measured with respect to the fibre axis, k is the plane wave propagation constant in the medium outside of the fibre and β is the propagation constant of the mode in the fibre. Scattering from random core-cladding interface irregularities can be described in terms of an autocorrelation function of the interface irregularity.[19, 39] The scattering losses are highest for correlation length of the order of magnitude of the light wavelength. The r.m.s. deviation of the fibre core diameter must be kept very small. If the random diameter changes have a correlation length that is roughly equal to the light wavelength, an r.m.s. deviation of the core radius of 0·001 µm causes 10 dB/km radiation loss. Fortunately, real fibres seem to have core-cladding interface irregularities that consist mainly of very low spatial frequencies. The correlation length is correspondingly much longer than the light wavelength and larger r.m.s. deviations can be tolerated. The angular distribution of the scattered power permits the determination of the Fourier spectrum of the core-cladding interface irregularities. E. G. Rawson has measured this distribution and found that forward scattering predominates.[42] This result seems plausible since any inhomogeneity that may be present in the material is stretched out to long length by the fibre pulling process. Consequently, the Fourier spectrum consists of components with long periods or low spatial frequencies. However, even though fibres with low scattering losses have been made, the core-cladding interface tolerance requirements must be considered in the fibre fabrication process.

Finally, a third radiation loss mechanism must be considered. All dielectric waveguides, and consequently also optical fibres, lose power by radiation when they are bent.[19, 43, 44] The fact that bending of the fibre causes radiation losses can be explained qualitatively as follows. We have seen that total internal reflexion keeps the light rays confined in the fibre core. However, the exact treatment of the guided wave problem shows that some power is carried outside the fibre core. The electromagnetic field cannot drop abruptly to zero at the core boundary: a small portion of its energy reaches beyond the fibre core and decays exponentially as a function of distance from the core. This evanescent field tail moves along with the field in the core. When the fibre is bent the field tail on the far side from the centre of curvature of the bend is forced to move faster to keep up with the field in the core. At a certain distance from the core the outside field would be forced to move faster than the velocity of light in the outside medium. The field resists being dragged along at such high speeds by radiating away. The amount of radiation loss depends on the field intensity at the distance where the evanescent field tail would exceed the speed of light and is very strongly dependent on the radius of curvature of the fibre. Below a certain threshold value the curvature losses are negligible, beyond this value they increase so rapidly that practically all the power is lost in a very short distance. This critical radius of curvature is determined by the equation[44]

$$R = \frac{3}{2}\frac{\beta^2}{\Gamma^3}. \qquad (9)$$

The parameter β is the propagation constant of the mode. Its value is bounded by the plane wave propagation constants in the core and cladding materials

$$n_2 k < \beta < n_1 k. \qquad (10)$$

The parameter Γ determines the decay rate of the evanescent field tail and is defined as

$$\Gamma = (\beta^2 - n_2^2 k^2)^{1/2}. \qquad (11)$$

To obtain a numerical estimate of the expected order of magnitude of the critical radius of curvature we set $\beta = n_1 k$ and have

$$R \simeq = \frac{3 n_1^2 \lambda}{4\pi (n_1^2 - n_2^2)^{3/2}}. \qquad (12)$$

For $\lambda = 1$ μm, $n_1 = 1{\cdot}5$ and $n_1/n_2 = 1{\cdot}01$ we find $R = 58$ μm, for $n_1/n_2 = 1{\cdot}003$ we have $R = 345$ μm. The critical radius of curvature is therefore very small for well-guided modes. However, for modes close to their cut-off value (that means close to the point where they are no longer guided in the core) the allowed radius of curvature may become much larger since β approaches $n_2 k$. Curvature losses of fibres must be considered when fibres are twisted around each other or some other supporting structure in a fibre cable. Unlike transmission in metal wires optical fibres cannot be bent arbitrarily abruptly even if their brittleness would allow us to do so.

4 Mode Coupling and Multimode Pulse Propagation

Practical experience has shown that the full pulse width predicted by equation (5) for multimode fibres is usually not realized.[45-48] This more favourable performance of multimode fibres can be explained by two different effects. If we assume that the modes of the fibre remain uncoupled from each other we still have to consider the possibility that their losses may depend on the mode number. In fact most scattering mechanisms, with the exception of Rayleigh scattering, cause high-order modes to suffer more radiation losses than lower-order modes.[49] As a result, the high-order modes with their longer group delay do not contribute as much to power transport so that the delay distortion, predicted by (5) on the basis that all modes carry equal amounts of power, is actually improved.

However, there is a second effect that reduces the pulse distortion of multimode operation. If the guided modes are coupled among each other they exchange power continuously. Some fraction of power, moving at high speed carried by low-order modes, may be shifted to higher-order modes where it continues at lower speed. Conversely, some other fraction of the total power that is initially carried by slow, high order modes shifts over to lower-order modes and continues its journey at higher speeds. Mode coupling causes a mixing of the power among the guided modes. This mode mixing results in an average velocity for the power transport and a corresponding reduction in the pulse broadening. The theory of coupled multimode pulse transmission predicts that the width of the pulse is given by the equation[50, 51, 52]

$$\Delta T = \frac{n}{c}\sqrt{L L_s}. \qquad (13)$$

In the presence of mode coupling the pulse width no longer increases proportionally to the length of the fibre but only as the square root of its length L. The parameter L_s is determined by the coupling strength. However, coupling among the guided modes also causes some scattering of power out of the fibre core resulting in radiation losses. The amount of loss depends on the shape of the Fourier spectrum of the waveguide irregularities that cause the coupling. For a given shape of this spatial Fourier spectrum the expression

$$S = \left[\frac{\Delta T}{\Delta t}\right]^2 \alpha L \qquad (14)$$

is independent of the magnitude of the Fourier amplitudes. The parameter α is the power loss per unit length. The ratio of ΔT (given by (13)) and Δt (given by (5)) is the relative improvement of the pulse width of coupled to uncoupled mode operation. The quantity S determines the loss penalty αL that must be paid for a given improvement in pulse width due to mode coupling. S may typically be 0·1 dB. This would mean that an improvement of $\Delta T/\Delta t = 0{\cdot}1$ would have to be paid with an excess scattering loss of 10 dB. This loss penalty is independent of length and depends only on the amount of relative pulse width reduction.

5 Fibre Pulling Methods

Having described the physical operation of optical fibres and some of their properties we now turn to the question of how optical fibres are made.

Most low loss optical fibres are presently made of glass (The one notable exception will be discussed in the next section.) There are two methods used to pull fibres. The simple rod-in-tube method is illustrated schematically in Fig. 6. A rod made of a glass with the desired refractive index of the fibre core is inserted in a tube made of a glass with the lower refractive index of the cladding. The concentric preform is heated in an electrical furnace or a gas burner to the softening point of the glasses. The fibre is pulled down and fastened to a rotating drum on which it is wound.[7, 34] The fibre diameter resulting from this process depends on the drawing speed and the temperature. The ratio of the outer radius of the tube to that of the rod is maintained in the fibre. The rod and tube method has the disadvantage that imperfections of the inner surface of the tube and the outer surface of the rod tend to cause core-cladding interface irregularities in the fibre that result in scattering losses.

To prevent the interface problem an alternate process is often used. Figure 7 shows the schematic of the double crucible arrangement.[53] Molten glass is held in two concentric crucibles. The inner crucible contains the core material and the cladding material is filled into the outer crucible. The openings of both crucibles are arranged concentric to each other so that the glasses flow out forming a clad fibre. The double crucible method offers the added advantage of allowing wider margins for the core-cladding radius ratios, making it easier to pull single mode fibres with very narrow core and relatively thick cladding.

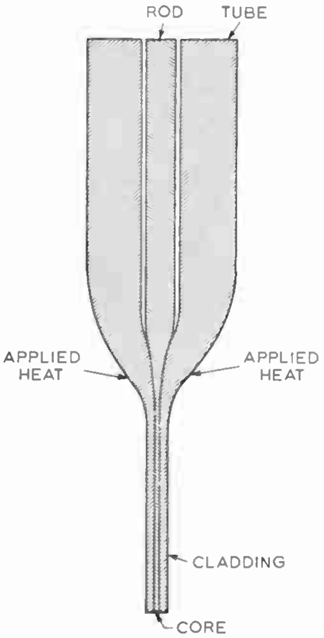

Fig. 6. Rod-in-tube method for fibre pulling.

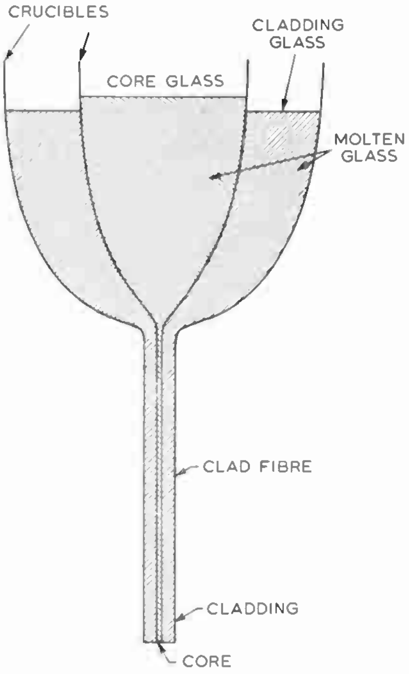

Fig. 7. Double crucible method for fibre pulling.

6 Special Fibres

Dielectric optical waveguides can be realized in many other ways. Of the many possibilities we describe two special types that are of considerable interest.

Liquid core fibres consist of a glass tube that is filled with a liquid of slightly higher refractive index.[54-56] The tube is usually a fused silica capillary that is filled with an organic liquid. Losses as low as 6·5 dB/km at 1·28 μm have been reported for fused silica tubes filled with tetrachloroethylene.[55] This fibre shows a number of absorption peaks that are attributed to overtones and combination tones of the OH vibration. A loss minimum at 1·09 μm reaches the low value of 7·5 dB/km.

Liquid core fibres have the advantage that liquids can be easily purified by repeated distillation. Their principal disadvantage is the large temperature dependence of the refractive index of liquids which limits the temperature range over which a liquid core fibre can be operated. Heating of the core decreases its index and as soon as the core index decreases below the value of the capillary tube, guidance is lost.

A particularly interesting special type of fibre is made entirely out of the same material justifying its name 'single material fibre'.[57] The best optical material known so far is fused silica. It has exceptionally low absorption and scattering losses. Unfortunately, fused silica also has an extremely low index of refraction ($n \simeq 1{\cdot}458$) so that it is very hard to find cladding materials having both low losses and an even lower index of refraction. This dilemma led to the invention of the single material fibre. Figure 8 shows its basic geometry. A slab of dielectric material is enlarged, defining a long, narrow ridge along the slab. Contrary to simple-minded intuition the enlarged region is capable of guiding electromagnetic radiation along the ridge, preventing it from escaping

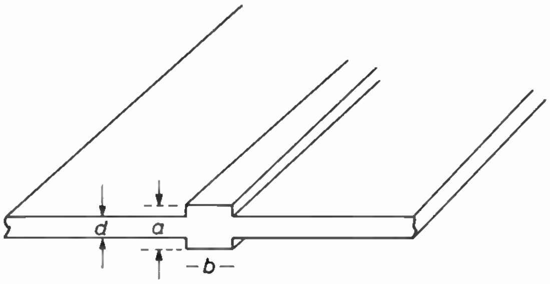

Fig. 8. Schematic diagram of a single material fibre.

into the space outside of the slab or even from spreading in the plane of the slab. A waveguide composed of a single material is thus formed. The precise shape of the enlargement of the slab is not important, any type of bulge serving the desired purpose.

Actual single material fibres are being enclosed in a tube made of the same material. In actual practice the single material fibres were made of fused silica. A slab of this material is inserted into a tube and a narrow rod is placed on the slab in the centre of the tube. This whole arrangement is then pulled into a fibre. The heating necessary for fibre pulling fuses the different parts together. Under proper drawing conditions the geometrical proportions are not changed during the pulling process. Figure 9 shows the geometry of a single material fibre.[57] The outer tube serves only as a shield to prevent contamination of the guiding region by dust or other external influences such as handling, and supporting the fibre. Single-material fibres can be made to support a single mode or many modes depending on the ratio of the enlargement to the slab thickness. Multimode fibres with a guiding rod of 15 µm and a slab thickness of from 3 to 4 µm have been made. A 300 µm section of this fibre had a loss of 28 dB/km at a vacuum wavelength of $\lambda = 1{\cdot}06$ µm. An enlargement of 6·5 µm in a slab of 4 µm thickness results in a single mode fibre. A loss of 55 dB/km has been obtained at 1·06 µm with a single mode fibre.

For multimode single material fibres the number of modes is approximately given by[57]

$$N = \frac{\pi}{2}\frac{ab}{d^2} \tag{15}$$

with a, b and d defined in Fig. 8.

7 Optical Glass Research

Owing to the need to reduce fibre losses to the lowest possible value, considerable efforts at producing low loss optical glasses are being expended in several places.[34, 56–60] Because of its exceptional properties most efforts are directed towards producing low-loss vitreous silica for the fabrication of low-loss optical fibres.

Vitreous silica in its pure form has the lowest losses of any solid or liquid material observed today. Figure 10 shows a spectral loss curve for Suprasil W1 made by Amersil-Heraeus.[61, 62] This material contains only several parts per million of OH so that the OH absorption bands are weak. Losses of approximately 10 dB/km are achieved between 0·7 and 0·9 µm. The loss increases near 0·95 µm because of the third harmonic of the OH vibration and decreases for longer wavelength. The lowest value near 1·1 µm of 6 dB/km does not seem to belong to a minimum so that even lower values are expected at longer wavelength. The curve of Fig. 10 was measured by P. Kaiser using unclad fibres drawn from Suprasil W1. Even lower losses, 3 dB/km at 1·06 µm, have been reported.[33]

It was mentioned earlier that vitreous silica has an exceptionally low refractive index. Two methods have been developed to change the index of the material in order to achieve the necessary core and cladding regions. Corning has patented a process which uses doping of fused silica to increase the refractive index to a value that is suitable for the fibre core. Index ratios of $n_1/n_2 = 1{\cdot}003$ have been achieved by this method.[59]

Addition of boron to the fused silica tends to reduce the refractive index slightly below the already low value of pure silicon dioxide.[63] This borosilicate glass is suitable as a cladding for a pure fused silica core. Fibres

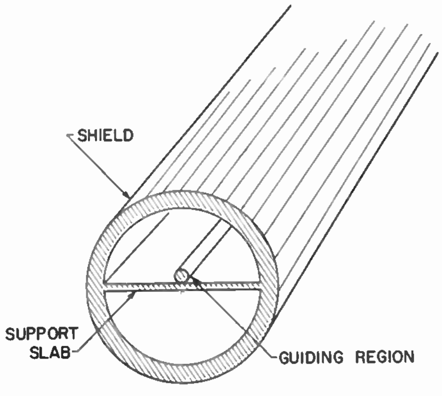

Fig. 9. Single material fibre with protective shield.

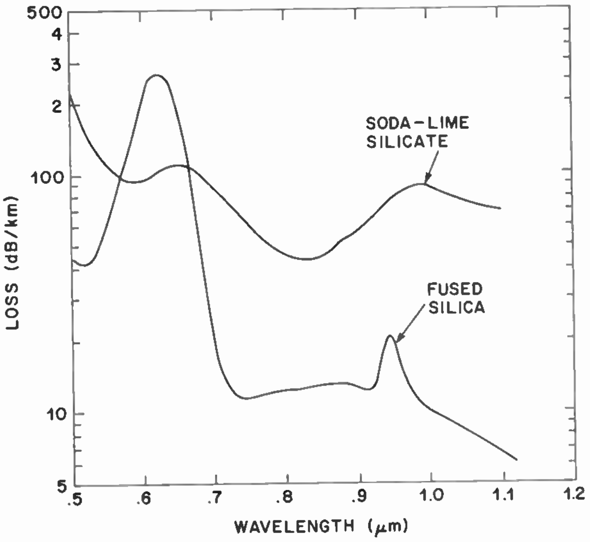

Fig. 10. Loss curves for soda-lime silicate glass and fused silica.

with losses of 10 dB/km near 0·7 μm have been achieved so far.[64]

The lowest fibre loss achieved to date comes from the Corning group reporting a multimode fibre with 4 dB/km loss at $\lambda = 1{\cdot}06$ and 0·8 μm.[13]

A disadvantage of fused silica fibres is the very high softening temperature of this material. For ease of fibre pulling it would be advantageous to have a material with lower melting point. For this reason efforts are being made to reduce the losses of soda-lime silicate glasses. By changing the composition of such glasses the refractive index can be readily varied so that core and cladding can be produced from the same material with slightly different compositions. So far the losses of soda-lime-silicate glasses are not yet as low as those of fused silica. Figure 10 shows a spectral curve of soda-lime silicate glass losses.[61] Approximately 45 dB/km have been obtained near 0·7 and 0·8 μm. There seems in principle to be no reason why the losses of these glasses cannot be reduced further. The main source of loss appears to be contamination with transition metal ions. It is hoped that further purification of the glass will result in lower loss.

8 System Considerations

Since no optical communications system does yet exist we can only speculate at the shape such a system may assume. We have concentrated on the transmission medium and outlined the state of the art of fibre waveguides. The answer to the question as to the most suitable source of light must await the outcome of efforts to improve the gallium arsenide injection laser. Heterojunction lasers consisting of sandwiches of aluminium-gallium arsenide on either side of a gallium arsenide layer have proved successful in achieving continuous laser operation at 0·9 μm wavelength at room temperature.[14] Unfortunately, these lasers age quickly and cease to oscillate after tens of hours.[15] Because of its small size the gallium arsenide injection laser would be an ideal source of coherent light for communications purposes. The incentive is strong to overcome the lifetime problem. Other possible contenders are neodymium glass lasers or so-called neodymium YAG (yttrium-aluminium garnet) lasers that need an optical light source as a pump.[16, 17] It is possible to pump these lasers with luminescent diodes. The laser emits at 1·06 μm.

For high signalling rates it would be necessary to use external light modulators. For lower rates of the order of up to 1000 Mbit/s it is possible to modulate an injection laser simply by modulating its supply current.

If lasers are used as signal sources relatively high signalling rates can be achieved. The rate is limited by material dispersion in the glass of single mode fibres. This limit might eventually allow signalling rates of 10^{11} bits per second for a fibre of 1 km length.[22]

For more modest signalling rates light emitting diodes are available as attractive light sources.[18] L.e.d.s do not suffer from the life time problems that plague the injection laser. However, they are incoherent sources and it is impossible to excite a single mode fibre efficiently with an incoherent light source. We are thus forced to use multimode fibres to transmit the light of an l.e.d. The signalling rate is limited by the pulse dispersion in multimode step fibres to roughly 100 Mbit/s. L.e.d.s can be modulated by modulation of the drive current at rates exceeding the signalling rate permissible by this transmission medium. Light modulators are thus not necessary in this case.

Light detectors do not pose any particular problem in the visible and near infra-red region of the spectrum. Reverse biased p-n junctions draw a current when light is absorbed in the junction region. By increasing the bias voltage near the avalanche breakdown point it is possible to achieve some gain in the detector since each absorbed photon causes an avalanche of hole–electron pairs. Avalanche gain is desirable to overcome the thermal noise in the low frequency circuits and keep the detection process shot noise limited.

9 Conclusions

Communications systems based on light transmission through optical fibres promise to become competitive with other methods of long distance communications. The components of a fibre communications system appear to be inherently cheap. Potentially, there is an enormous bandwidth available since light has a frequency of 3×10^{14} Hz (at $\lambda = 1$ μm). However, at present there is no prospect of utilizing this enormous bandwidth. In fact, a communications system using light emitting diodes and multimode step index fibres would not exceed a signalling speed of the order of 100 Mbits per second. However, even at this low rate an optical fibre system may be competitive with coaxial cable or even waveguide systems since an individual optical fibre takes up very little space so that a very large number of fibres can be incorporated in a fibre cable. In the distant future it may be possible to utilize the wide bandwidth potential of the light carrier more fully. In this case equalization methods may be used to compensate fibre dispersion or other types of transmission media may become necessary.

At present, optical fibre communications systems are still in the research stage. Problems of splicing fibres and constructing fibre cables need to be solved. The shape that a practical fibre system will assume is still not determined. Perhaps several of the potential possibilities, such as single or multimode transmission, may be used for different applications.

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The challenge of fibre-optical communication systems

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SUMMARY

The paper reviews the demands made on technology by the potential impact of fibre-optical communication systems. The reasons for the interest in this technology are discussed in terms of discounted cash flow for wideband transmission installation. The primary system requirements are identified and research aimed at developing fibre-optic cable, light sources and terminal sub-systems is described.

Some typical system parameters which can reasonably be related to the present state of art are given. It is concluded that there is now little doubt that fibre optic communication will make a major impact on transmission technology. It is further suggested that the greatest impact may be in opening up new fields of telecommunication.

1 Introduction

Fibre-optical communications represents a potential leap-frog in the state-of-the-art in transmission systems by bringing together new system concepts and new technology. As such, it provides a substantial research and development challenge because it draws on a very wide range of skills and facilities to advance the technology to the level where the full competitive potential of the new system can be realized. Fortunately, the incentive for advancing many of the technologies required for fibre-optical communications stems from a much broader business base. Thus the development of glass as an electronic material, the advances in III–V semiconductor technology, and the evolution of digital communications and techniques, have independently contributed to the present situation where system planners are considering fibre-optical communications as one of the principal wideband transmission systems of the future. This paper reviews the challenge in the context of the work being carried out at STL. More general surveys have been published elsewhere[15–18].

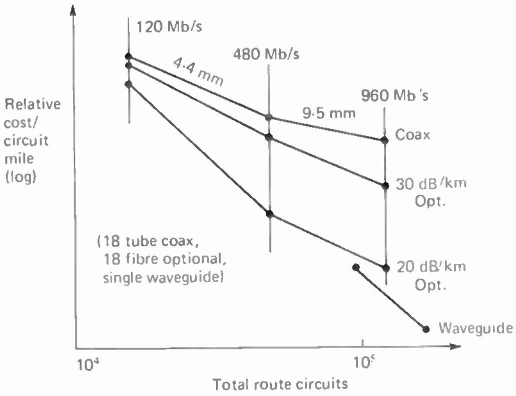

Fig. 1. Relative circuit costs.

In order to understand why the initial analysis of dielectric-fibre waveguides[1] aroused the interest of transmission engineers, one must briefly consider the economics of wideband communication. Figure 1 compares the relative cost per circuit-mile of three types of digital transmission system. Although they all show a cost reduction as the capacity is increased, the coaxial cable system shows a pronounced decrease in the cost reduction slope at the higher bit-rates. This is because at the highest speeds the transmission medium either becomes more expensive, as indicated by the change in the diameter of the cable from 4·4 mm to 9·5 mm, or the repeater spacing has to be reduced. On the other hand, the optical system shows a much more pronounced downward trend of costs between 480 and 960 Mbits/s. In this case there is still a change in slope due to the higher repeater costs associated with the greater bit rates: however, the diagram illustrates the cost advantage of using a transmission medium where the actual line costs are virtually independent of the bandwidth at relatively high bit rates. This is also very strongly indicated for the long-haul millimetre waveguide.

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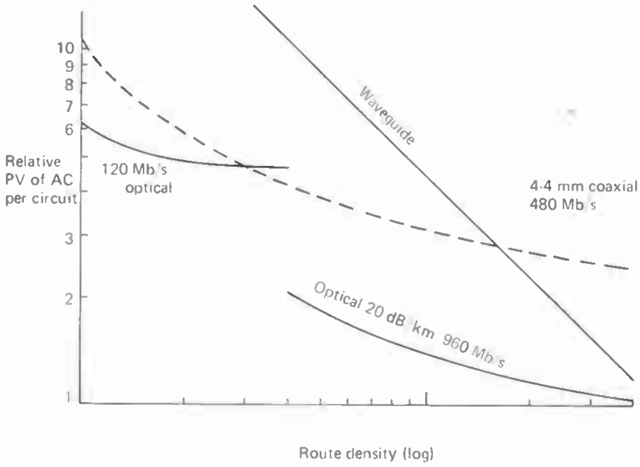


Fig. 2. Discounted installation costs.

It is necessary to consider discounted cash flow to understand the relative merits of millimetre waveguide and optical waveguide wideband communication. Figure 2 compares the relative present value of annual charges (PV of AC) of the systems discussed above. This notation enables one to compare the true cost of providing equipment where part of the installation has to be made well in advance of the projected growth in traffic.

For example, if one assumes a 10% per annum growth in traffic demand and an effective capitalization period of 20 years, then in the case of the millimetre waveguide much of the costly installation would have to be carried out at the beginning. The optical-fibre system on the other hand, despite its high bandwidth capability, has a similar installation and line cost pattern to the coaxial cable. Thus it does not have the steep rise in PV of AC shown by the millimetre waveguide at the lower route densities. The values in Fig. 2 are relative, but an impression of absolute values can be obtained by noting that the route density at which the plot for the waveguide becomes competitive corresponds to the projected future traffic between London and one of the major cities. Thus Fig. 2 shows that there may be a few routes in a country like the United Kingdom where the millimetre waveguide appears economically attractive but because of the greater installation flexibility of the optical-fibre communication system the potential impact is much greater. Indeed the interest in this new technology extends well down to the lower capacity trunk applications and beyond to local area transmission. Apart from the civil telecommunications requirements, the technique is now being considered for many short link applications where the light weight, small size and immunity to interference are of interest.

2 The Basic System Requirements

The most common types of fibre are summarized in Fig. 3. It is convenient to consider the various modes of optical propagation in terms of the simple geometric ray picture. Although only two rays are shown in the multimode fibre, in practice for the example chosen, over 1000 modes could propagate at 800 nm wavelength. In the case of the single-mode fibre, where the core diameter is comparable to that of the wavelength, most of the energy would be in the HE_{11} mode. For the Selfoc (self-focusing) fibre the behaviour can be either multimode or single mode, depending on the choice of the graded refractive index profile. Figure 4 shows the actual mode patterns for the first few orders. The patterns are obtained by observing the radiation from the end of a 'single mode' fibre and reducing the wavelength of the launched light from the cut-off condition. An image intensifier is used to enable the photograph to be taken. For a multimode fibre of the type shown in Fig. 3, θ would be about 1·5° for the highest order group shown in Fig. 4, i.e. the group HE_{22} etc.

The key parameter of these fibres, namely the attenuation, is in all cases primarily determined by absorption in the core material and Rayleigh scattering by inhomogeneities. To a first order, this is the same for multimode or single-mode fibres. The single mode has more of the energy travelling in the cladding and hence the attenuation is influenced to a much greater extent by the cladding material than in the case of the multimode. However, the main difference between the two types of fibre is in the dispersion, or, as far as the digital system is concerned, pulse spreading, which is caused by the different effective velocities of the different modes. This is indicated by the (rather exaggerated) difference in the path lengths of the two rays shown at the top of the diagram. The dispersive effect of the longer path of the outer curved ray in the Selfoc fibre is partially compensated by the fact that it moves through a region of lower refractive index and hence higher velocity. Clearly most system requirements

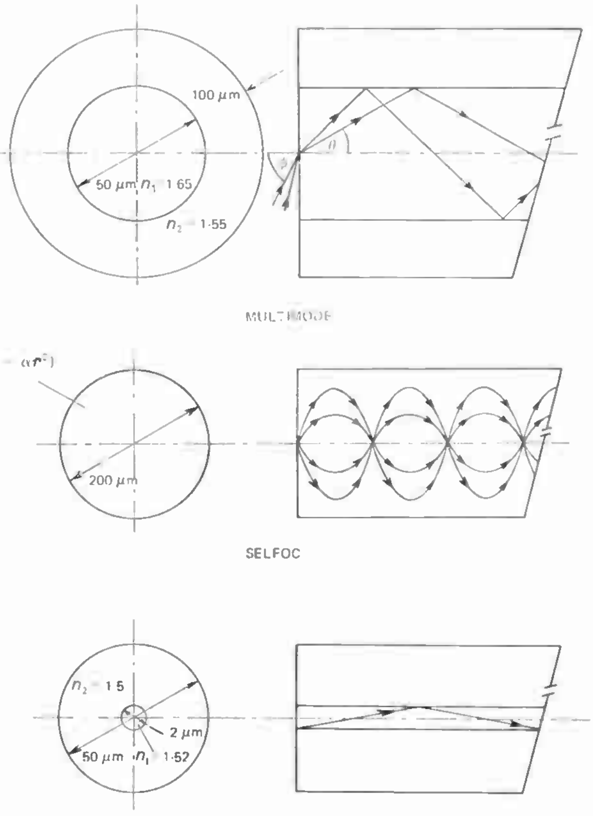
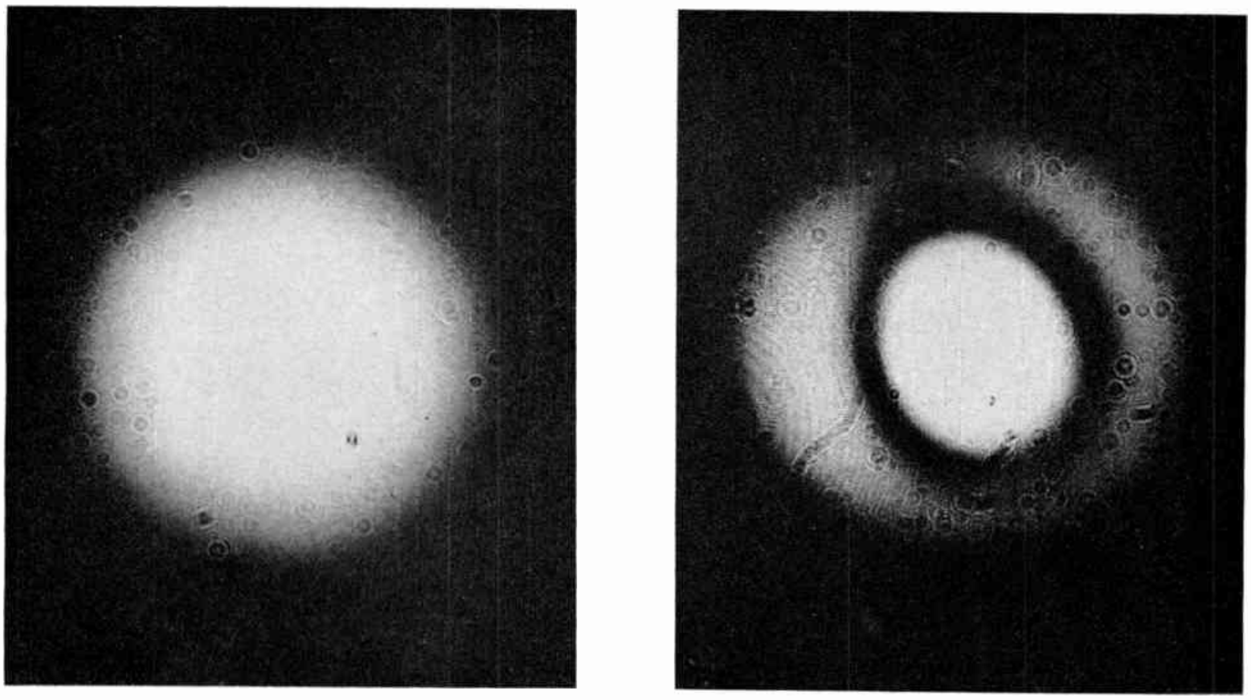
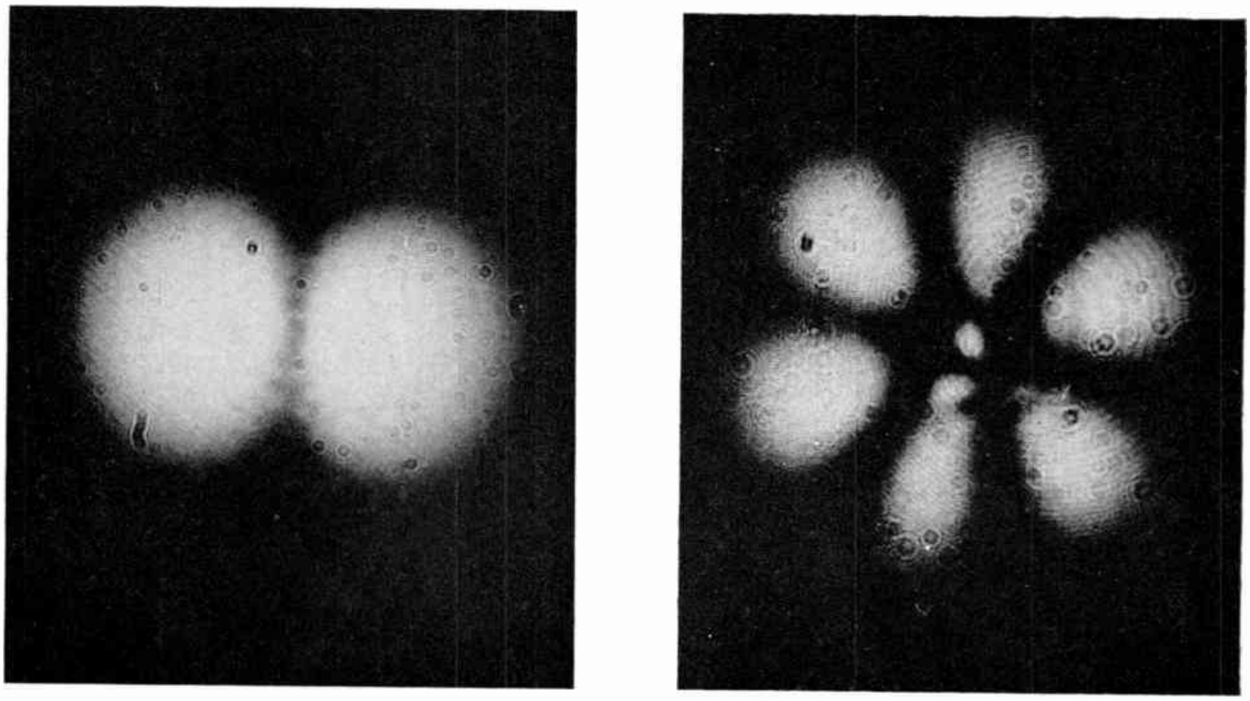


Fig. 3. Typical fibre structures.



HE_{11} HE_{12}



HE_{21}, TE_{01} and TM_{01} HE_{22}, TE_{02}, TM_{02} etc.

Fig. 4. Mode patterns.

will demand the lowest possible attenuation which may ultimately be limited by Rayleigh scattering at around 1 to 2 dB per kilometre. From the system point of view, the other major difference between multimode and single-mode fibres is in their 'impedance match' to the light source.

Figure 5 is an attempt to relate the various combinations of sub-system which can be used to form typical fibre-optical communications. A trunk communication system would use the arrangement shown at the top which has a semiconductor laser coupled to a single-mode fibre. In the context of this system, the word 'laser' implies not so much a coherent source but a source with a high luminous intensity over an area comparable with the core of the single-mode fibre. Also, since in this system pulse spreading would be due to material dispersion only,

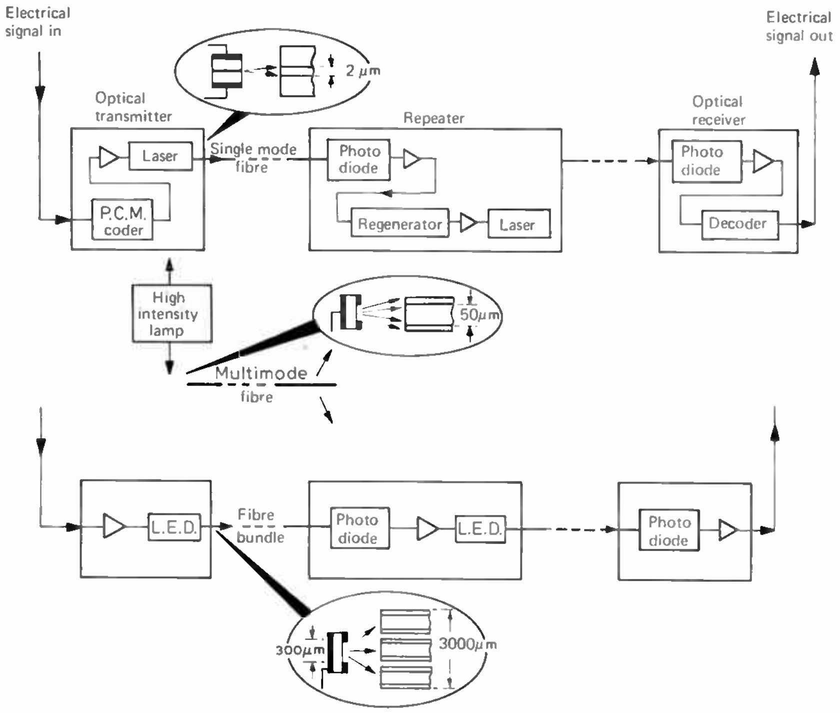


Fig. 5. Basic fibre-optical communication systems.

'laser' implies a reasonably monochromatic source.

At lower bit rates the laser might be replaced by a high-intensity semiconductor lamp using laser technology, but radiating from the direction perpendicular to the junction from an area which is comparable to the diameter of the core of the multimode fibre. At the other extreme is the light-emitting diode which has a relatively low intensity broad source. Unlike the laser, which has a very non-linear response above the lasing threshold, the light-emitting diode can be amplitude-modulated. Its area is a reasonable impedance match to a bundle of multimode fibres which are suitable for short distance applications. The high-intensity lamp and single-strand multimode fibre can also be used in the analogue system shown at the bottom of Fig. 5.

3 The Transmission Medium

By a systematic approach to the purification of glass, in a way similar to that used for semiconductor crystals, the relation between the concentration of transition elements which cause absorption and attenuation in soda-lime silica glasses has been identified. In many respects the purification process is more difficult than for semiconductors because the concentration required for the impurity is of the same order, but zone refining cannot be used with an amorphous material. However, by using processes, and an approach to cleanliness, more common to semiconductor operations than traditional glass working, bulk glass attenuation approaching the values indicated in Fig. 1 have been obtained.

Figure 6 shows a glass melting apparatus in which the problem of crucible contamination is avoided by coupling the r.f. power directly into the glass in a 'cold' silica crucible.[4] The powdered components are first heated by using the graphite susceptor beneath the crucible. At a temperature of approximately 1000°C a crude soda-lime glass is formed which then couples directly to the r.f. field. At this point the temperature is maintained at above 1000°C by the r.f. coupling and the susceptor is removed. Thus during the (several hours) period of homogenization, the glass is the hottest part of the system. After homogenization the glass is pulled from the crucible in the form of rods. Bulk losses

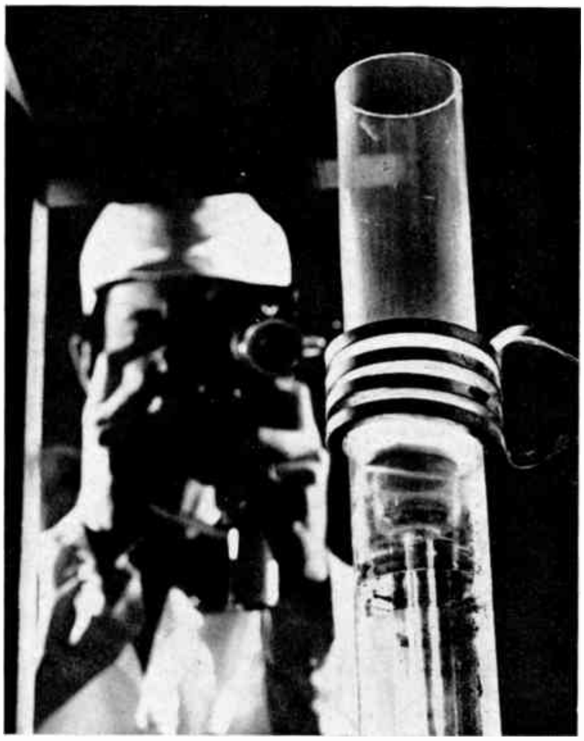


Fig. 6. R.f. melting of glass.

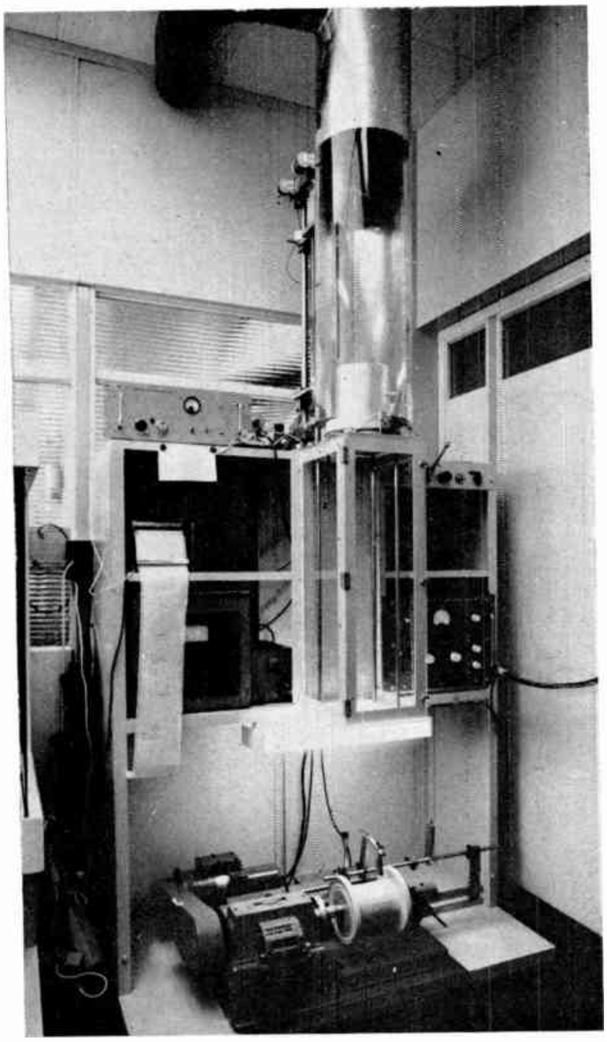


Fig. 7. Fibre-pulling apparatus.

below 30 dB per kilometre at 1060 nm have been obtained by this technique.

To preserve the quality of the glass after it has been formed into a fibre it is desirable to minimize its handling and processing after melting. An important step towards this at STL has been the co-location of the clean room for glass melting with the clean room housing the fibre-pulling apparatus, shown in Fig. 7. This type of apparatus is capable of producing fibre by either the 'composite rod' or the 'double-bushing' techniques indicated in Fig. 8. The advantage of the double-bushing process is that it can take glass directly from the melting process, and in principle can be operated as a continuous mass production process. The composite rod process avoids the possibility of crucible contamination, but may need some intermediate processing unless the composite is formed as part of the glass preparation process.

Figure 9 gives attenuation as a function of wavelength for a fibre produced by the double-bushing process using r.f.-melted soda-lime silica glass with a bulk attenuation of 45 dB per kilometre at 900 nm. Figure 10 gives the results for a fibre produced by the composite-rod process from a cladded glass rod, again using the r.f. melting technique used for producing the glass. In this case the core attenuation of the bulk material was 57 dB per kilometre. Although in both cases additional loss is introduced during the fibre fabrication process, these results indicate that considerable progress has been made in minimizing the contamination of glass en-route between raw material and finished fibre. These, and

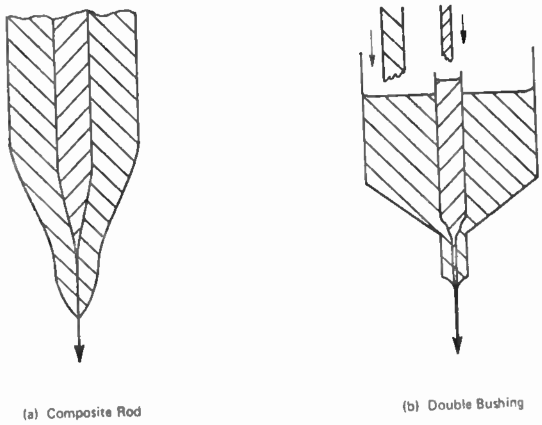


Fig. 8. Fibre-pulling methods.

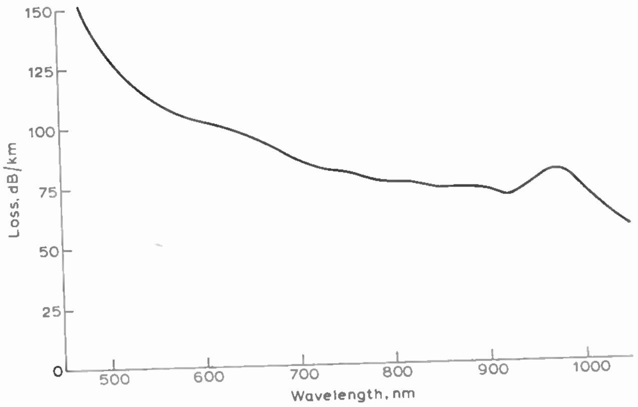


Fig. 9. Total fibre loss for 40 µm core in 70 µm cladding produced by double bushing.

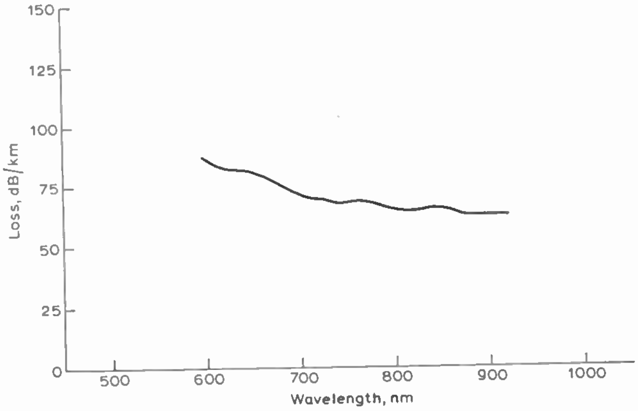


Fig. 10. Total fibre loss for 30 µm core in 65 µm cladding produced by composite rod.

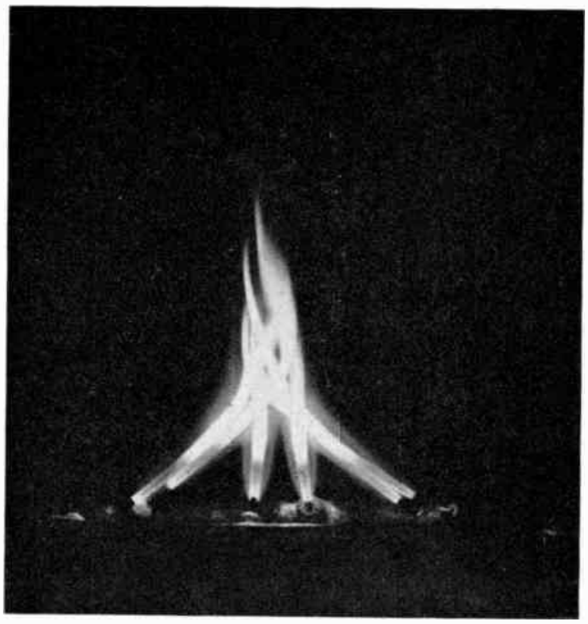


Fig. 11. Luminous propane flames to demonstrate configurations for pulling fibre. 6-jet swirl-flame burner.

similar experiments, show fairly consistent correlation between the final fibre attenuation and the attenuation of the bulk glass before fibre pulling. Thus, further refinement of both processes, to lower the bulk glass loss and to reduce the gap between bulk and fibre attenuation is likely to lead to fibres meeting the requirements for optical communication systems.

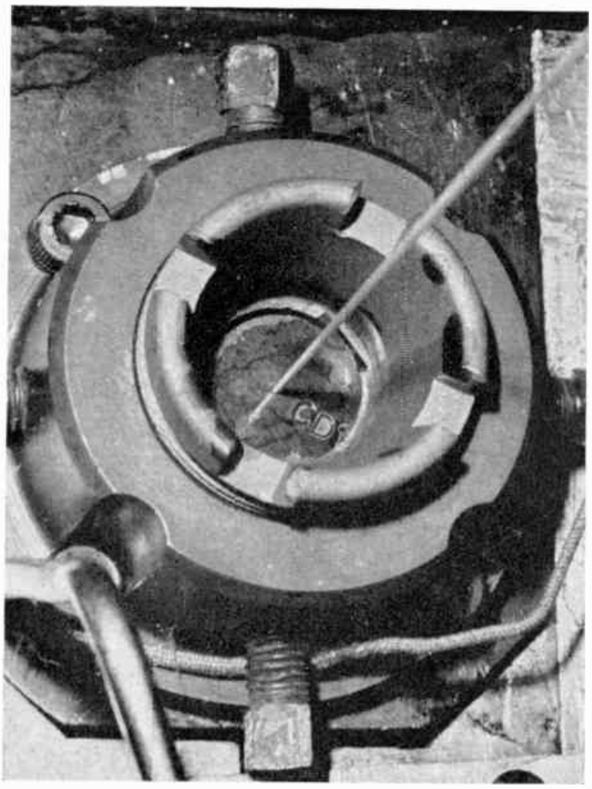


Fig. 12. Extrusion head for plastic-coated fibre.

The material which has consistently shown the lowest attenuation is pure silica.[5] Fibres using a doped silica core giving a loss less than that obtained by any other materials have also been reported.[6] The refractory nature of silica, which probably accounts for its resistance to contamination, also makes it more difficult to fabricate into fibre. Compared with refined r.f. melting techniques, described for homogenizing glass, silica composite rod has to be heated for fibre pulling by the relatively unsanitary process of an oxy-hydrogen flame, or by means of a very high power gas laser. A six-jet swirl flame used for silica fibre pulling is shown in Fig. 11. (In this case propane gas has been used for the purpose of the photograph.)

The results obtained with silica certainly encourage one to pursue this technological route in addition to the purification of the glasses discussed above.

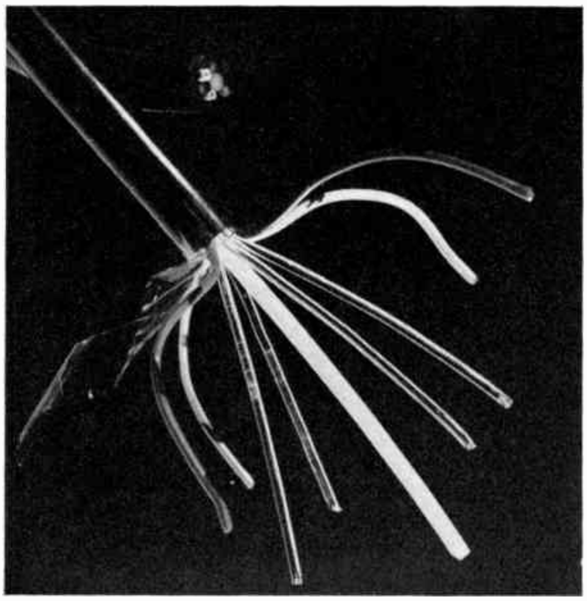


Fig. 13. Cable showing optical and power leads.

The final stage in the realization of the optical transmission medium is the fabrication of the cable. Figure 12 shows a cable extrusion head which has been adapted to take a single-strand multimode fibre and coat this with a thermoplastic polyester. This improves the mechanical handling properties of the fibre and increases the tensile strength to several kilogram-weight for a typical 100 μm fibre encased in 1 mm (diameter) of plastic. Figure 13 shows a cable comprising four of these plastic-covered fibres with some power leads. When one considers the bandwidth capability of each of these strands it places space-division multiplex in a more favourable light compared with time-division multiplex for wideband systems.

4 Light Sources

The multi-layer heterojunction technology[7] has provided most of the design flexibility needed for fabricating the high-intensity sources discussed in Section 2, since it has enabled the double-heterojunction laser[8] to be

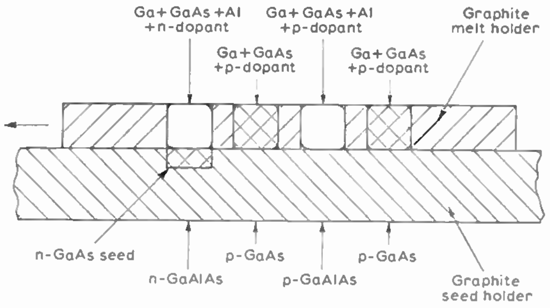


Fig. 14. Sliding boat epitaxy.

realised. By using a sliding boat multi-layer epitaxy technique (see Fig. 14), the refractive index of the crystal on either side of the central GaAs region is modified by alloying aluminium. The optical guide, created by the interface between the p-type GaAs and the p and n-type GaAlAs, establishes a photon-electron confinement which enables sufficiently low threshold current densities for efficient c.w. operation to be obtained. The main characteristics of this device are shown in Fig. 15(a). The scanning electron micrograph of one of these devices is shown in Fig. 16, on which the three layers have been indicated.

The sliding-boat technology can be extended to give more complex carrier-refractive-index profiles. Figure 17 shows a photograph of a sliding boat with provision for 11 layers. This has been used to produce the localized gain region (l.g.r.) 5-layer lasers illustrated in Fig. 15(b). These lasers have enabled a further substantial reduction in threshold current density to be achieved.[9] Further reduction in the thickness of the GaAs layer, shown in Fig. 15(a), does not increase the confinement of the photons or reduce threshold current density. The l.g.r. structure shown in Fig. 15(b) uses the two additional GaAlAs layers to provide a waveguide structure for optimum optical confinement, independent of the much narrower GaAs region which confines the carriers.

The characteristics of three different device configurations, using the double-heterojunction technology, are summarized in Fig. 18. The broad contact device is suitable for relatively low bit-rate multimode applications. The mesa laser, with its low threshold current and small area, is specifically designed for high bit-rate single-mode applications. The high-intensity Burrus diode[10] is suitable for multimode digital or analogue systems (provided that the linearity is acceptable).

Although such problems as the repair of damaged optical cables in the field may provide challenges for the future, undoubtedly the major technical challenge in optical communication research today is laser life: serious degradation occurs, with c.w. equivalent operation of less than 1000 hours for most of the lasers produced. The degradation mechanisms are not yet satisfactorily identified. On the other hand, one can take some comfort from the fact that the degradation process can be fairly consistently characterized: it increases with the threshold current density, and is more or less directly proportional to 'on' time, i.e. halving the duty cycle doubles the operational life. There is no significant degradation during shelf-life test. Certainly, with great attention to manufacturing detail, such as protection of the resonator mirrors and reduction in strain, the life of double-heterojunction lasers does seem to be improving. Further reductions in threshold current density by methods such as the l.g.r., should also help to increase laser life. However, there is no doubt that a breakthrough in understanding the degradation mechanism is needed. One must hope that this understanding will not reveal any fundamental degradation process, but will point the way to soundly based improvements.

5 Subsystems Considerations

Assuming the required progress in the technologies described, one can design fibre-optical communications systems along the lines indicated in Table 1, which summarizes the main parameters of three types of system.

The calculations have been based on p.c.m. with an error rate of 1 in 10^9. Avalanche detector sensitivity is assumed, as in a pulse system the higher noise of this device does not weigh significantly against the much

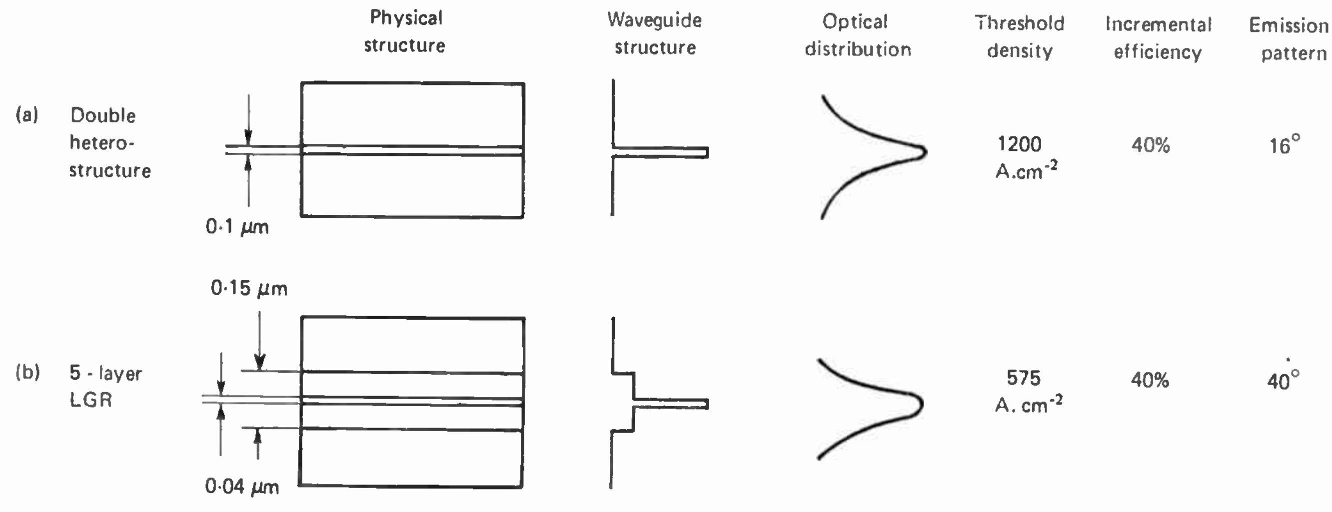


Fig. 15. Double-heterostructure representation.

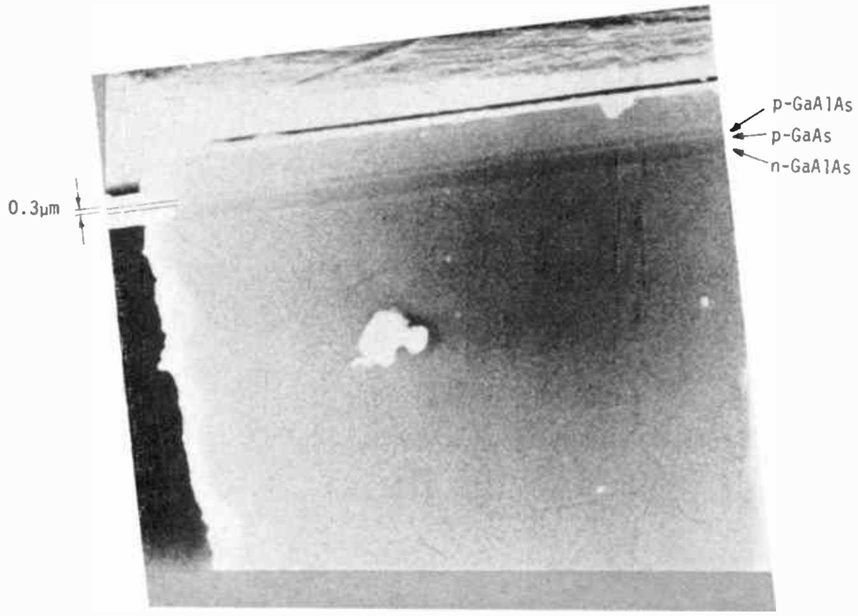


Fig. 16.
Scanning electron micrograph of double-heterostructure laser.

higher gain. This does not apply in an analogue modulated system where the noise of an avalanche detector may not be acceptable.

Although very low loss glass or silica fibres are not yet readily available, useful design information has been obtained from experiments with low-loss liquid-filled fibres.[11] Results indicate that in multimode fibres the dispersion over realistically useful lengths is determined largely by the behaviour of the modes initially launched. Thus if the launching of higher-order modes is limited by restricting the divergence of the input light, i.e. θ and ϕ in Fig. 3, then there appears to be little conversion to higher-order modes (which would of course cause greater dispersion). The values of dispersion given in the final column of Table 1 have been calculated on the purely geometrical considerations indicated in Fig. 3. In practice these give, somewhat surprisingly, a conservative rather than an optimistic result. Some of this 'conservatism' may be due to attenuation of the higher-order modes, i.e. by scattering into the cladding. However, attenuation measurements show that scattering

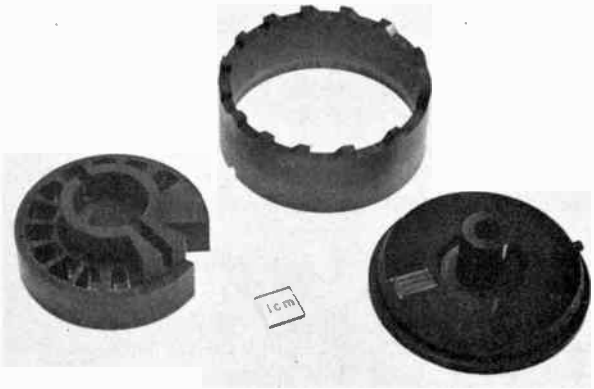
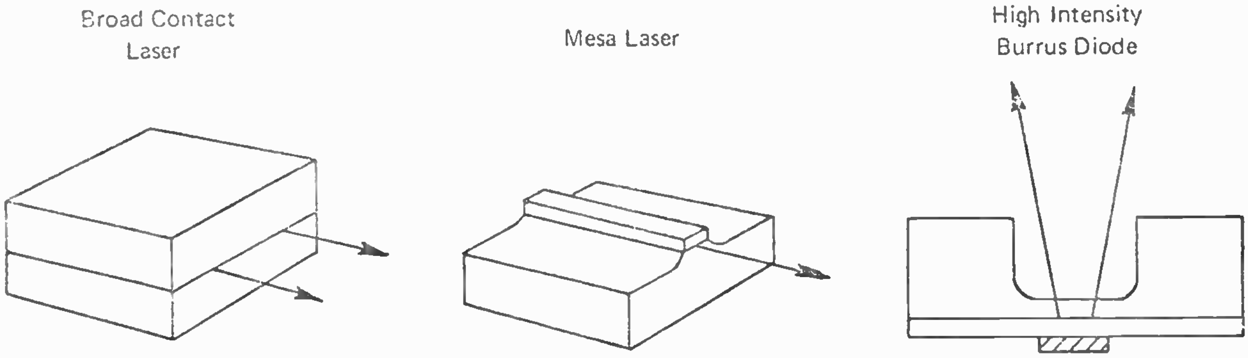


Fig. 17. Circular sliding boat using graphite.



Source size	1·0 µm × 75 µm	1·0 µm × 15 µm	50 µm diameter
Threshold current	0·5 A	0·1 A	—
Drive current	~1·0 A	~0·2 A	~0·2 A
Output	100 mW peak	10 mW mean	1·0 mW mean
Duty factor	10%	up to c.w.	up to c.w.
Modulation rate	~10 M bit/s	⩽ 1·0 G bit/s	100–200 M bit/s

Fig. 18. Laser and lamp characteristics.

Table 1 Typical optical system parameters

System bit rate Mbit/s	Fibre loss dB/km	Repeater spacing km	Duty cycle %	Power received μW	Power launched mW	Line loss dB	Launching angle ϕ°	Propagation angle θ°	Dispersion ns/km
35	8	6·28	4	0·01	0·9	50	3·4	2·3	4·0
	16	3·22			1·3	52	4·8	3·2	7·8
120	8	5·9	14	0·03	1·8	47	1·9	1·3	1·2
	16	3·1			2·5	49	2·7	1·8	2·4 (max. for 480 Mbit/s)
480	4	8·2	25	0·2	0·4	33	—	(HE_{11})	(0·21)
	8	4·1							(0·42)
	20	1·7							(1·1)

is not a prime mechanism; more plausible is the selective conversion to the lower-order modes.[12] This suggests that for longer distances the linear relationship between pulse spreading and distance leads to somewhat conservative figures for the bandwidth capability of multimode fibres.

The 35 and 120 Mbit/s systems have been considered in terms of multimode fibres similar to those shown at the top of Fig. 3, and the 480 Mbit/s system has been considered in terms of single-mode fibre, as shown at the bottom of Fig. 3. Comparison of the 8 and 16 dB parameters shows that repeater spacing is not proportional to fibre loss in the multimode case, because at the lower attenuation it is necessary to limit the launching of higher order modes by reducing ϕ. This means that a lower line loss must be specified to keep the received power the same. Similar considerations apply at 120 Mbit/s; in this case the permitted line loss is rather less than the 35 Mbit/s because the received power at the detector diode may have to be greater at the higher bit rate from signal/noise considerations.

The calculations have been carried out for lasers. If a diode were used rather less power could be launched (see Fig. 18). This will again be reflected in the line loss column of Table 1. Referring to the emission pattern column of Fig. 15, it can be seen that the double-heterostructure laser is in fact a better match to ϕ than the l.g.r. device despite its much lower threshold. This suggests that for the multimode systems the double-heterostructure would be better, but for the single-mode fibre the 40° emission pattern of the l.g.r. five-layer device would not be a serious disadvantage compared to the advantage of the low threshold current.

The significance of the duty cycle column of Table 1 is that in the 35 Mbit/s case the 4% duty cycle represents a considerable operational advantage from the standpoint of laser life. Some life advantage should also be obtained in the 480 Mbit/s case, also due to the 25% duty cycle. However, in this case the situation is less straightforward. Figure 19 shows the component parts of the transmitter the design of which enables mesa lasers to be modulated at 1 Gbit/s, by direct modulation of the laser rather than by using electro-optical modulators. This has been achieved by biasing the laser below threshold and applying a differential modulation to take it in and out of the lasing condition. Similar technique would also be used at about 500 Mbit/s. If the degradation processes are associated with the actual lasing condition, rather than purely the current density in the 'on' condition, then laser life should be extended by operating at an effective 25% duty cycle. However, from a power dissipation point of view the laser is effectively in a c.w. condition at these higher bit rates. Present evidence, from a comparison of the behaviour of high-intensity lamps and lasers operating at similar current densities, indicates that it is the lasing condition which accelerates degradation. Hence it is reasonable to assume that even at the 480 Mbit/s rate the laser life is greater than in the c.w. condition.

In the single-mode case the launching efficiency will generally be less than in the multimode case because even with mesa lasers there is still a significant mismatch between the active area of the laser and the core diameter. In the example chosen, about 20% efficiency has been taken. This, and the higher required received power at the higher bit rate, is again reflected in the line loss and hence

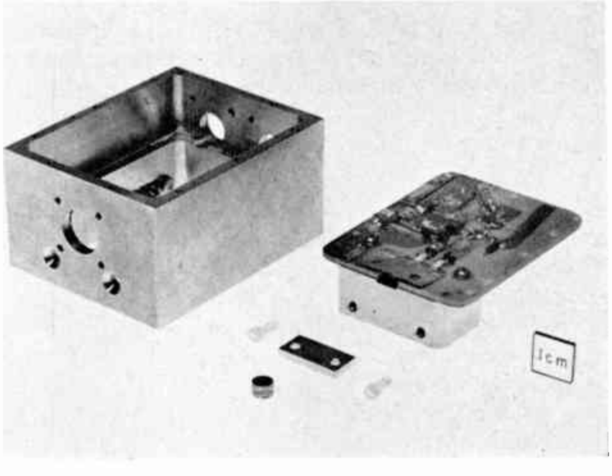


Fig. 19. Component parts of optical transmitter.

Table 2

Material dispersion data

Material	Laser $\delta\lambda = 20$ Å	Lamp $\delta\lambda = 200$ Å
Glass (soda lime silica)	0·4 ns/km	2 ns/km
Silica	0·18 ns/km	0·9 ns/km

the repeater spacing. These factors tend to erode some of the advantages of the cost/bandwidth relationship discussed earlier, and stimulate attention to improved launching methods, more sensitive detectors, and lower joint losses in the system.

Reference to Table 2 shows that, even in the single-mode case, dispersion must be considered at the higher bit rates. The Table shows that even if lamps could be modulated at the higher bit rate these devices would not be suitable because of their spectral line width. The laser is satisfactory until glass of better than 8 dB per kilometre is obtained. Thus, as lower loss glass and silica fibres become available, and higher digital modulation rates are used, attention to reducing the spectral spread of the laser will be worthwhile.

Finally, a comment on integrated optical subsystems is appropriate. Whereas proposals for repeaters giving gain in the optical region are still very speculative, local area applications may accelerate the requirements for optical switches and junction devices. Work on thin-film electro-optics technology[14] is likely to produce devices which can provide useful optical switching. If fibre-optical systems are used for communication between terminals sufficiently close for no electronic gain to be required to make up the fibre losses, then the incentive for remaining in the optical regime to carry out simple switching becomes much greater.

6 Conclusions

By coordinated research on materials, components and subsystems on a fairly broad front, the status of fibre-optical communication has been transformed from an 'if' situation to a 'when' situation. This paper has tried to indicate the types of system which can today be envisaged with a reasonable degree of confidence, and to outline the demands which these will make on the technology.

In considering the potential impact of fibre-optical communications there is a natural tendency to make comparisons with established systems and techniques. Although some of these comparisons involve considerable extrapolations, it is likely that we still lack imagination, in predicting the future of this technology. The impact of new technologies is frequently much greater when they open up entirely fresh fields rather than providing new ways of meeting existing requirements. The fibre technology is likely to provide new telecommunications opportunities, but experimental systems may have to be in use for some time before the pattern emerges.

7 Acknowledgments

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Use of laser amplifiers in a glass-fibre communications system

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SUMMARY

A glass-fibre communications system using travelling-wave amplifiers operating on the laser principle is studied. The propagation characteristics of such a system are investigated theoretically with the aid of the mean value and variance of the statistical distribution of the number of photons. When linear optical amplifiers are used, the noise increases at each stage on account of spontaneous emission in the amplifier, thereby limiting the length of the transmission range. The maximum transmission range increases with the magnitude of the input signal and decreases with the gain per stage and the number modes.

1 Introduction

In future glass-fibre communications systems the signal level will have to be regenerated at repeater stations installed at intervals of several kilometres on account of the optical losses in the glass-fibre. The present paper will consider the realizability of laser amplifiers for repeater stations. Such amplifiers amplify optical signals without the need for prior conversion into electric signals. In principle they can be broadbanded low-noise models. The noise of a travelling-wave laser amplifier, in which only a single mode is able to propagate, can for instance approach the limit given by the quantum noise. Solid-state lasers, which are of particular interest in the present context, have a relative bandwidth of the order of 0·01% to 0·1%, a figure which, at the high optical carrier frequency used (approx. 350 THz), corresponds to a bandwidth of 35 GHz to 350 GHz. These two features lend interest to laser amplifiers wherever there are only a few modes and large bandwidths have to be amplified.

Direct optical amplifiers may be compared to indirect amplifiers of conventional design. In the latter the signal is converted by a photodiode from the optical frequency range to the baseband, amplified, regenerated and finally applied to an optical transmitter. This double conversion (opto-electrical, electro-optical) requires the provision of more components, e.g. a regulated power supply of 100 V to 200 V for an avalanche photodiode, than the optical amplifier. Furthermore, technical difficulties are to be expected with the low-noise baseband amplifiers at large bandwidths. With increasing bandwidth the photodetector noise produced by dark current, limited quantum efficiency, carrier multiplication and the transistor amplifier likewise increases. These problems encountered with the electronic broadband amplifier prompted us to study the feasibility of using an optical amplifier in fibre communications systems.

2 Characteristics of a Glass-fibre Communications System with Linear Laser Amplifiers

2.1 *Characteristics of a Repeater Section*

This section will first treat the characteristics of a straightforward linear-gain laser waveguide in a glass-fibre communications system, whereby the noise characteristics are of particular interest. An ideal photon detector is provided at the end of the communications system for demodulating the optical signal (energy detection). The solid angle and time domain covered by the detector during detection is characterized by the number of modes (m) detected. The number of photons in all m modes of a count is denoted by z and fluctuates between one count and the next and so determines the noise in the communications system. The frequency distribution $H(z)$, and in the limiting case the probability distribution $W(z)$, will be characterized in the following by the two parameters, the mean value† n of the random variable z and the standard deviation σ. Both parameters

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† n is therefore the mean number of photons in the space and time range comprised in the measurement.

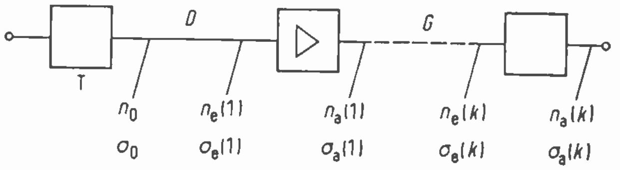

Fig. 1. Model of a communications system with k repeater sections. Each section consists of a fibre with the attenuation factor D and a travelling-wave amplifier with the power gain G.

will be calculated for a communications system consisting of k cascaded repeater sections as shown in Fig. 1. Each repeater section consists of a fibre and a linear travelling wave amplifier. The mean value n and the variance σ^2 of the photon distribution $W(n)$ will first be given for one such repeater section. The two parameters have been calculated and are discussed elsewhere[1,2] for a linear laser amplifier. If $n_e(1)$ is the mean value and $\sigma_e^2(1)$ is the variance at the input of the laser amplifier, and m modes are detected by the detector, the corresponding parameters $n_a(1)$ and $\sigma_a^2(1)$ at the output may be defined:

$$n_a(1) = Gn_e(1) + rm, \tag{1}$$

$$\sigma_a^2(1) = G^2[\sigma_e^2(1) - n_e] + Gn_e(1)(1+2r) + mr(r+1). \tag{2}$$

G denotes the gain by which the mean number of photons $n_e(1)$ representing the signal is multiplied when passing through the amplifier, and r the average number of photons produced by spontaneous emission per mode. References 1 and 3 give

$$r = (G-1)\frac{\dfrac{\varrho_2}{\gamma_2}}{\dfrac{\varrho_2}{\gamma_2} - \dfrac{\varrho_1}{\gamma_1}} \tag{3}$$

ϱ_2 and ϱ_1 denote the occupation densities in the upper and lower levels of the laser transition and γ_2 and γ_1 denote the degeneration of the corresponding state. Since $\varrho_1 = 0$ holds for an ideal 4-level process (as encountered, for example, in Nd^{3+}-doped yttrium-aluminium-garnet crystals)

$$r = G - 1. \tag{4}$$

The number m can be divided into a factor m_t which depends on the angular range (transverse modes) and a factor m_l which depends on the sampling interval (longitudinal modes) so that

$$m = m_t m_l. \tag{5}$$

If no light can reach the detector except through a waveguide, m_t will be given by the number of transverse modes excited in the waveguide. If the waveguide is designed without disturbing reflexions, m_l will be[3]

$$m_l = \Delta f \tau. \tag{6}$$

Thus Δf is the optical bandwidth of the system (given, for example, by an optical filter or by the bandwidth of the laser transition) and τ is the sampling interval over which the photons are counted. For pulse transmission the maximum pulse repetition rate in bauds is given by $1/\tau$.

It is seen from (1) that the mean value $n_e(1)$ (representing, for example, the signal) increases according to the gain of the laser amplifier, but that a noise component proportional to the number of modes m is added at the same time. In the small-signal range, this noise is independent of the signal. The mean noise power is

$$P_r = rm\frac{hf}{\tau}. \tag{7}$$

For a single transverse mode, eqns. (3) and (6) yield, for example,

$$P_r = (G-1)\frac{\dfrac{\varrho_2}{\gamma_2}}{\dfrac{\varrho_2}{\gamma_2} - \dfrac{\varrho_1}{\gamma_1}}\,\Delta f.\,hf. \tag{8}$$

The variance at the output of the amplifier is composed according to eqn. (2) of 3 terms. The first summand equals zero when $n_e(1)$ exhibits a Poisson distribution ($\sigma_e^2(1) = n_e(1)$) and so represents an ideal laser signal. The last summand gives the variance of the spontaneous emission corresponding to a Bose distribution.[3] The middle term describes the increase in variance during amplification as the result of the interference of noise and the amplified signal.[4]

In the communications system shown in Fig. 1 the glass-fibre is represented by the attenuation factor D. The mean value $n_e(1)$ of the photon distribution at the output of the fibre is derived from the mean value n_0 at the input corresponding to

$$n_e(1) = Dn_0. \tag{9}$$

D is assumed to include all the additional losses, e.g. at connectors or joints. References 1 and 2 give for the variance at the end of the fibre

$$\sigma_e^2(1) = D^2(\sigma_0^2 - n_0) + n_0 D. \tag{10}$$

This calculation assumes that no additional noise will be added along the fibre. Noise of this type is particularly liable to be produced by radiation external to the waveguide coupled in at bends and scattering centres. Let us, however, proceed from the assumption that the fibre is protected from external optical power pickup by opaque cladding.

Since the thermal noise produced at optical frequencies by the fibre attenuation is likewise negligibly small as compared to the noise from spontaneous emission, the latter represents the only considerable noise source.

The mean value $n_a(1)$ and variance $\sigma_a(1)$ at the end of the first repeater section are calculated by substituting (9) and (10) in eqns. (1) and (2):

$$n_a(1) = \varepsilon n_0 + mr, \tag{11}$$

$$\sigma_a^2(1) = \varepsilon^2\sigma_0^2 + n_0(\varepsilon + 2r\varepsilon - \varepsilon^2) + mr(1+r). \tag{12}$$

n_0 and σ_0^2 denote the mean value and variance of the input signal. The parameter $\varepsilon = DG$ indicates the extent to which the line loss is compensated by the laser amplifier. Complete compensation is reached when ε equals unity. With $\varepsilon > 1$, reflexions (e.g. at connectors) may produce oscillations if no isolator is provided.

Reflexions are liable to appear at all guide discontinuities (e.g. connectors). Also imperfections in the glass fibre (scattering centres) will produce a reflexion which is distributed along the line. The in-phase superimposition of all these reflexions results in a frequency-dependent reflexion coefficient $R(f)$ with maxima and minima.

The threshold of laser oscillation is reached when, at a frequency in the gain band Δf, coefficients R_1 and R_2 for the optical power propagating in either a forward or backward direction satisfy the condition

$$\sqrt{R_1 R_2} = \frac{1}{\varepsilon}. \tag{13}$$

For $\varepsilon > 1$, an isolator with a sufficient isolating capability must therefore be provided in order to realize a repeater section with satisfactorily stability. The rather complex optical isolator can be omitted if ε is less than unity. The repeater section will then have a small residual attenuation and the transmission range is consequently shorter.

For the mathematically straightforward case of ε equals unity, eqns. (11) and (12) are transformed into

$$n_a(1) = n_0 + mr, \tag{14}$$

$$\sigma_a^2(1) = \sigma_0^2 + 2n_0 r + mr(1+r). \tag{15}$$

It should be noted that the reflexions referred to above not only degrade system stability, but that the double reverse reflexion increases the signal distortion.

2.2 *Characteristics of Communications System*

When several repeater sections are cascaded, the output variables of one repeater section can be substituted for the input variables of the next. Appropriate substitution of eqns. (11) and (12) and the simplification of the resulting sums yields the mean value $n_a(k)$ and the variance $\sigma_a^2(k)$ after k repeater sections:

$$n_a(k) = \varepsilon^k n_0 + mr\frac{\varepsilon^k - 1}{\varepsilon - 1} \tag{16}$$

$$\sigma_a^2(k) = \varepsilon^{2k}\sigma_0^2 + \varepsilon(1-\varepsilon+2r\varepsilon)\left(\frac{1-\varepsilon^k}{1-\varepsilon}\right) \times \left[n_0\varepsilon^{k-1} + \frac{mr}{\varepsilon+1}\,\frac{1-\varepsilon^{k-1}}{1-\varepsilon}\right] + mr(1+r)\frac{\varepsilon^{2k}-1}{\varepsilon^2-1}. \tag{17}$$

For the important case of ε equals unity, these equations are far simpler:

$$n_a(k) = n_0 + mrk \tag{18}$$

$$\sigma_a^2(k) = \sigma_0^2 + 2rkn_0 + mkr(kr+1). \tag{19}$$

Along the communications system the noise increases due to the repeated cascading of attenuation and linear gain from one repeater section to the next. The mean value is composed according to (16) or (18) of the laser signal and a noise component which increases linearly with the number of repeater sections. The ratio Q_1 of the mean values of signal and noise provides a first clue to the quality of propagation along the fibre. Equation (16) yields

$$Q_1 = \frac{\varepsilon^k n_0(\varepsilon - 1)}{mr(\varepsilon^k - 1)} \tag{20}$$

and for $\varepsilon = 1$

$$Q_1(k) = \frac{n_0}{mkr}. \tag{21}$$

This ratio Q_1 must not be confused with the conventional signal/noise ratio. Although it indicates the extent to which the mean value of the signal lies above the mean value of the noise, it fails to take account of the scatter around the mean values.

According to (20) and (21), Q_1 increases with the number of signal photons n_0 injected into the front-end of the communications system. n_0 is given by the radiated power p_0 of the laser transmitter and the measurement interval τ

$$n_0 = \frac{p_0}{hf}\tau \tag{22}$$

p_0 cannot be chosen arbitrarily large because all lasers have a certain power limit and the power density in the fibre must lie distinctly below the limit for the destruction of the dielectric (or the order of 50 to 500 MW/cm^2) and below the threshold for disturbing stimulated processes (Raman or Brillouin effect).

If p_0 is maintained constant, Q_1 will increase according to (20) and (22) with the sampling interval τ and is thus inversely proportional to the electrical bandwidth.

For all further calculations the numerical value $n_0 = 10^7$ is assumed for the number of signal photons. This requires, for example, a mean radiated power p_0 of 2·34 mW over a measurement interval $\tau = 1$ ns for a 353 THz carrier (corresponding to a wavelength of 850 nm).

Q_1 is plotted in Fig. 2 as a function of the number k of repeater sections for various values of m and ε (r according to (4)). Q_1 decreases continuously along the communications system as is typical for the repeated cascading of linear attenuation and gain. Figure 2 allows only a rough estimate of the maximum transmission range. A more accurate determination is possible only after taking account of the change in variance along the transmission line. The variance increases from repeater section to repeater section according to eqns. (17) and (19) and the scattering of the number of photons around the mean value likewise increases. This divergence of the

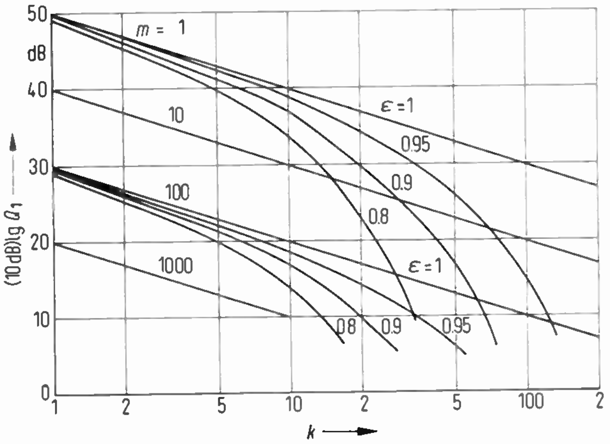

Fig. 2. Decrease in ratio Q_1 (mean value of signal/mean value of noise) with k repeater sections for various numbers of modes m calculated for an initial number of photons n_0 of 10^7, a gain of 20 dB and various values of ε.

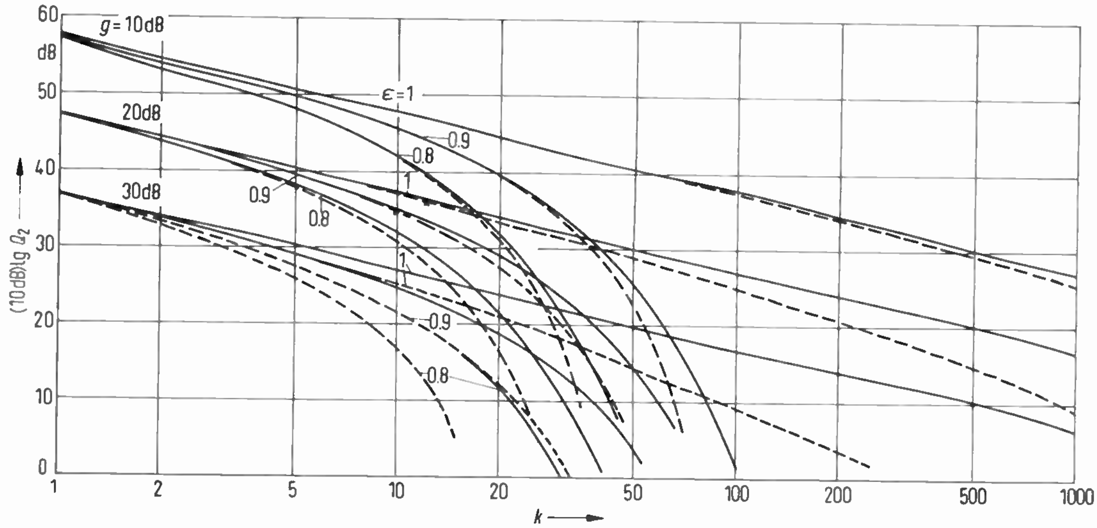

Fig. 3. Q_2 as a function of k repeater sections for $n_0 = 10^7$, $g = 20$ dB, various numbers of modes m and various values of ε.

signal distribution can be expressed by the ratio

$$Q_2(k) = \frac{n_S^2(k)}{\sigma_a^2(k)}. \tag{23}$$

n_S is the average number of signal photons behind the kth amplifier

$$n_S(k) = \varepsilon^k n_0. \tag{24}$$

Q_2 can be regarded as the ratio of the signal power to the noise power in a resistance behind an ideal photon detector if the mean noise power is electronically suppressed. Q_2 is plotted in Fig. 3 as a function of the number k of repeater sections for various numbers of modes and various values of ε, and decreases continuously along the transmission line.

To calculate the maximum possible number of repeater sections it is necessary to choose a criterion for the photon distribution $W(z)$ which still just assures the desired quality of signal transmission. A binary channel is taken as an example (Fig. 4). Noise appearing at the output of the kth laser amplifier in the blanking intervals ($n_0 = 0$) has according to (18) and (19) the mean value

$$n_P = mrk \tag{25}$$

and the variance

$$\sigma_P^2 = mrk(rk+1) \tag{26}$$

corresponding to a Bose distribution in m modes. The threshold n_S of the pulse detector must now be chosen so high above the mean noise power that the false alarm rate is sufficiently low. This requirement can be expressed by the relation

$$n_S = n_P + \eta_1 \sigma_P. \tag{27}$$

The coefficient η_1 determining quality of propagation is of the order of 5 to 10.

During the signal pulse, numbers of photons are counted whose mean value η_I is given by (18) and whose scattering is given by the variance σ_I according to (19). The difference between the threshold value and the mean signal value must be so chosen as to have a low nodetection probability. This requirement is expressed by the relation

$$n_S + \eta_2 \sigma_I = n_I. \tag{28}$$

The two requirements (27) and (28) determine the maximum possible number of repeater sections and, after eliminating n_S, yield

$$n_I = n_P + \eta_1 \sigma_P + \eta_2 \sigma_I \tag{29}$$

and, after taking (26) and (19) into consideration

$$n_0 = \eta_1 \sqrt{mrk(rk+1)} + \eta_2 \sqrt{\sigma_0^2 + 2rkn_0 + mrk(rk+1)}. \tag{30}$$

To simplify the calculation, assume $\eta_1 = \eta_2 = \eta$.

Whence it then follows from (30) that

$$n_0 = \eta^2[1 + 2rk] + 2\eta\sqrt{mkr(kr+1)}. \tag{31}$$

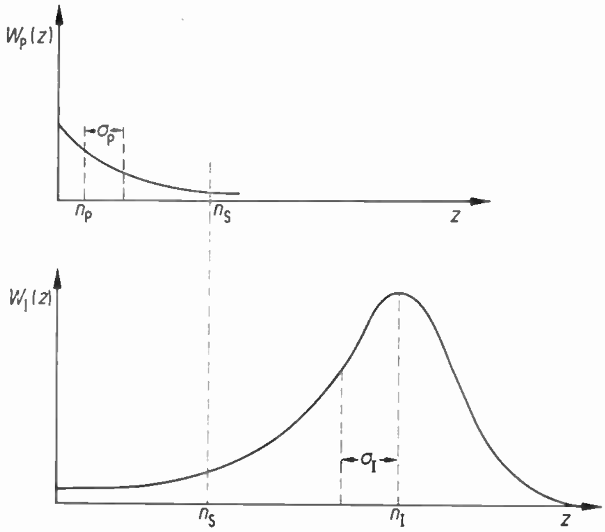

Fig. 4. Probability of z photons being counted during the blanking interval ($W_P(z)$) and the pulse ($W_I(z)$) and determination of threshold n_S.

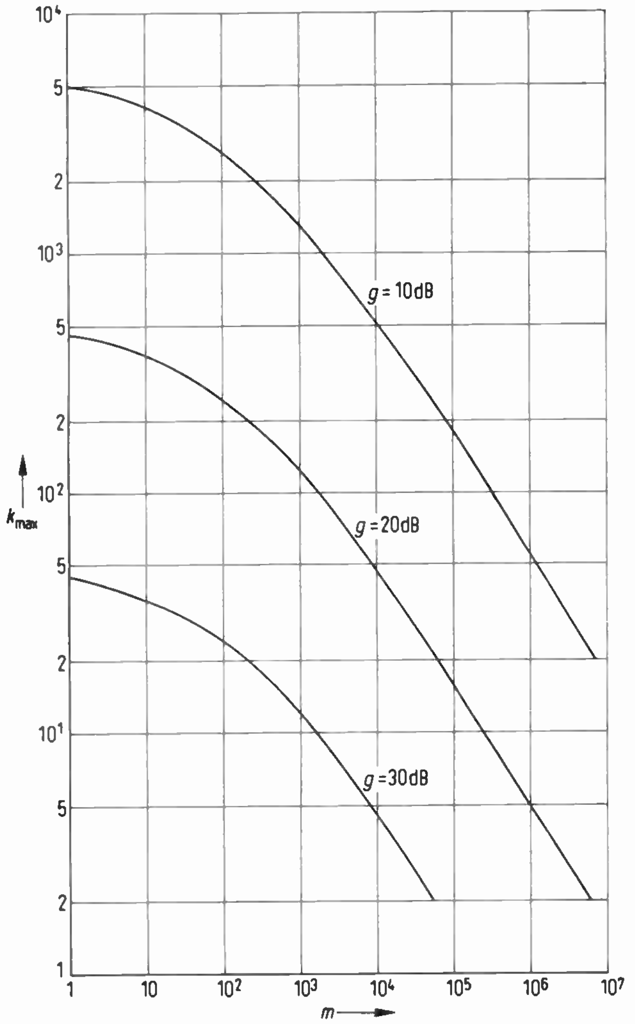

Fig. 5. Maximum possible number k_{max} of repeater sections as a function of the number of modes m for various values of g ($n_0 = 10^7$).

In the numerical examples calculated below, $k_{max} r \gg 1$. For this case (31) yields

$$k_{max} = \frac{n_0}{2r(\eta^2 + \eta\sqrt{m})}. \qquad (32)$$

The maximum possible number k_{max} of repeater sections increases with the magnitude of the input signal, but is lower for high spontaneous emission r (i.e. for high gain) and a large number of modes m.

k_{max} is plotted in Fig. 5 as a function of the number of modes m for various values of the gain $g = 10 \log G$ (r as per (4); $\eta = 10$).

The transmission range is greatest when $m = 1$. This favourable case assumes that only one transverse mode is excited in the communications system and the optical bandwidth of the system is optimally matched to the (electrical) signal bandwidth according to (6). Although a communications system of this type is feasible, it would require technologically complicated components (frequency-stable laser transmitter and narrowband filter, monomode splicing technique, etc.). It is therefore remarkable that with up to 100 modes the possible number of repeater sections decreases only slightly and that useful results are obtained even with technologically simpler multimode systems. The effect of a relatively large number of modes will be illustrated by the example of a GaAs waveguide amplifier (see below) with an evaporation-deposited dielectric filter having a bandwidth of 1 nm. This leads according to (6), and assuming a time resolution of $\tau = 1$ ns, to $m_1 = 415$ axial modes. If about 100 transverse modes propagate in the communications system, then $m = m_1 m_t = 41\ 500$.

It will be seen from Fig. 5 that, given a gain g of 20 dB, some 30 repeater sections are still possible, whereas with only 100 modes 250 repeater sections could be cascaded.

The transmission range of the communications system here treated can be calculated from the maximum possible number k_{max} of repeater sections and from the repeater section length l given by the fibre attenuation. Assuming the attenuation factor D to be due largely to the line loss α (in dB/km) of the fibre, then

$$D = 10^{-0.1\alpha l}. \qquad (33)$$

For $\varepsilon = DG = 1$

$$l = \frac{10 \log G}{\alpha} \qquad (34)$$

and the length L_1 of the transmission range attainable with linear laser amplifiers is

$$L_1 = k_{max} l = k_{max} \frac{10 \log G}{\alpha}. \qquad (35)$$

This length L_1 will now be compared with the length L_2 of the range that can be covered by an optical signal without repeaters. Without repeaters a laser signal will undergo continuous attenuation along the transmission path, while the Poisson distribution remains unchanged.[5] The maximum range is attained when, in correspondence with the criteria noted above, the mean value n of the signal is larger than the standard deviation σ by a given factor η

$$n = \eta\sigma = \eta\sqrt{n}. \qquad (36)$$

With $n = n_0 10^{-0.1\alpha L}$, L_2 can be calculated from

$$L_2 = \frac{1}{\alpha}(10 \log n_0 - 20 \log \eta). \qquad (37)$$

The use of linear laser amplifiers is meaningful whenever the ratio L_1/L_2 is considerably greater than unity.

Equations (33), (36), and (37) yield

$$\frac{L_1}{L_2} = \frac{\log G}{G-1} \; \frac{\frac{n_0}{2\eta^2}}{\log\left(\frac{n_0}{\eta^2}\right)} \; \frac{1}{1 + \frac{1}{\eta}\sqrt{m}}. \qquad (38)$$

This ratio decreases continuously with G and m, but increases with η_0/η^2. For $G = 100$ we obtain, for example, $n_0 = 10^7$, $\eta = 10$, $m = 10^3$, $L_1/L_2 = 48$.

In summary it may be said that although the maximum length of the transmission range possible with linear laser amplifiers is limited, it may be of some practical interest in the case of a high input level and relatively small number of modes.

The aspect of limited transmission range is also important in relation to the bandwidth of the communications system. Given a communications system composed of cascaded lossy repeater sections with linear amplifiers, the usable bandwidth will generally decrease with path length (even if optical equalizers are provided) as the result of delay distortion. Thus the range is likewise limited by a given bandwidth and, in the case of pulse transmission, regenerative amplifiers must be provided at appropriate intervals to regenerate the shape and phase of the pulses.

Both considerations favour the use of the simple linear laser amplifier between the terminals of transmission lines of limited length or between two pulse-regenerating non-linear amplifiers.

3 Non-linear Optical Amplifiers

Non-linear laser amplifiers used in optical pulse transmission systems have to perform the following functions:

(a) Noise reduction as a result of their nonlinear response (e.g. with threshold value).

(b) Amplification of pulse amplitude up to a given value (limiting).

(c) Shortening of signal pulses broadened by delay distortion.

(d) Pulse phase regeneration.

(e) If necessary, branching, distribution and mixing of signals.

Whereas the straightforward amplification of optical signals in a laser is readily possible by stimulated emission, these additional problems cannot be solved without the complicated methods of integrated optics. Since this technology is still in its infancy, we must here confine ourselves to enumerating briefly a few simple ways and means by which pulse regeneration can be realized in optical communications.

The first step is obviously to exploit the large signal response of laser amplifiers. If the input signal becomes so large as to be capable of drawing from the laser amplifier an amount of energy of the order of magnitude of the energy stored in the amplifier, the following effects may appear:

(a) The power gain G may decrease with increasing signal amplitude. Amplitude fluctuations of the signal pulses can in this way be reduced.

(b) Since the signal energy draws from the amplifier a large proportion of its stored energy, less remains for spontaneous emission. Thus the spontaneous emission produces less noise.[6]

(c) The pulse shape undergoes a change. In particular, pulses which began at a definite instant with the amplitude zero are shortened (for details see, e.g., Ref. 7). In a glass-fibre communications system this always occurs when proper pulses are injected into the front-end of the system and have not already merged due to the fibre dispersion.

These effects make it possible to design optical amplifiers which, although their pulse transmission quality is superior to that of linear amplifiers, do not completely satisfy the requirements enumerated above. The regeneration of the shape and phase of pulses is probably easier to realize by utilizing other non-linear effects. Saturable absorbers, which are quickly switched by a sufficiently strong optical signal from a low initial transmission to a high final transmission, are here of great interest. Substances such as solutions of organic dyes have already been used as passive Q switches in giant-pulse lasers.

Used in combination with a laser amplifier and an optical cavity, these Q switches along with other optical–optical logic circuits also open the way to an optically triggerable laser pulse transmitter with a definite threshold response.[8] The amplitude and duration of the output pulse are there independent of the input pulse and a dead time given by the pump rate of the laser and the relaxation time constant of the Q switch results over which no new input pulse will be accepted, so allowing the straightforward regeneration of the blanking interval. It should also be noted that the cascading of the laser amplifier and the saturable absorber provides a certain measure of isolation[9] because an amplified signal opens the Q switch wider than a reflected and non-amplified signal behind the Q switch.

Q switches, laser amplifiers and passive devices (cavities, filters) may thus be used in combination to build the pulse regenerating optical repeaters of the future.

For the future technical developments GaAs injection lasers and other semiconductor devices which can be used in the fabrication of lasers, laser amplifiers,[10] devices with a non-linear transmission characteristic[11] and complex optical circuits and gates[12] will be of particular interest.

4 First Experimental Results

4.1 *Experimental Set-up*

In the systems conceived for glass-fibre communications, GaAs double heterostructure (d.h.s.) injection lasers are favoured because they are small and can be easily modulated with the pump current. It was therefore logical to use these lasers as both transmitters and travelling-wave amplifiers in our experimental model. In GaAs d.h.s. injection lasers, of course, a population inversion occurs in a thin dielectric slab waveguide[13] with typical dimensions of about 1 μm and 10 μm to 100 μm across and a length of 200 μm to 300 μm.

Since the fibre guides are usually of circular cross-section, it is first necessary to solve the problem of matching the fields of two interconnected guides. No patent solution to this problem of field transformation is so far known, but various relatively complicated methods of eliminating mismatch more or less effectively are conceivable: smooth tapered transitions between the two guide cross-sections, the imaging of both cross-sections with a cylindrical lens, the use of a guide with a graded refractive index, or the use of complicated holograms with a corresponding matching effect. The whole problem can be sidestepped by choosing a fibre of rectangular core cross-section.[14] High-grade glass-fibres of this type can be fabricated.[15]

In a glass-fibre communications system operating with GaAs d.h.s. injection lasers, fibres of rectangular core cross-section provide an ideal solution to the problem of matching the laser to the fibre. Favourable matching between the two guides (laser and fibre) is obtained if both core cross-sections are approximately equal.[16]

Various experiments were performed with a glass-fibre repeater section using a GaAs d.h.s. injection laser as a transmitter, several metres of glass-fibre of rectangular core cross-section as transmission path and a GaAs d.h.s. injection laser as amplifier.

The end faces of the fibre (created by simply breaking off the fibre) were placed against GaAs front faces and adjusted with fine precision to the guide from the transmitter and the guide from the amplifier (Fig. 6). The interface was treated with appropriate fluids or adhesives to minimize reflexion. To avoid thermal problems the transmitter and amplifier injection lasers were not continuously pumped but pulsed in synchronism at a duty cycle of 10%.

Except in the saturation experiment the laser transmitter was pumped just above the threshold, so that the emission was almost completely polarized (TE modes) and very few transverse modes were injected into the fibre. The output of the amplifier could be observed with an infra-red television camera and a broadband photodiode, or connected to a further glass-fibre.

4.2 *Measurements*

The gain in active optical guides was first measured as a function of the pump current I. The power gain G resulting from stimulated emission may be expressed[12]

$$10 \log G = g = \beta I^{\delta} - \gamma \qquad (39)$$

where δ is assumed to be of the order of 2 and γ denotes the losses of the laser guide. To determine g, the pump pulse of the amplifier was chosen longer than that of the transmitter so that the amplified signal could be distinguished from the added noise with an oscillograph during the time pattern of the optical pulse at the amplifier output. Since the gain was determined by the turning on and off of the amplifier pump current, the reference point $g = 0$ includes the coupling losses and the residual absorption of the guide.

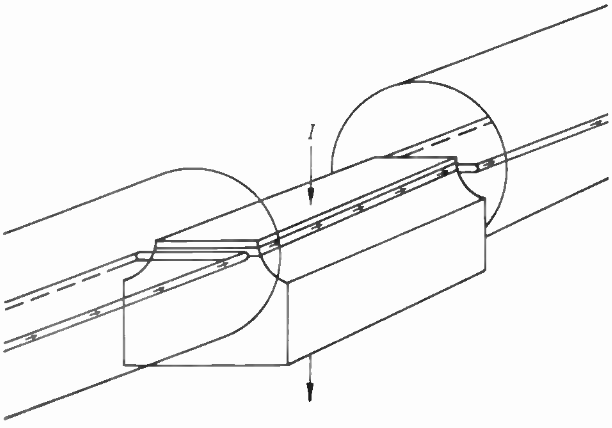

Fig. 6. Configuration of optical amplifier composed of rectangular-core glass-fibres and a GaAs d.h.s. injection laser.

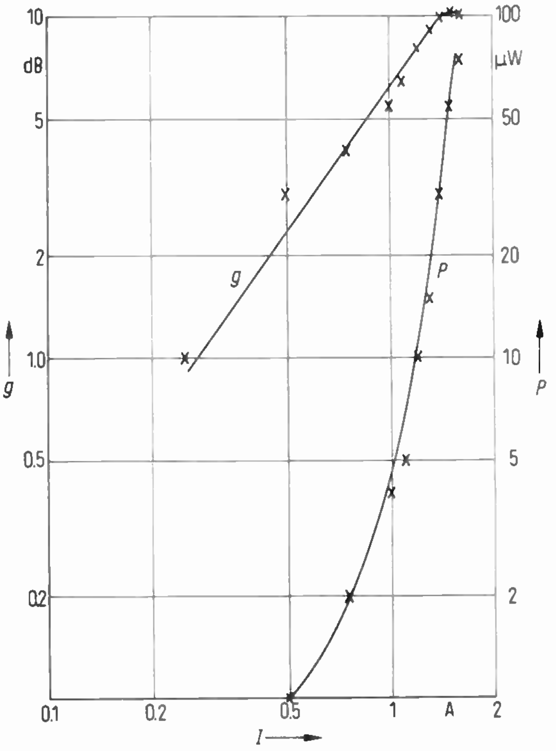

Fig. 7. Gain g and noise power P_r as a function of the pump current. The noise power was measured over an optical bandwidth of ~6 THz.

Figure 7 shows the result of this measurement. g increases with the current in correspondence with (39). For the GaAs d.h.s. injection lasers here used, Fig. 7 indicates the gain exponential δ to be 1·4.

For this measurement the faces of the amplifier injection laser were treated with a fluid with a refractive index n of 1·68, which according to the Fresnel equations represents a reflexion coefficient R of 12% per amplifier end face. The positive feedback from this reflexion limited the single-pass gain to $G = 8{\cdot}5$. Higher overall gain values could be measured due to resonance amplification. The reflexion at the double interface, GaAs-adhesive-fibre, can be reduced by depositing a $\lambda/4$ film upon the face of the GaAs injection laser by evaporation. After the deposition of an SiO film a gain of over $g = 15$ dB was measured without the appearance of any oscillations.

The mean noise power that the amplifier superposes upon the signal as the result of spontaneous emission is plotted in Fig. 7.

The increase in noise power in Fig. 7 with the pump current is approximately expressed by (8) and (39) up to just below the oscillation threshold, whereas still closer to the threshold the spontaneous emission rises sharply due to the onset of positive feedback.

It should be noted that the measured noise power corresponds to the total GaAs bandwidth ($\Delta f \simeq 6$ THz) and the noise power can be reduced by using optical filters to reduce the bandwidth.

In a further experiment the large signal response of the amplifier was investigated. With a fixed pump current the amplifier input power was varied over several orders

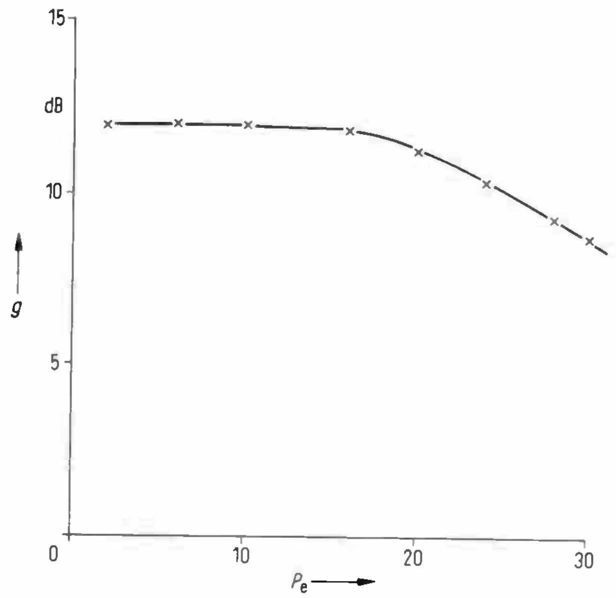

Fig. 8. Saturation of gain g at high input levels P_e.

of magnitude and the signal gain measured. Figure 8 shows the decrease in gain at high input levels due to the travelling-wave amplifier being driven into saturation.

The non-linearity was still very slight and no notable pulse distortion appeared.

The response of the amplifier is interesting when the pump pulse is made shorter than the signal pulse. With the pump current removed the amplifier exhibits according to (39) a slight absorption and the signal is somewhat attenuated. When the pump current is applied, the gain rises sharply with the current. When the pump current is again removed, the signal causes the rapid decay of the existing population inversion. The signal pulse at the output of the amplifier is amplified only for the duration of the pump pulse. This gain switching might perhaps be used for pulse regeneration.

5 Conclusions

Laser amplifiers can be used to great advantage in glass-fibre communications systems whenever there is a requirement for a broadband signal to be transmitted and only a small number of modes are excited. When repeater sections with linear laser amplifiers are cascaded, the total transmission range is limited by the summation of the noise and by the bandwidth, which depends on the length of the transmission line. Laser amplifiers are also practical for such transmission lines of limited length if the signal has a narrow electrical bandwidth as compared with the optical carrier bandwidth, so that a great many modes are excited. Optical communications systems of infinite length will become possible as soon as pulse-regenerating optical amplifiers have been realized. Such amplifiers will require in particular an optically-triggerable pulse laser.

Preliminary measurements showed GaAs injection lasers to make suitable optical amplifiers. The use of an optical fibre of rectangular cross-section is proposed. For further practical research on optical amplifiers it will be necessary to develop appropriate optical filters, isolators and delay equalizers.

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Effect of loss on propagation in multimode fibres

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SUMMARY

An analysis is given of propagation in multimode optical fibres having finite transmission loss. Using experimentally-deduced values for the cladding loss, fibre dispersions are calculated for a range of input beam widths and are compared with experiment. It is shown that the core loss has little effect on dispersion. Cladding loss reduces dispersion for a relatively small increase in attenuation, while dispersion and attenuation become non-linear functions of length.

1 Introduction

Apart from attenuation the most important parameter to consider in the application of optical fibres to long-distance communications is that of bandwidth. For single-mode cladded fibres the bandwidth depends on mode and material dispersions as well as on the frequency spread of the source and estimates have been made for various configurations[1,2] resulting in predicted values of tens of gigahertz per kilometre for a monochromatic source, falling to less than 1 GHz/km for semiconductor lasers of practical linewidths. In graded-index fibres very large bandwidths are possible[3] when a monochromatic Gaussian input beam is correctly matched to the characteristic mode of the fibre and is launched accurately along the axis. However, for less than perfect launching conditions, or in the presence of mode conversion, effectively multimode propagation occurs so that the dispersion is determined by the group delay between modes and the predicted bandwidth[3] falls to a few gigahertz per kilometre and this has been confirmed by experiment.[4,5]

The bandwidth to be expected of cladded multimode fibres was originally expected to be quite small, < 10 MHz/km, but recent measurements have shown[6] that with a monochromatic Gaussian beam, and with careful optimization of the launching conditions, values approaching 1 GHz/km are possible, although there seems to be a limiting effect due to mode conversion caused by bending of the fibre. In addition the fibres on which the measurements were made, even though exhibiting the low attenuation of 5·8 dB/km, have a high-loss cladding thereby producing a mode-filtering action which enhances[7,8] the bandwidth. In order to estimate the effect of such mode filtering on pulse dispersion and overall attenuation, and to examine whether the technique can be usefully applied in practice, we present here an analysis for propagation in a cladded, multimode optical fibre having a finite cladding and core loss.

2 Generalized Analysis

A modal analysis is complicated by the large number of possible modes in the fibres under consideration which may have core diameters between 30 μm and 100 μm compared with input wavelengths of less than 1 μm. A simplified approach is therefore taken in which the input beam is represented by a bundle of rays which are assumed to propagate along the fibre by geometrical reflexion at the core/cladding interface, provided that they fall within the numerical aperture of the fibre. The details of the theory have already been reported[9] and will not be repeated here. It has been used to obtain the impulse response for sources having Gaussian and Lambertian spatial distributions and for a source having a Gaussian temporal, as well as spatial, distribution. Numerical results have also been shown graphically for typical lossless fibre parameters and good agreement is obtained with experiment for cases where the effect of fibre loss is not serious. The theory is now applied to fibres having finite core and cladding loss and numerical results are derived for a particular combination. After deducing a value for the cladding loss from a

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probing beam experiment, a comparison is made between theory and measurement for dispersion when the input pulses are derived from a TEM_{00} mode-locked helium/neon laser.

3 Application of Theory to Lossy Fibres

As indicated above, a ray propagation model is adopted in which a ray at an angle θ to the axis, in a core of refractive index n_1, propagates by successive reflexions at the interface between the core and the cladding (refractive index n_2) providing that it falls within the numerical aperture of the fibre. The analysis assumes that propagation is by meridional rays, i.e. that there are no skew rays, which is reasonable for a Gaussian beam launched axially. Secondly, it implies that the fibre axis is straight and there is no mode conversion. In most multimode fibres the latter process could well be the dominant factor determining the bandwidth but it is ignored here since the purpose of the analysis is to isolate the effect of mode filtering and to examine its influence on dispersion and attenuation. The results are applicable to the fibres made in these laboratories, since the degree of mode conversion due to geometrical imperfections has been shown to be small, but may not necessarily apply to other fibres.

For a spatially Gaussian input beam of wavelength λ and spot size ω_0 at the beam waist, the corresponding far-field semi-angular beam width is $\theta_0 = \lambda/\pi\omega_0$. If the input pulse amplitude varies with time as

$$G_i(t) = G_1 \exp(-2t^2/a^2) \tag{1}$$

and the total input pulse energy is E_0 so that

$$E_0 = \int_{-\infty}^{\infty} G_i(t)\,dt \tag{2}$$

then for a fibre of length L the output pulse amplitude varies[9] with time as

$$G_0(t) = 2\alpha E_0 \exp[\alpha(1-2t^2/a^2\beta^2\gamma)] \times \int_1^{n_1/n_2} x \exp[-\gamma(\beta x - 2t^2/a^2\gamma)]\,T[\theta(n_1 Lx/c)]\,dx \tag{3}$$

where $\alpha = 2/\tan^2\theta_0$

$\beta = n_1 L/c$

$\gamma = \alpha/\beta^2 + 2/a^2$

and $T(\theta)$ is the transmission factor of the rays along the fibre as a function of θ It should again be noted that for axial launching of Gaussian beams there are no skew rays so that the transmission and reflexion factors are a function only of θ. For a skew ray they are functions of the angle which the ray makes with the interface as is emphasized in Section 4. The main causes of ray attenuation are absorption in the core, scattering in the core and at the core/cladding interface, together with imperfect reflexion at the interface due to finite cladding loss. The effects of absorption and scattering in the core can be represented by an effective core loss coefficient α_c and of imperfect reflexion and scattering at the interface by a reflexion factor $R(\theta)$. The ray transmission factor can thus be written

$$T(\theta) = [R(\theta)]^{\frac{L\tan\theta}{d}} . \exp(-\alpha_c L \sec\theta) \tag{4}$$

where d is the core diameter, $(L/d)\tan\theta$ is the number of reflexions at the interface and $L\sec\theta$ is the path in the core.

For a given shape of input pulse the output pulse distribution, and hence the dispersion, may be found by substituting eqn. (4) into eqn. (3) and evaluating the resulting expression.

The reflexion factor $R(\theta)$, however, depends on the state of polarisation of the reflecting ray. Thus if the complex refractive index of the cladding is denoted by $(n_2 - ik)$ then for rays with polarization perpendicular to the plane of incidence on the core/cladding interface the reflexion factor is

$$R_\perp(\theta) = \left|\frac{n_1 \sin\theta - i[n_1^2\cos^2\theta - (n_2-ik)^2]^{\frac{1}{2}}}{n_1 \sin\theta + i[n_1^2\cos^2\theta - (n_2-ik)^2]^{\frac{1}{2}}}\right|^2 \tag{5a}$$

and for polarization parallel to the plane of incidence

$$R_\parallel(\theta) = \left|\frac{(n_2-ik)^2 \sin\theta - in_1[n_1^2\cos^2\theta - (n_2-ik)^2]^{\frac{1}{2}}}{(n_2-ik)^2 \sin\theta + in_1[n_1^2\cos^2\theta - (n_2-ik)^2]^{\frac{1}{2}}}\right|^2 \tag{5b}$$

where k is the loss coefficient in the cladding and is related to the bulk cladding absorption coefficient α_k by $\alpha_k = 4\pi k/\lambda_0$. In the analysis which follows the light is assumed to be randomly polarized so that $2R(\theta) = R_\perp(\theta) + R_\parallel(\theta)$.

4 Dispersion in Liquid-core Fibre

The theory is now applied to the case of a fibre[10] consisting of hexachlorobuta-1,3-diene in Chance-Pilkington ME1 cladding and made in these laboratories. The minimum attenuation is[11] 5·8 dB/km, which gives an upper limit to the minimum core attenuation, and while it is known that the cladding loss is several orders of magnitude larger than this the actual value is not known accurately.

Measurements of dispersion have been made with a mode-locked helium/neon laser operating in the TEM_{00} mode at 0·633 μm and producing pulses of 0·65 ns half-width. The core diameter was 50 μm, $n_1 = 1{\cdot}551$ and $n_2 = 1{\cdot}4846$. The dispersion calculations therefore assume these values.

At this wavelength the total fibre attenuation is 35 dB/km and even if this were all due to the core the resulting effect on dispersion would be negligible. Thus for a core loss of this magnitude the greater path length of a ray at an angle to the axis of, say, 5° compared with that of the axial ray, produces a differential attenuation of only 0·1 dB/km. It may be concluded therefore, and more detailed calculations have also shown, that the core attenuation has little effect on the dispersion and only reflexion loss at the interface need be considered.

Since the exact value for the loss of the internal surface layer of the fibre cladding was not known, values for the reflexion loss were obtained directly by measuring the variation in attenuation of a narrow beam as the input angle of incidence θ was changed. A white light source, followed by a 0·64 μm filter with sideband suppressors, was stopped with an 0·16 cm aperture and placed 10·2 cm from the input end of the fibre to give a probe beam of 0·5° angular width. For this measurement the fibre length (47 m) was made relatively short to minimize

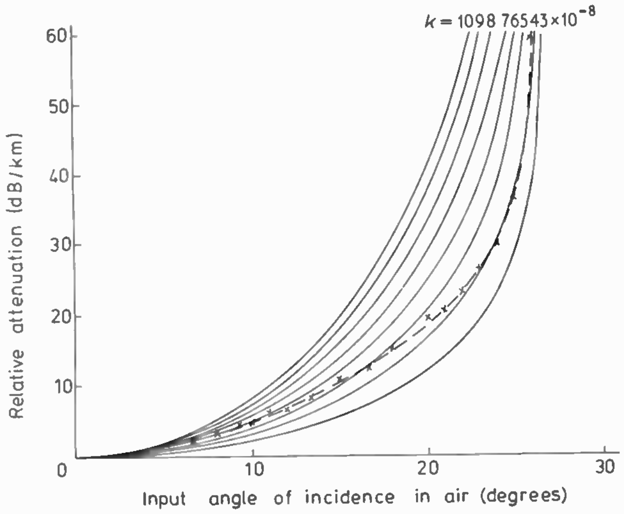

Fig. 1. Increase of attenuation of a probing beam as the angle of incidence on the fibre is increased. The solid lines are calculated for the values of k shown; the points are measured at 0·64 μm for a fibre of length 47 m and core diameter 38 μm.

the effects of mode conversion and the bend radius was 30 cm. For each value of θ the increase of transmission loss compared with the axial ray was measured and is indicated by the crosses in Fig. 1. For this method of launching most of the input power is launched in the form of skew rays so that equations (5a) and 5(b) are not directly applicable and a skew ray analysis must be used. This involves substituting in equations (5a) and (5b) the angle of incidence for a given skew ray, followed by integration over the appropriate energy distribution in the fibre (see Appendix). Curves for various values of loss coefficient k have been computed and are indicated by the solid lines in Fig. 1.

It can be seen that the ray attenuation rises relatively slowly for angles of incidence up to about 10° but increasingly sharply at larger angles particularly at about 25°. At small angles the experimental points lie along the $k = 6 \times 10^{-8}$ curve and move gradually to that for $k = 4 \times 10^{-8}$, the corresponding cladding attenuation values being 5000 dB/km and 3400 dB/km respectively. There will be some residual mode mixing present and other evidence[6] suggests that the effect will be more marked at smaller angles. Thus we feel that greater reliance should be placed on the attenuation deduced from large angle values, namely ~3400 dB/km.

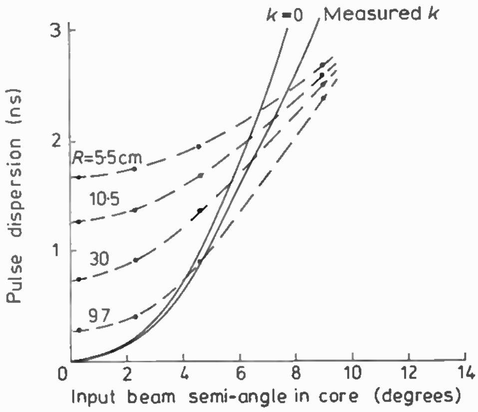

Fig. 2. Pulse dispersion as a function of beam angular width in the core for a fibre of length 180 m and core diameter 50 μm at 0·633 μm. The solid lines are calculated for the coefficients indicated and the points are measured for fibre coiled on drums of radii R shown.

The experimentally deduced values of k as a function of θ were now used in conjunction with equations (3), (4) and (5) to obtain output pulse shapes for the experimental length (180 m) of fibre for various angular widths of input beam. From the pulse half-widths the dispersions shown by the lower solid curve in Fig. 2 were plotted. For comparison the upper solid curve shows the dispersion calculated under identical conditions for the same fibre assumed to be lossless. It can be seen that the effect of cladding loss, and the resultant mode filtering due to preferential attenuation of higher-order modes (high-angle rays), is to cause a reduction in dispersion particularly in input beams of large angular width. It is not easy to make a direct comparison with experiment since the above theory assumes that the fibre is straight and that no mode conversion occurs. Experimental measurements are most conveniently carried out with the fibre coiled on a drum which can cause[6] mode conversion. Figure 2 also shows measured dispersions for the fibre wound on drums of different radii and it may be seen that at small angular widths the experimental results for increasing bend radius R_0 move asymptotically towards the theoretical curve. Unfortunately it has not been possible to lay out the fibre in a straight line in the laboratory.

At large angles the experimental dispersions are still less than those calculated for the lossy fibre and the reason for this is not certain. However, our tests have shown that the fibres from which the results were obtained were sufficiently tightly wound on the supporting drums to cause additional mode conversion, as shown by an increase in the angular width of the propagating rays.

Furthermore in a curved fibre the energy distribution is not symmetrical about the axis, as required by the analysis, but is displaced outwards and this will have an effect on the measured dispersion.

Another possible explanation is that the cladding loss is higher than that indicated by the probing beam experiment. However, further computations indicate that in order to obtain agreement at large angles with experimental dispersions extrapolated to a straight fibre it would be necessary to invoke a cladding loss of 20 000 dB/km. In fact α_k has also been measured by an independent method. Firstly, the ME1 tubing is drawn in such a way that a solid-core fibre is formed. The attenuation of this unclad fibre is then measured by hanging it horizontally and without supports along its length. The results are given in Fig. 3 and show that at 0·633 μm the loss is 2030 dB/km.

This method measures the average loss over the cross-section of the fibre whereas the probing beam indicates the loss of the internal surface layer of the liquid-filled ME1 tube which could be rather different. However, both results indicate that the cladding loss is more likely

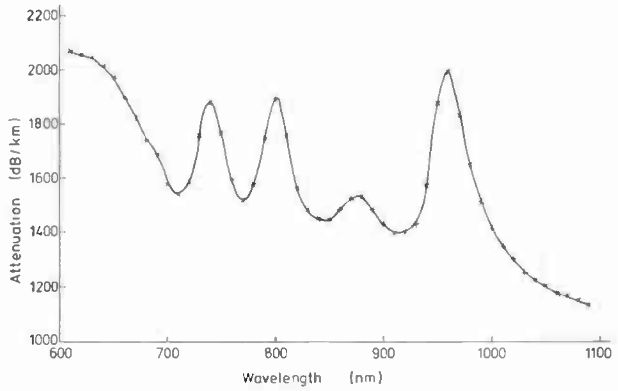

Fig. 3. Loss of unclad ME1 fibre.

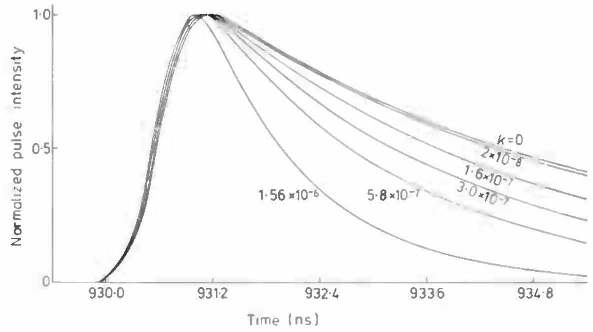

Fig. 4. Variation of output pulse shape with cladding loss coefficient at 0·633 μm for a fibre of length 180 m and core diameter 50 μm. The beam width in the core is ±8° and the Gaussian input pulse is of half-width 0·65 ns.

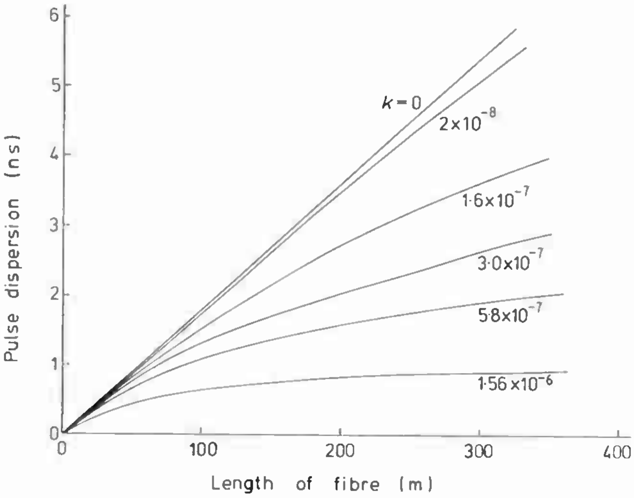

Fig. 5. Variation of pulse dispersion with length for various values of cladding loss coefficient at 0·633 μm for a fibre of core diameter 50 μm and a beam width in the core of ±8°.

to be in the region of 2–3000 dB/km rather than 20 000 dB/km, which in itself is surprising since ME1 is a sealing glass and is not intended for optical use. The low experimental dispersion for input beams of large angular width therefore cannot be completely explained in terms of high cladding loss. Nevertheless, the theory is in qualitative agreement with experiment and may be applied to the case of unstressed fibres with reasonable confidence.

5 Effect of Cladding Loss on Dispersion

Having examined the influence of cladding loss on dispersion for a particular fibre it is of interest to investigate the effect more broadly. The core loss largely determines the overall fibre attenuation but, as shown above, has little effect on dispersion for attenuations at least as high as those discussed here. Figures 1 and 2 clearly illustrate the mode filtering action and Fig. 2 also shows how the dispersion can depend on bending of the fibre. The computations have been repeated for a range of k values and Fig. 4 shows that with increased cladding loss the effect on the output pulse shape is marginally to sharpen the leading pulse edge and to shorten considerably the trailing edge. This is because the rays which are first to reach the output are those travelling at small angles to the axis and are thus least affected. As may be concluded also from Fig. 2 the pulse broadening in our liquid-core fibre, for the length and launching conditions shown, is roughly four-fifths that of an equivalent lossless fibre at an input angular width of ±8°.

The calculated variation of pulse dispersion with fibre length, Fig. 5, confirms earlier results that in a straight lossless fibre, and in the absence of mode conversion, the pulse width varies linearly with length. However, in the presence of cladding loss, when mode filtering takes place, the curves have a characteristic shape consisting of an initial rise which gradually becomes a broad maximum followed by a steady fall (for lengths greater than those shown in the figure). The length at which the maximum occurs falls with increasing cladding loss and, as the theory also shows, with smaller core diameters. A comparison of the results obtained for different angular widths of the input beam indicates that for small lengths and cladding losses the dispersion is predominantly controlled by the launching conditions while for longer lengths and high cladding loss it is the latter which is the determining factor. The dispersion thus becomes a non-linear function of length and we obtain the interesting result that the differential pulse dispersion can be made zero or even negative, as the higher-angle rays are progressively removed. A method therefore seems to be available for selecting a desired fibre dispersion or bandwidth, for a given length, by appropriate choice of cladding loss, core diameter and launching conditions.

Mode filtering occurs because of the penetration of the electromagnetic field into the lossy cladding, the depth of which is determined by the refractive index difference. We have accentuated the magnitude of the effect by substituting tetrachloroethylene, having a refractive index difference of 1·501 at 0·633 μm, for the core liquid and using a Pyrex cladding, of refractive index 1·475 and loss $\sim 2\times10^5$ dB/km. The numerical aperture was thus changed from 0·45 to 0·28 and this enabled the non-linear variation of dispersion with

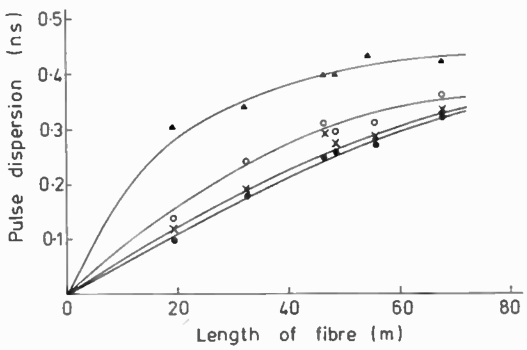

Fig. 6. Variation of pulse dispersion with length for various angular beam widths in the core as follows: △ 9·3°; ○ 4·83°; × 2·5°; ● 0·5°. The fibre consisted of tetrachloroethylene in Pyrex tubing of internal diameter 63 μm. The radius of curvature of the fibre was 5·5 cm.

length to be observed even in short lengths of fibre, as in Fig. 6. For an input beam width of 9·3° the pulse width is approaching a maximum at a fibre length of about 70 m. The measurements given in Fig. 6 are thus also in (qualitative) agreement with theory.

6 Effect of Cladding Loss on Total Attenuation

Although cladding loss can increase the bandwidth of a fibre it is important that the resulting increase in total attenuation should not be unacceptably high. There are two forms of attenuation which must be considered, namely (i) c.w. attenuation due to absorption and scattering, and (ii) peak pulse reduction due to dispersion. A normal attenuation measurement gives a value for the former which also describes the effect on total pulse energy. However, in a pulse-modulated communication system the detector normally responds to pulse height which suffers a reduction due to both absorption and dispersion of energy. As discussed above, and shown in Fig. 4, the cladding loss attenuates mainly the trailing edge of the pulse and has only a small effect on the pulse height, producing very little peak pulse reduction.

The contribution of the cladding to the c.w. transmission loss of our liquid-core fibres is only 1 or 2 dB/km at the most[11] yet it has a significant effect on the dispersion. If the cladding loss were to be increased to, say, 20 000 dB/km ($k = 3\times 10^{-7}$) then for an input beam of angular width $\pm 8°$ Fig. 5 indicates that the dispersion would be ~ 3 ns for a 0·4 km length compared to 7 ns in a lossless fibre and, because of the non-linear dependence with length, the relative difference is much greater over a kilometre. Thus the zero-loss dispersion would be 18 ns at 1 km compared with 5 ns for a 20 000 dB/km cladding.

Figure 7 shows that under the same conditions the contribution of the cladding to the c.w. fibre attenuation would be 3·6 dB for a 0·4 km length and less than 8 dB perhaps as little as 5 dB, over 1 km. This is small considering the large change in dispersion. The change in peak pulse height due to the cladding is, of course, small. An input beam of smaller angular width would experience a smaller attenuation and a corresponding lower dispersion. In fact, the cladding attenuation could be tailored to match the source used and the bandwidth and attenuation desired.

Figure 7 also shows that the attenuation per metre depends on the length over which it is measured as well as the width of the input beam and both should therefore be specified when a fibre loss is quoted. The loss falls with increasing length as the higher-angle rays are progressively removed by the cladding.

Summarizing, therefore, if a conventional detector is used, and in the absence of mode conversion, the output pulse height is not appreciably reduced by increasing the cladding loss. However, even in the lossless case the pulse energy has been dispersed so that the output pulse height is reduced although there is no loss of energy.

7 Conclusions

A theory for propagation in lossy multimode optical fibres has been developed and used in the interpretation of experimental results obtained in fibres having a high-loss cladding. At large beam angles the measured dispersion is less than that predicted by theory, possibly due to mode conversion related to distortion of the fibre on the supporting drums.

It is shown that providing the core loss is not excessively large it has little effect on dispersion. The cladding loss has been obtained using a probing beam technique giving a value of ~ 3400 dB/km which is in reasonable agreement with that found by measuring the attenuation of the ME1 tubing drawn down into an uncladded fibre. In the presence of cladding loss the dispersion is reduced for a relatively small increase in c.w. attenuation. Both dispersion and c.w. attenuation become non-linear functions of length. While the presence of cladding loss reduces the total pulse energy the pulse amplitude is little changed.

In a fibre with zero mode conversion it has been shown that the presence of cladding loss reduces the dispersion caused by rays launched at large angles to the axis. In this relatively simple case a similar result could also be

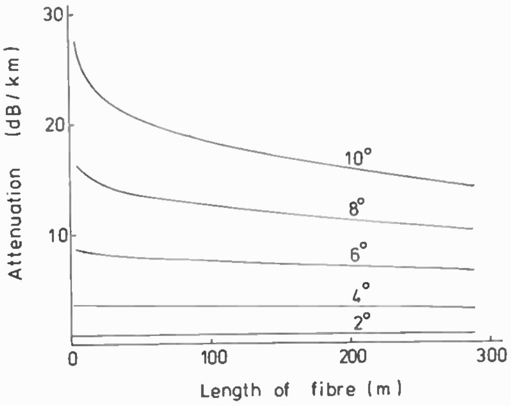

Fig. 7. Contribution of cladding loss to total fibre attenuation for the beam widths in the core shown on the curves, and for the relatively large cladding loss coefficient of 3×10^{-7}: (20 000 dB/km) at 0·633 μm. The core diameter is 50 μm.

achieved by the use of apertures[7] at the input or output or, where possible, by changing the numerical aperture of the fibre. Since each ray travels at a constant angle to the axis a better technique would involve some form of equalization of propagation times so as to reconstitute the input pulse with no loss of energy. A possible method has already been suggested.[12] In a practical fibre the situation will be complicated by mode conversion due to bends, scattering and inhomogeneities. In normal fibres where mode conversion may have a significant but not dominant effect over lengths of practical interest, cladding loss can be used to reduce the dispersion at a modest cost in attenuation of total pulse energy. On the other hand, instead of counteracting mode conversion an alternative approach[13] is to enhance it in a highly-controlled manner so as to cause a sufficient scrambling of rays (modes) over practical lengths that they have the same propagation time for any launching angle. The hazard of this approach is that rays 'converted' to angles greater than that corresponding to the numerical aperture are coupled into the cladding and constitute a loss. Whether a practical low-loss mode scrambling technique can be developed remains to be seen.

Finally, it must be commented that the ray analysis progressively fails as the number of propagating modes becomes small. The more correct modal analysis[14] must then be used.

8 Acknowledgments

The work was supported by the Science Research Council to whom grateful acknowledgment is made. One of us (H.M.) is indebted to the British Council for the provision of supplementary support.

10 Appendix: Reflexion Loss of a Probe Beam Injected into a Fibre at angle θ

A generalized skew ray analysis may be used[15] to determine the additional fibre attenuation caused by a high-loss cladding, but the numerical integration required is complex. Considerable simplification can be made in the case of a plane wave of uniform intensity entering the fibre core at an angle θ to the axis.

Each of the two symmetrical halves of the input cross-section of the core are divided into j ($j = 1, 2, \ldots m$) incremental strips of width $h = d/2m$ parallel to the plane of incidence of the input beam. For unit power incident on the core the power P_j on each of the jth strips is

$$P_j = 1 + \frac{8}{\pi d^2}\left[jh\left\{\left(\frac{d}{2}\right)^2 - (jh)^2\right\}^{\frac{1}{2}} - \left(\frac{d}{2}\right)^2 \cos^{-1}(2jh/d) - (P_{j-1} + \ldots P_2 + P_1)\right] \tag{6}$$

where $P_j = 0$ for $j \leqslant 0$.

All rays contained within the jth strip propagate down the fibre successively reflecting from the core/cladding interface at the same angle to the interface ξ_j, given by

$$\xi_j = \sin^{-1}(\cos\beta_j \sin\theta) \tag{7}$$

where

$$\beta_j = \sin^{-1} 2(j-1)h/d$$

The loss at each reflexion is found by substituting ξ_j for θ into equations (5a) and (5b), while the number of reflexions made by rays contained within this strip is $L \tan\theta/d \cos\beta_j$.

Finally, the total transmitted power $P(\theta)$ is found by numerical integration over all strips,

$$P(\theta) = \sum_{j=1}^{m} P_j R(\xi_j)^{\frac{L}{d}\frac{\tan\theta}{\cos\beta_j}} \tag{8}$$

from which the cladding induced loss in dB/km for an injection angle θ follows.

The effect of the cladding on the total fibre attenuation may be found by angular integration over a particular input distribution, weighted by results such as those shown in Fig. 1 for the loss as a function of angle.

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Design of a staggered-p.r.f. moving target indication filter

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and

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SUMMARY

To avoid blind speed phenomena, staggered-p.r.f. m.t.i. systems are used in modern radars for detection of high-speed targets. In this paper a procedure is presented which considers both the effect of filter weights and the interpulse durations. The goal of this procedure is to yield an m.t.i. response with minimum variations in the pass-band and maximum attenuation in the stop-band.

1 Introduction

An m.t.i. (moving target indicator) filter commonly known as a comb filter exhibits a periodic property. Its frequency response folds over at the radar's pulse repetition frequency (p.r.f.). Thus a target having a Doppler frequency which is an integer multiple of the radar p.r.f. will be treated as stationary clutter and will be rejected by the m.t.i. filter. This blind speed is a function of the radar operating frequency and of the p.r.f. which in turn determines the unambiguous range. For a modern radar which operates in the microwave range (L-band or above) and which is designed to have a large unambiguous range, it is difficult to avoid this blind speed problem, in particular when the radar is required to detect targets with very large velocities. One of the approaches to alleviate this problem is to use a staggered-pulse m.t.i. system in which the interpulse durations vary from pulse-to-pulse.

The design of such filters has been discussed in the literature.[1-3] However, these design approaches are concerned only with the filter weights, with the assumption that the interpulse durations are fixed. Furthermore, most other procedures determine a set of weights which yield a good clutter rejection in the stop-band region. The ripples and variations of the filter response in the pass-band are generally ignored. Although Jacomini[4] recently considered variation of the interpulse periods, his criterion of optimality is based on minimizing a linear cost function and differs from the criterion presented in this paper. The criterion of this paper is to select the weighting function which minimizes the pass-band ripple from the class of weightings which maximize the m.t.i. improvement factor. For good filter performance, the smoothness of the response in the pass-band is just as important as the attenuation in the stop-band. That is, target detectability should be uniform in the pass-band. Thus the existing methods solve only part of the problem.

A desirable solution yields a filter with maximum attenuation in the stop-band and minimum ripples in the pass-band region. To achieve this kind of characteristic, more degrees of design freedom are required. In the case of the fixed-p.r.f. m.t.i. system, these are provided by including feedback loops in the delay line network. By properly adjusting the locations of the poles and zeros of the transfer function, it is possible to design a filter having specified characteristics. This aspect has been adequately treated in the literature. However, feedback loops generally increase the filter transient response time which is undesirable for radar m.t.i. applications. For a staggered-pulse m.t.i. transversal filter, an additional degree of design freedom is provided by variation of the interpulse durations. For a given blind speed requirement, the interpulse intervals are constrained to certain permissible values which are integer multiples of the reciprocal of the blind speed. However, there are many choices of the interpulse periods satisfying the blind speed requirement. Since the various choices primarily affect the m.t.i. pass-band characteristics, we investigate a means of selecting a good interpulse code and a weighting for this code which maximizes the m.t.i. improvement factor. Thus at least in theory, one should be able to use this additional degree of freedom to one's advantage.

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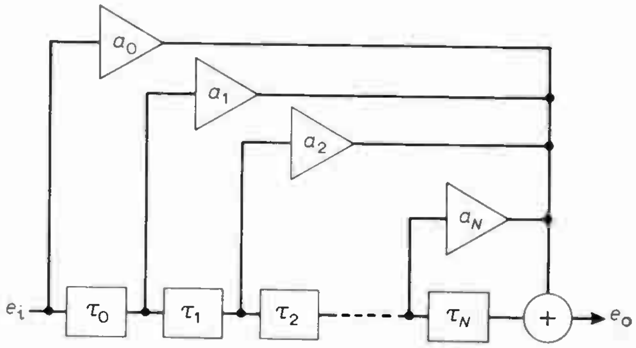

Fig. 1. A staggered-p.r.f. m.t.i. filter.

However, achievement of an optimal solution to such a problem is extremely difficult. In this paper, an attempt is made to develop an approach for solving such a problem. A solution which is based on a search technique is presented which at least yields a locally optimal solution for the imposed conditions.

2 Power Transfer Function and Average Gain

The impulse response of a staggered m.t.i. filter, shown in Fig. 1, can be represented as:

$$f(t) = \sum_{n=0}^{N} a_n \delta(t - T_n) \tag{1}$$

where

$$T_n = \sum_{i=0}^{n-1} \tau_i, \qquad T_0 = 0$$

and τ_i represents the interpulse duration.

The Fourier transform of the above expression is:

$$F(f) = \sum_i a_i \exp(-j2\pi f T_i) \tag{2}$$

where f is the Doppler frequency of the radar return.† The power transfer function is

$$|F(f)|^2 = \sum_i \sum_j a_i a_j \cos 2\pi f(T_i - T_j). \tag{3}$$

The design goal is then to find a set of weights a_i and a set of interpulse durations τ_i such that the power transfer function has the desired characteristics.

Before investigating the design conditions and procedures, it will be informative to discuss first some properties of the power transfer function.

(i) Since we are only interested in the relative amplitude of this function, the weights can be normalized by any constant and the characteristics of the function will not be changed.

(ii) If the interpulse durations are constrained such that

$$f_r \tau_i = l_i \tag{4}$$

where l_i are integers for all i then f_r is the first blind Doppler frequency. This is equivalent to saying that the frequency response of the filter is a periodic function having a period of f_r, or

$$|F(f)|^2 = |F(f + nf_r)|^2. \tag{5}$$

This becomes evident if one replaces f in equation (3) by $f + nf_r$. For the case that the interpulse durations are fixed having a value τ, then equation (3) becomes

$$|F(f)|^2 = \sum_i \sum_j a_i a_j \cos 2\pi f(i - j)\tau. \tag{6}$$

Therefore, in this case, the first blind Doppler frequency occurs at the p.r.f. which is the reciprocal of the interpulse duration τ.

(iii) The power transfer function is symmetric within each period, that is

$$|F(f_r - f)|^2 = \sum_i \sum_j a_i a_j \cos [2\pi (f_r - f)(T_i - T_j)] = |F(f)|^2. \tag{7}$$

(iv) The average value of this power transfer function is

$$\overline{|F(f)|^2} = \frac{1}{f_r} \int_0^{f_r} \sum_i \sum_j a_i a_j \cos 2\pi f(T_i - T_j) \mathrm{d}f = \sum_i a_i^2. \tag{8}$$

(v) If one assumes that white noise having a unit spectral density function

$$n_0 = 1 \tag{9}$$

is applied to this filter, the ensemble average output noise power is equal to $\sum_i a_i^2$. Thus if the filter weights are normalized by a factor $\sqrt{\left(\sum_i a_i^2\right)}$ the average output is just unity. Since both the average gain of the power transfer function and the noise power gain are equal to $\sum_i a_i^2$, the average signal/noise ratio is not degraded. Thus, one may define this as an average gain of the power transfer function.

3 Clutter Output

Assume that the power spectrum density function of the clutter return is $G(f)$, then the clutter output power is

$$C(T_i, a_i) = \int_{-\infty}^{\infty} \sum_i \sum_j a_i a_j G(f) \cos 2\pi f(T_i - T_j) \, \mathrm{d}f. \tag{10}$$

Changing the order of summations and integration, one finds

$$C(T_i, a_i) = \sum_i \sum_j a_i a_j \rho_{ij} \tag{11}$$

where

$$\rho_{ij} = \int_{-\infty}^{\infty} G(f) \cos 2\pi f(T_i - T_j) \, \mathrm{d}f,$$

which is seen to be the clutter covariance function at time $T_i - T_j$. One of the most desirable optimization conditions is to minimize this clutter power output. However, if one attempts to do so, it will lead to a trivial solution $a_i = 0$. To avoid this situation, one may require that the average gain of this clutter output be some fraction of the average gain of the power transfer function such that

$$\sum_i \sum_j a_i a_j \rho_{ij} = \lambda \sum_i a_i^2 \tag{12}$$

where λ represents the clutter suppression factor. In order to find a set of a_i with a given set of T_i such that λ is

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† The summation indices in eqn. (2) and subsequent equations vary from 0 to N unless stated otherwise.

minimum, the following conditions must be satisfied

$$\sum_j a_j \rho_{ij} - \lambda a_i = 0 \qquad (13)$$
$$i = 0, \ldots, N$$
$$j = 0, \ldots, N.$$

This is a typical eigen-value problem. There are $(N+1)$ such a_i sets which satisfy the above equation. Each of these sets comprise the eigen-vector associated with an eigen-value λ. Thus the eigen-vector associated with the minimum λ is the desired optimal solution.[1] In References 1 and 3, the average signal/clutter gain of the m.t.i. filter is optimized. This is referred to as the reference gain and is numerically equal to the reciprocal of λ. Alternatively, this procedure is equivalent to maximizing the clutter attenuation of a filter normalized to the average power gain or to maximizing the m.t.i. improvement factor.

As an example, assume that $G(f)$ has a uniform distribution function with a limited bandwidth such that

$$G(f) = \frac{1}{f_u - f_l}, \qquad f_l \le f \le f_u \qquad (14)$$
$$= 0, \qquad \text{otherwise}$$

then

$$\rho_{ij} = \frac{1}{f_u - f_l} \int_{f_l}^{f_u} \cos 2\pi f(T_i - T_j)\, \mathrm{d}f$$
$$= \frac{\sin 2\pi f_u(T_i - T_j) - \sin 2\pi f_l(T_i - T_j)}{2\pi(f_u - f_l)(T_i - T_j)}. \qquad (15)$$

If $f_l = 0$, this then becomes

$$\rho_{ij} = \frac{\sin 2\pi f_u(T_i - T_j)}{2\pi f_u(T_i - T_j)}. \qquad (16)$$

If the $G(f)$ is Gaussian and has a mean f_0 and variance σ^2, then

$$\rho_{ij} = \cos 2\pi f_0(T_i - T_j) \exp\left[-2\pi^2(T_i - T_j)^2 \cdot \sigma_c^2\right]. \qquad (17)$$

In the case $f_0 = 0$, this becomes the familiar case treated by Murakami.[3]

4 Average Gain in the Pass-band Region

Next, consider the stop-band and the pass-band regions of the power transfer function. Let the stop-band region extend from $f = 0$ to f_u and the pass-band region extend from $f = f_u$ to $f_r - f_u$. Then

$$\int_0^{f_r} \sum_i \sum_j a_i a_j \cos 2\pi f(T_i - T_j)\, \mathrm{d}f$$
$$= 2 \int_0^{f_u} \sum_i \sum_j a_i a_j \cos 2\pi f(T_i - T_j)\, \mathrm{d}f +$$
$$+ 2 \int_{f_u}^{f_r/2} \sum_i \sum_j a_i a_j \cos 2\pi f(T_i - T_j)\, \mathrm{d}f. \qquad (18)$$

Formulation of equation (18) uses the fact that the filter power transfer function is symmetric about $f_r/2$.

The last term in the above equation expresses the output of a signal which has a uniform spectrum in the pass-band. This merely states that the Doppler shift of the return target signal is not known *a priori*. Furthermore, the first term in equation (18) is equal to $\sum_i a_i^2$ from equation (8), irrespective of the value of a_i. Under this condition, the average output in the pass-band region is determined from equation (18) as

$$\frac{1}{f_r/2 - f_u} \int_{f_u}^{f_r/2} \sum_i \sum_j a_i a_j \cos 2\pi f(T_i - T_j)\, \mathrm{d}f$$
$$= \frac{f_r/2}{f_r/2 - f_u} \cdot \sum_i a_i^2 - \frac{1}{f_r/2 - f_u} \times$$
$$\times \int_0^{f_u} \sum_i \sum_j a_i a_j \cos 2\pi f(T_i - T_j)\, \mathrm{d}f. \qquad (19)$$

By regarding the a_i to be derived for a uniform spectral density function of unit amplitude from 0 to f_u and by noticing the relations of equation (16) and (13), one finds that

$$\frac{1}{f_r/2 - f_u} \int_{f_u}^{f_r/2} \sum_i \sum_j a_i a_j \cos 2\pi f(T_i - T_j)\, \mathrm{d}f$$
$$= \left[1 \pm \frac{2f_u(1 - \lambda)}{f_r - 2f_u}\right] \sum_i a_i^2. \qquad (20)$$

The left side of this equation represents the average response of this filter in the pass-band region. In the case $f_u \ll f_r$, which usually is the case when the clutter spectrum is concentrated in the region close to $f = 0$, the left side of the above equation can be approximated by the average gain $\sum_i a_i^2$. Although equation (20) was derived for a uniform spectral density over the stop-band region, the effect of any other spectral density is to modify λ. For any practical filter, $\lambda \ll 1$, so that the above relation can be generalized to include all other cases.

The mean-squared deviation from this average value in the pass-band is then

$$\varepsilon_r^2 = \frac{1}{f_r/2 - f_u} \int_{f_u}^{f_r/2} \left[\underset{i \pm 1}{\sum_i \sum_j} a_i a_j \cos 2\pi f(T_i - T_j)^2\right] \mathrm{d}f. \qquad (21)$$

In order to form a filter that has a response curve with minimum variations in the pass-band region, this error function must be a minimum. In order for this to be achieved by choice of a set of T_i, it is required that:

$$\frac{\delta \varepsilon_r^2(T_i)}{\delta T_i} = 0 \qquad i = 0, \ldots, N. \qquad (22)$$

Then, at least in theory, with the condition of equations (22) and (13), one should be able to solve for a set of T_i and a_i which yields a filter design having the desired characteristics; i.e. minimum ripple in the pass-band and maximum attenuation in the stop-band.

Unfortunately, the condition of (22) is not readily solvable in closed form by known methods.

5 A Search Solution

In view of the difficulty in seeking a solution for the above problem, we suggest here an alternative means to seek an approximate solution. This method is based on a search algorithm.

The procedure of this approach is shown in Fig. 2. An initial vector consisting of a set of interpulse periods, τ_i, is selected in a manner which is later described. The index m refers to the number of initial interpulse period vectors. The a_j represents the filter weights, which are derived from the τ_i and the clutter covariance function. With the

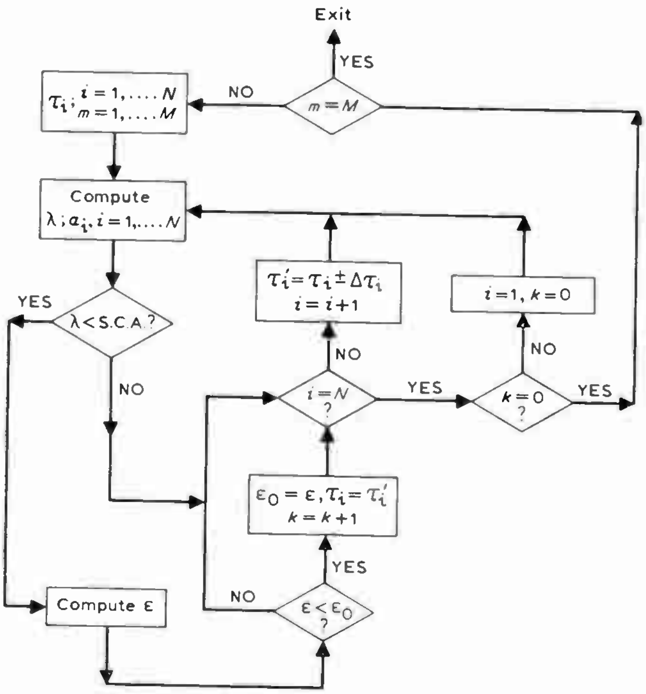

Fig. 2. Flow chart of search program.

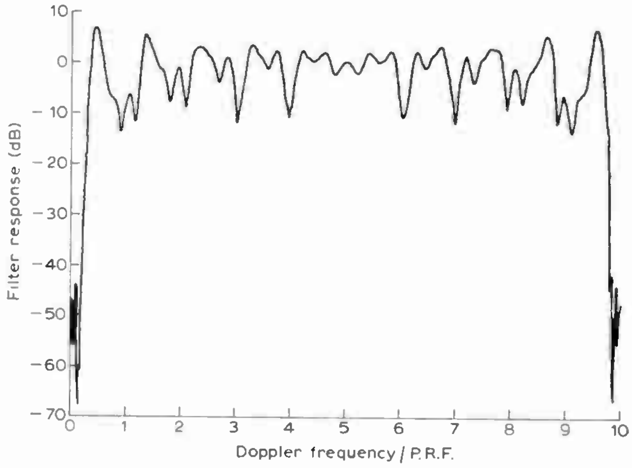

Fig. 3. Seven-pulse staggered m.t.i. filter frequency response curve. ($f_l = 0$, $f_u = 0{\cdot}2$, $N = 7$, $\lambda = 0{\cdot}000031$, $\varepsilon_r = 0{\cdot}654677$; $\tau_i = 1{\cdot}1$, $1{\cdot}3$, $1{\cdot}1$, $1{\cdot}0$, $1{\cdot}2$, $1{\cdot}0$; $a_i = -0{\cdot}060069$, $0{\cdot}198400$, $-0{\cdot}434329$, $0{\cdot}639978$, $-0{\cdot}552371$, $0{\cdot}262550$, $-0{\cdot}080401$.)

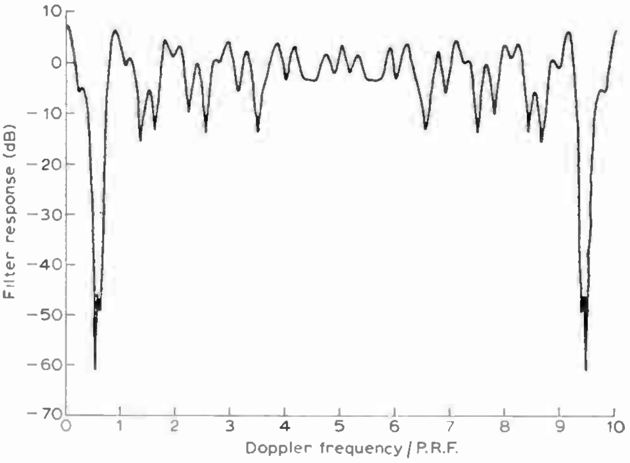

Fig. 4. Seven-pulse staggered-p.r.f. m.t.i. filter frequency response curve. ($f_l = 0{\cdot}5$, $f_u = 0{\cdot}657$, $N = 7$, $\lambda = 0{\cdot}000018$, $\varepsilon_r = 0{\cdot}695458$; $\tau_i = 1{\cdot}2$, $1{\cdot}2$, $1{\cdot}0$, $1{\cdot}1$, $1{\cdot}3$, $1{\cdot}0$, $a_i = -0{\cdot}071971$, $-0{\cdot}148622$, $-0{\cdot}532457$, $-0{\cdot}709161$, $-0{\cdot}333465$, $-0{\cdot}245260$, $-0{\cdot}122314$.)

assumed interpulse durations and by use of the relation of equation (14), a set of optimal weights with an associated eigen-value, λ, are found. If this λ is less than the specified clutter attenuation (s.c.a.), these weights are inserted in (21) to compute the value of the error function. Both the eigen-value and the error function are recorded. The next step is then to increase and decrease the interpulse durations by an amount $\Delta\tau_i$. This is done one interpulse-duration at a time. Again the eigen-value λ is computed which is then compared with the s.c.a. If it is less than the s.c.a. the error function ε is computed and the result is compared with the previous recorded ε. If ε is improved the new τ_i value, denoted as τ', is retained and another of the interpulse spacings is varied. This procedure is repeated until no further improvement in ε is obtained and the function has apparently reached a local minimum. A new set of initial values of τ_i' is assumed and the same procedure repeated again. It is assumed that an adequate number of pulses are available to achieve the required clutter attenuation. Thus, the solution obtained is the one which for a given blind speed requirement (which in turn restricts the interpulse spacings) corresponds to the minimum pass-band error from among the class of weightings which maximize the improvement factor. These weightings are determined from the interpulse periods and the clutter covariance function as previously described.

Equation (4), $f_r \tau_i = l_i$, sets the boundary conditions on the interpulse duration. Here τ_i is the interpulse duration and l_i is an integer. Hence, the variations of τ_i are limited to discrete values at steps which are the reciprocal of f_r. In practice, in order to obtain sufficient energy on a target, the interpulse time variations are bounded and therefore only a few values of τ_i need be considered in using the search program. Moreover, the minimum interpulse time is dictated in practice by the maximum range of the clutter. The constraint which requires that τ_i be discrete does not necessarily satisfy the condition of equation (22). By relying on the search method discussed above, we may find a solution which is not necessarily absolutely optimal; however the solution is always physically realizable. Furthermore, this procedure can be repeated, each time with a set of different initial values, until satisfactory results are achieved. In the following section we present several examples.

6 Numerical Examples

All examples are for 7-pulse, staggered-p.r.f., m.t.i. cancellers. To reduce local minimum problems, the search routine has been repeated about ten times. Each time a different set of initial interpulse periods is used. The best results are then plotted as a frequency response curve shown in these illustrations. The initial τ_i values are generated by a random number generator as discussed in the previous section. The whole search routine takes about two minutes on a CDC 3800 machine.

Figure 3 shows a design with a stop-band extending from $f = 0$ to $f = 0{\cdot}2$ of f_d where f_d is a normalizing factor corresponding to the reciprocal of the smallest interpulse period. This is equivalent to a filter notch of approximately 120 knots for an L-band radar at an unambiguous range of 80 nautical miles. The first normalized blind speed occurs at $f = 10$. The eigen-value which represents the clutter suppression factor is about 0·000031 and the average r.m.s. error in the pass-band is 0·654677. The interpulse durations shown in that figure are normalized with respect to the smallest τ_i, while the filter weights are normalized to the square root of the average gain. At 0 dB level, the gain of the power transfer function is equal to the average gain. In the second example shown in Fig. 4, the stop-band is set at a range from $f = 0{\cdot}5$ to $f = 0{\cdot}657$. This curve clearly demonstrates the periodic property of this type of filter. In these examples, it is assumed that the input clutter power density function is uniform, that is

$$G(f) = 1, \quad f_l \leq f \leq f_u$$
$$= 0, \quad \text{otherwise.}$$

However this same procedure can be applied to a clutter power density function with an arbitrary distribution.

7 Conclusion

The first-order effects of varying the interpulse periods and weights in an m.t.i. transversal filter are to alter the filter response in the pass-band and rejection regions respectively. The paper has shown how an acceptable filter design which is compatible with a radar waveform requirement may be achieved. By systematically varying the interpulse periods, the response variation in the pass-band or visibility region of the filter is minimized among those weights which maximize the m.t.i. improvement factor.

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Contributors to this issue*

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Since 1969 Mr. Sandbank has been Manager of the Communications Systems Division where much of the research on fibre optic communication technology at STL is carried out. He participates in several University and SRC activities aimed at improving the Industrial impact of post-graduate research and technology. He received, with co-authors, the IERE Heinrich Hertz Premium in 1965 and the Lord Rutherford Award in 1966.

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Professor W. A. Gambling (Fellow 1964, Member 1958) is Dean of the Faculty of Engineering and Applied Science at the University of Southampton, where he has been Professor of Electronics since 1964. He graduated with first-class honours in electrical engineering at the University of Bristol and was awarded the Alfred Fry Prize. He holds the degrees of Ph.D. and D.Sc. of the Universities of Liverpool and Bristol respectively.

After a period as lecturer in electrical engineering at the University of Liverpool Dr. Gambling spent two years as postdoctoral research fellow at the University of British Columbia, Canada and then joined the University of Southampton. He has also held appointments as Visiting Professor at the Bhabha Atomic Research Centre, Bombay, and at the University of Colorado. His present appointments include membership of the Technology Sub-Committee of the University Grants Committee. As well as serving as Vice-President of the Institution from 1970 to 1973, Professor Gambling has chaired the organizing committees of Institution conferences on 'Microwave Applications of Semiconductors', 'Lasers and Opto-Electronics' 'Infra-red Techniques' and the Southern Section of the IERE. He has also been a co-author of papers which have been awarded the A. F. Bulgin and Lord Rutherford premiums. Professor Gambling leads a research group which has made many contributions in the field of optical-fibre communications.

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* See also pages 664, 693 and 702.

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The frequency limitations of a gyrator circuit

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SUMMARY

An approximate analysis of the Rao–Newcomb gyrator circuit is presented. It is justified by comparison of its predictions for the y-parameters of the gyrator with predictions based on computation by general-circuit-analysis program and with predictions based on admittance measurements. The unconditional stability, passivity and performance limitations of the gyrator are discussed.

1 Introduction

At low frequencies, a gyrator with equal gyration conductances and equal short-circuit input and output conductances has a y-matrix of the form[1]

$$\begin{pmatrix} \varepsilon g & g \\ -g & \varepsilon g \end{pmatrix} \qquad (1)$$

where ε and g are real and positive.

If one port of the gyrator is terminated in a capacitance C, the driving-point impedance at the other port may be shown to have at low frequencies a reactive part which is positive and proportional to frequency, thus resembling the reactance of an inductor. The inductance of an inductor simulated in this way is C/g^2. Its Q-factor varies with frequency in a similar manner to the Q-factor of a coiled-wire inductor[2] and may be shown to have a maximum value of approximately $1/2\varepsilon$, provided that $\varepsilon \ll 1$.[1,2,3]

The realization of a transistor gyrator is usually based on two oppositely directed transconductance amplifiers connected in parallel. These amplifiers form a negative feedback loop, which at low frequencies has a very large loop gain of $1/\varepsilon^2$ when the gyrator is unterminated.

As the frequency is increased, the performance of the transconductance amplifiers is increasingly affected by capacitances within them. In particular there are parasitic capacitances across the ports of the gyrator and phase shifts associated with its transconductances. One parasitic capacitance gives a lower limit to the inductance which a given gyrator can simulate, while the other adds a parasitic capacitance across the simulated inductor giving an upper limit to the frequency at which the simulation is useful. The effect of phase shifts in the transconductances is to increase Q-factor at high frequencies (see Sect. 5) and they may produce instability of the negative feedback loop unless the loop gain is arranged to decrease sufficiently slowly with increase of frequency.[4]

Since the loop gain of a gyrator depends on its port terminations, the prevention of instability in a circuit containing gyrators must take into account the terminations with which the gyrators are to be used. The stability of networks containing gyrators is discussed further in Section 2.

2 Stability of Networks Incorporating Gyrators

A two-port network is passive if at all frequencies its y-parameters satisfy the three conditions:[5]

$$\mathrm{Re}(y_{11}) \geqslant 0 \qquad (2)$$

$$\mathrm{Re}(y_{22}) \geqslant 0 \qquad (3)$$

$$4\mathrm{Re}(y_{11})\mathrm{Re}(y_{22}) \geqslant |y_{21}+y_{12}^*|^2 \qquad (4)$$

If one or more of these conditions is not satisfied, the network is active, i.e. it is capable of supplying net real signal power and can be combined with passive elements to form an unstable network.

A gyrator described by equation (1) is passive; but at frequencies at which the phase angles of y_{12} and y_{21} are significant it is possible for the condition of equation (4) not to be satisfied, so that the gyrator is active. If gyrators are to be used as circuit elements in passive

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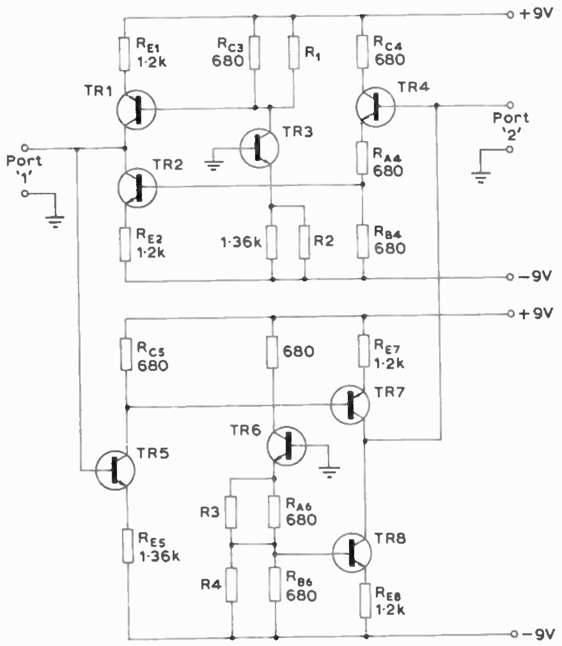

Fig. 1. Experimental gyrator circuit.

network synthesis, it is always possible to ensure the stability of complete networks by designing any gyrators within them to be passive at all frequencies.

If gyrators are used with resistors and capacitors in such a way that there is no signal flow between the ports of any gyrator other than through that gyrator,† the stability of a complete R-C-gyrator network may be ensured by designing its gyrators to be unconditionally stable. A two-port network is said to be unconditionally stable if it is stable for all combinations of passive port terminations.

A two-port network is unconditionally stable if at all frequencies its y-parameters satisfy Llewellyn's three conditions:[5]

$$\mathrm{Re}(y_{11}) \geqslant 0 \tag{5}$$

$$\mathrm{Re}(y_{22}) \geqslant 0 \tag{6}$$

$$2\mathrm{Re}(y_{11})\mathrm{Re}(y_{22}) \geqslant |y_{12}\,y_{21}| + \mathrm{Re}(y_{12}\,y_{21}) \tag{7}$$

It may be shown that the condition of equation (4) implies that of equation (7), but not conversely. Hence unconditional stability is less restrictive than passivity. The condition of equation (7) is sometimes written

$$S \geqslant 1 \tag{8}$$

where

$$S = \frac{2\mathrm{Re}(y_{11})\mathrm{Re}(y_{22})}{|y_{12}\,y_{21}| + \mathrm{Re}(y_{12}\,y_{21})} \tag{9}$$

and is known as Stern's stability factor.

It has been argued[3,6,7] that the synthesis method which leads to RC-active filters with the least sensitivity to changes in element values is that of first designing a doubly terminated LC network and then simulating each inductance by a capacitively terminated gyrator. In this important application it is not necessary for gyrators to be unconditionally stable and the complete filter will always be stable if each gyrator is designed so that when one port is terminated in a capacitance the driving-point admittance at the other port has a positive real part at all frequencies. It will be shown in Section 5 that when a gyrator has $\mathrm{Re}(y_{11}) > 0$ and port '2' terminated in a capacitance C then the real part of its driving-point admittance at port '1' is positive at all frequencies if C is sufficiently large. Hence this gyrator cannot cause instability of a complete filter provided that it is used to simulate an inductance exceeding a certain critical value. Although sufficient, this condition is not necessary for the stability of the complete filter, but if a gyrator were used to simulate an inductance below the critical value, an analysis of the complete filter would be necessary in an investigation of its stability.

The critical value of simulated inductance will be determined for the Rao–Newcomb gyrator circuit studied in this paper; but so that the gyrator may be used in other ways than as an inductance simulator its unconditional stability and passivity will also be investigated.

3 Analysis of Gyrator Circuit

The gyrator circuit to be analysed is based on that of Rao and Newcomb[9] and is shown in Fig. 1. In order to compensate for errors in bias voltages due to inaccurate resistances, resistors R1, R2, R3 and R4 are chosen experimentally.

3.1 *Transistor Equivalent circuit*

In view of its convenience for nodal analysis, the simplified hybrid-π equivalent circuit of Fig. 2 is used to represent the transistors. Experience has shown that the measured performance of a transistor circuit agrees well with the performance predicted by calculations based on this equivalent circuit at frequencies below one-tenth of the frequency of unit common-emitter short-circuit current gain. The transistors used are the complementary types BFW 31 and BFW 32. Transistors TR1, TR2, TR7 and TR8 of Fig. 1 operate with quiescent emitter

Table 1

Typical hybrid-π parameters for transistors in gyrator of Fig. 1

	I_E (mA)	$r_{bb'}$ (Ω)	$r_{b'e}$ (Ω)	g_m (S)	$C_{b'e}$ (pF)	$C_{b'c}$ (pF)
TR3, TR4, TR5, TR6,	6·25	15	520	0·25	200	7
TR1, TR2, TR7, TR8	3	15	990	0·12	96	7

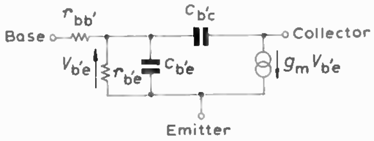

Fig. 2. Transistor equivalent circuit.

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† This would exclude their use with external feedback between their ports as in non-reciprocal cascade synthesis.[8]

currents of about 3 mA and the other transistors with quiescent emitter currents of about 6·25 mA. Typical hybrid-π parameters estimated from the manufacturer's data are listed in Table 1.

3.2 *Approximate Expressions for y-parameters*

In order to obtain insight into how the properties of the gyrator vary with frequency and with component parameters, it is helpful to find approximate expressions for the y-parameters of the complete gyrator. The approximations made appear reasonable in the case of the gyrator of Fig. 1.

All y-parameters are admittances calculated with a short circuit at one of the ports, and a short circuit at either port reduces the loop gain to zero. Hence both y_{11} and y_{22} are obtained by adding two admittances of the form of Y_2 of Fig. 3 and one of the form of Y_1 of Fig. 3. In terms of the hybrid-π parameters of Fig. 2,

$$Y_1 \simeq \frac{1}{r_{bb'}+r_{b'e}+(1+\beta_o)R_E}\cdot\frac{1+s(1+\beta_o)C_{b'c}(R_E+R_C)}{1+sC_{b'c}R_C} \quad (10)$$

$$Y_2 \simeq \frac{r_{bb'}+R_B+R_E}{R_E}\cdot\frac{sC_{b'c}}{1+sC_{b'c}(r_{bb'}+R_B)} \quad (11)$$

where s is complex frequency and $\beta_0(=g_m r_{b'e})$ is the low-frequency common-emitter short-circuit forward current gain of the transistor. Each of these expressions has been simplified by the cancellation of a zero by an approximately coincident pole and assumes that R_E, R_C and R_B are much larger than $1/g_m$ and $r_{bb'}$.

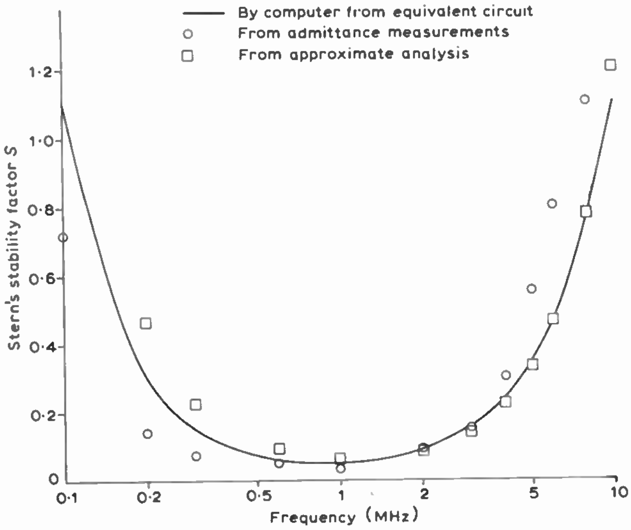

Fig. 5. Estimates of Stern's stability factor.

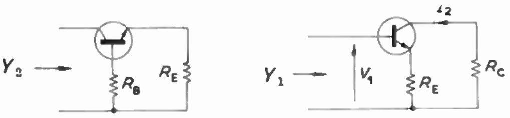

Fig. 3. Transistor stages for analysis.

In obtaining y_{12} and y_{21}, the two oppositely directed amplifiers are assumed unilateral. As a first step it is necessary to obtain approximate expressions for the transfer admittance I_2/V_1 of Fig. 3 both when R_C is large (680 Ω) compared with $1/g_m$ and $r_{bb'}$ and when R_C is zero. In terms of the hybrid-π parameters of Fig. 2, in the former case

$$\frac{I_2}{V_1} \simeq \frac{1}{R_E}\cdot\frac{1-sC_{b'c}R_E}{1+sC_{b'c}R_C} \quad (12)$$

and in the latter

$$\frac{I_2}{V_1} \simeq \frac{1}{R_E}\cdot\frac{1-sC_{b'c}R_E}{1+sC_{b'c}r_{bb'}}. \quad (13)$$

Each of these expressions has been simplified by the cancellation of a zero by an approximately coincident pole and assumes that R_E and $r_{b'e}$ are each much larger than $1/g_m$ and $r_{bb'}$.

By applying equation (10) to TR4 and TR5, equation (11) to TR1, TR2, TR7 and TR8, equation (12) to TR5 and equation (13) to TR2 and TR7, the y-parameters of the gyrator may be shown to be given in terms of the parameters of Fig. 1 and Fig. 2 by the following approximate expressions:

$$y_{11} \simeq \frac{1}{\beta_{o5}R_{E5}}\cdot\frac{1+s\beta_{o5}C_{b'c5}(R_{C5}+R_{E5})}{1+sC_{b'c5}R_{C5}}+\left(1+\frac{R_{C3}}{R_{E1}}\right)\frac{sC_{b'c1}}{1+sC_{b'c1}R_{C3}}+$$
$$+\left\{1+\frac{R_{A4}R_{B4}}{R_{E2}(R_{A4}+R_{B4})}\right\}\frac{sC_{b'c2}}{1+sC_{b'c2}\dfrac{R_{A4}R_{B4}}{R_{A4}+R_{B4}}} \quad (14)$$

$$y_{22} \simeq \frac{1}{\beta_{o4}(R_{A4}+R_{B4})}\cdot\frac{1+s\beta_{o4}C_{b'c4}(R_{C4}+R_{A4}+R_{B4})}{1+sC_{b'c4}R_{C4}}+\left(1+\frac{R_{C5}}{R_{E7}}\right)\frac{sC_{b'c7}}{1+sC_{b'c7}R_{C5}}+$$
$$+\left\{1+\frac{R_{A6}R_{B6}}{R_{E8}(R_{A6}+R_{B6})}\right\}\frac{sC_{b'c8}}{1+sC_{b'c8}\dfrac{R_{A6}R_{B6}}{R_{A6}+R_{B6}}} \quad (15)$$

$$y_{12} \simeq \frac{R_{B4}}{R_{A4}+R_{B4}}\cdot\frac{1}{R_{E2}}\cdot\frac{1-sC_{b'c2}R_{E2}}{1+sC_{b'c2}r_{bb'2}} \quad (16)$$

$$y_{21} \simeq -\frac{R_{C5}}{R_{E5}R_{E7}}\cdot\frac{1-sC_{b'c5}R_{E5}}{1+sC_{b'c5}R_{C5}}\cdot\frac{1-sC_{b'c7}R_{E7}}{1+sC_{b'c7}r_{bb'7}} \quad (17)$$

where the additional suffix to each hybrid-π parameter denotes to which transistor of Fig. 1 that parameter refers. In obtaining equations (14) to (17) it has been assumed that there is no voltage loss or phase shift in TR4 and that the collector load of TR5 is exactly R_{C5}.

It is interesting to note that the above expressions for the y_{ij} contain the collector-base capacitances for the transistors but not their emitter-base capacitances, so that to this approximation the y_{ij} are independent of the frequency at which any transistor has unit common-emitter short-circuit current gain. In addition in this

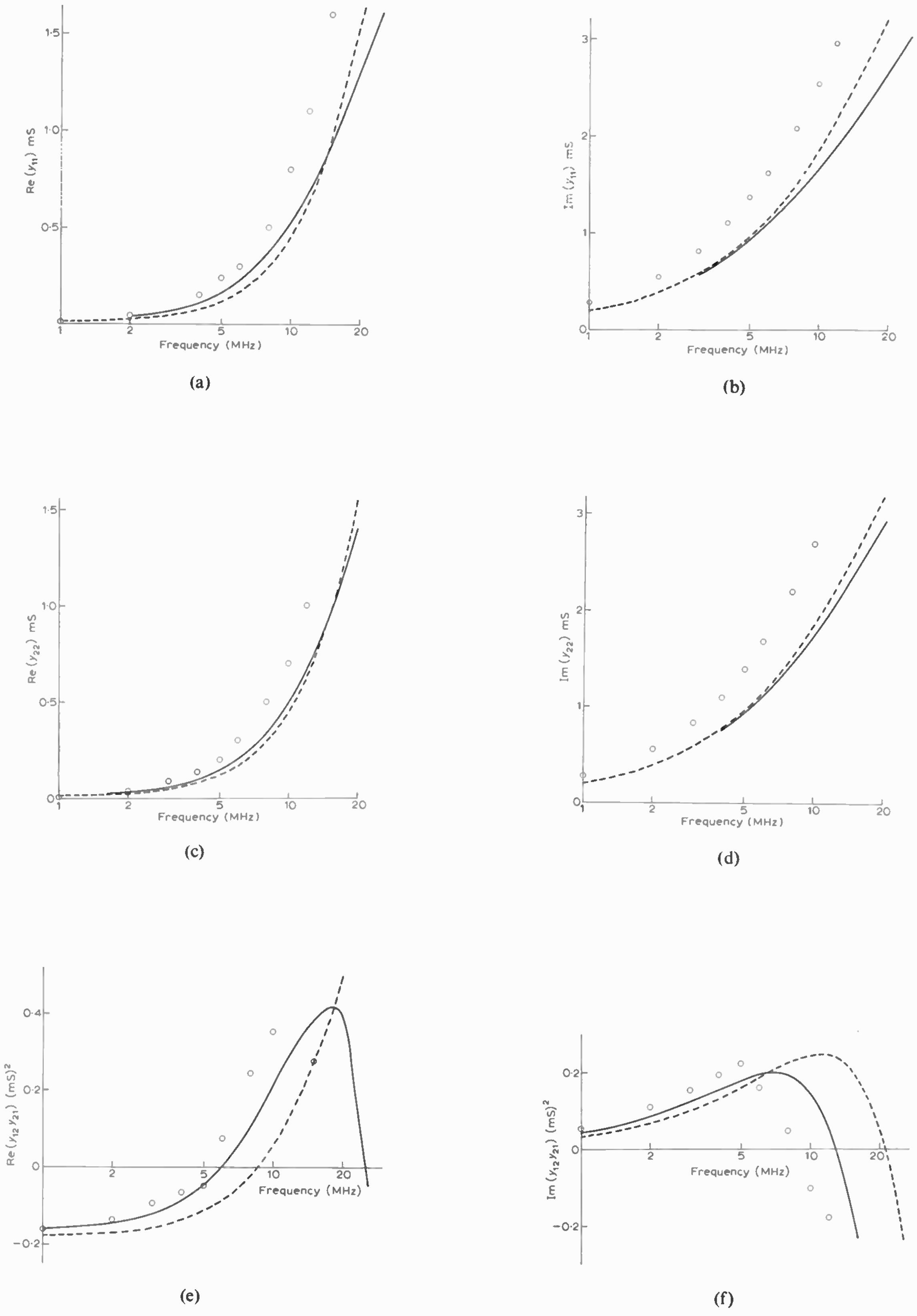

Fig. 4. Estimates of gyrator y-parameters.
---- from approximate analysis.
—— by computer from equivalent circuit.
○ from admittance measurements.

approximation, the product $y_{12}y_{21}$, which is important in Llewellyn's conditions, is independent of the low-frequency common-emitter short-circuit current gain of any transistor.

Using the numerical element values of Fig. 1 and Table 1 the following approximate expression is obtained for the y-matrix of the gyrator:

$$\begin{pmatrix} y_{11} & y_{12} \\ y_{21} & y_{22} \end{pmatrix} \simeq \begin{pmatrix} 0{\cdot}0057\,\dfrac{1+\dfrac{s'}{0{\cdot}029}}{1+\dfrac{s'}{42}} & 0{\cdot}42\left(1-\dfrac{s'}{19}\right) \\ -0{\cdot}42\,\dfrac{\left(1-\dfrac{s'}{17}\right)\left(1-\dfrac{s'}{19}\right)}{\left(1+\dfrac{s'}{33}\right)} & 0{\cdot}0057\,\dfrac{1+\dfrac{s'}{0{\cdot}029}}{1+\dfrac{s'}{42}} \end{pmatrix} \tag{18}$$

where s' denotes complex frequency in MHz and the y_{ij} are in mS. The expressions for y_{11} and y_{22} have both been simplified by replacing two real poles and one real zero which occur fairly close together by one real pole positioned so that the simplification does not affect y_{11} or y_{22} for large s'. In y_{12} and y_{21} poles occurring above 1 GHz have been omitted. For subsequent comparisons, $\mathrm{Re}(y_{11})$, $\mathrm{Im}(y_{11})$, $\mathrm{Re}(y_{22})$, $\mathrm{Im}(y_{22})$, $\mathrm{Re}(y_{12}y_{21})$ and $\mathrm{Im}(y_{12}y_{21})$ have been calculated from equation (18) and are shown as functions of frequency by the broken curves of Fig. 4(a)–(f).

It is evident from equations (14) and (15) that this approximate analysis predicts that $\mathrm{Re}(y_{11})$ and $\mathrm{Re}(y_{22})$ are positive at all frequencies. Hence the first two conditions for passivity (equations (2) and (3)) and Llewellyn's first two conditions (equations (5) and (6)) should be satisfied by the gyrator circuit of Fig. 1. In addition, as will be shown in Section 5, the input conductance of this gyrator will be positive at all frequencies when it is used to simulate a sufficiently large inductance. Substitution from equation (18) into equation (9) however predicts that Llewellyn's third condition (equation (8)) is not satisfied at frequencies between about 100 kHz and 9 MHz, so that the gyrator is not unconditionally stable and hence not passive. Predicted values of Stern's stability factor (equation (9)) in this frequency range are indicated by squares on the graph of Fig. 5.

3.3 *Exact Computation of y-parameters from Equivalent Circuit*

There are two principal sources of error in the expressions given in equation (18) for the y-parameters of the gyrator: inaccuracies in the equivalent-circuit representations of the transistors given in Fig. 2 and Table 1, and inaccuracies introduced by the simplifying assumptions described in Section 3.2. The accuracy of the transistor representations can only be verified by comparing the results of calculation with those of measurement, and this is done in Section 4.2; but the simplifying assumptions described in Section 3.2 were validated by computation in the following way.

The transistor representations of Fig. 2 and Table 1 were used to construct an equivalent circuit for the complete gyrator. The y-parameters of this equivalent circuit were then computed at a sequence of frequencies by nodal analysis using a general-circuit-analysis program (g.c.a.p.) on a digital computer. The values of $\mathrm{Re}(y_{11})$, $\mathrm{Im}(y_{11})$, $\mathrm{Re}(y_{22})$, $\mathrm{Im}(y_{22})$, $\mathrm{Re}(y_{12}y_{21})$ and $\mathrm{Im}(y_{12}y_{21})$ so obtained are shown by the continuous curves of Fig. 4(a)–(f). The differences between the broken and continuous curves of Fig. 4(a)–(f) show the errors introduced by the simplifying assumptions made in deriving equation (18) from the assumed equivalent circuit. Stern's stability factor as calculated from the computed y-parameters by means of equation (9) is shown by the continuous curve of Fig. 5.

4 Experimental Confirmation of Analysis

4.1 *Setting up of Gyrator*

The values of resistors R1 and R2 are chosen by experiment so that with port '2' terminated in a short circuit zero voltage appears at port '1'. Similarly R3 and R4 are chosen so that zero voltage appears at port '2' when port '1' is terminated in a short circuit. The large amount of d.c. negative feedback in this direct-coupled circuit ensures that each port voltage remains approximately zero even when the port terminations present open circuits to d.c. Since the gyrator is unstable with its ports unterminated, the effect on d.c. conditions of removing the short-circuit terminations cannot be investigated until the gyrator has been stabilized.

4.2 *Admittance Measurements*

Measurements of driving-point admittances were made at the ports of the gyrator at frequencies from 15 kHz to 30 MHz by means of Wayne-Kerr B601 and B701 bridges. At each frequency four admittances were measured:

(i) at port '1' with a short circuit at port '2', giving y_{11} directly,

(ii) at port '2' with a short circuit at port '1' giving y_{22} directly,

(iii) at port '1' with port '2' unterminated, giving $y_{\mathrm{oc}1}$, which is related to the gyrator y-parameters by

$$y_{\mathrm{oc}1} = y_{11} - \frac{y_{12}y_{21}}{y_{22}} \tag{19}$$

(iv) at port '2' with port '1' unterminated, giving $y_{\mathrm{oc}2}$, which is related to the gyrator y-parameters by

$$y_{\mathrm{oc}2} = y_{22} - \frac{y_{12}y_{21}}{y_{11}} \tag{20}$$

In all cases the gyrator was stable when connected to the measuring bridge.

From equations (19) and (20) two estimates of $y_{12}y_{21}$ were made and the geometric mean taken. The real and imaginary parts of this geometric mean, together with the real and imaginary parts of the measured y_{11} and y_{22}, are shown as circles on Fig. 4(a)–(f). Comparison with the curves of Fig. 4 shows that the hybrid-π equivalent circuit predicts the form of the variation of y_{11}, y_{22} and $y_{12}y_{21}$ with frequency. The tendency of predicted variations to occur at frequencies slightly higher than those at which they are observed to occur can be explained, at least in part, by the omission of wiring capacitance from the equivalent circuit.

Stern's stability factor (equation (9)), as calculated from the estimates of y_{11}, y_{22} and $y_{12}y_{21}$ made from admittance measurements, is shown by circles on Fig. 5. The good agreement between the three estimates of Stern's stability factor shows that any of the three estimates may be used as the basis of a study of unconditional stability at frequencies up to at least 10 MHz.

5 Limitations of the Gyrator as an Inductance Simulator

It may be seen from equation (18) that the susceptive parts of y_{11} and y_{22} are at low frequencies equivalent to a parasitic capacitance of about 30 pF at each port. One of these parasitic capacitances gives a lower limit of about 0·2 mH to the inductance which can be simulated by the gyrator of Fig. 1. The other parasitic capacitance appears across the simulated inductance, thus forming a parallel resonant circuit tuned to a frequency not exceeding about 2 MHz. Hence parasitic capacitance restricts the simulation of inductance to frequencies below 2 MHz, and this upper frequency limit decreases as the simulated inductance is increased above 0·2 mH.

It may also be seen from equation (18) that at low frequencies each of y_{11} and y_{22} approximates to a conductance of 5·7 μS and each of y_{12} and y_{21} to a conductance of magnitude 0·42 mS. Hence ε of equation (1) is approximately 0·0136. Thus, from the result given in Section 1, the gyrator should be capable of simulating an inductor with a peak Q-factor of about 37 in low-frequency applications. Gyrators with smaller low-frequency values of ε have been reported: for example, Sheahan and Orchard[10] have obtained peak Q-factors in excess of 400 by using a feedback pair containing a m.o.s. transistor at the input of each transconductance amplifier.

The effective value of ε increases with frequency and may be seen from equation (18) to have approximately twice its low-frequency value at 1 MHz. Hence at first sight equation (18) suggests that the quality of the simulated inductor would deteriorate unacceptably at frequencies above 1 MHz. It is well known however that the effect of phase shift in y_{12} and y_{21} is to increase Q-factors at high frequencies. To see the effect of this phase shift, assume that at a given frequency the gyrator (with parasitic capacitances removed) is described by the y-matrix

$$\begin{pmatrix} \varepsilon g & g\exp(-\mathrm{j}\phi_1) \\ -g\exp(-\mathrm{j}\phi_2) & \varepsilon g \end{pmatrix}$$

It may be deduced that when one port is terminated in capacitance C the impedance at the other port resembles that of an inductor with Q-factor

$$\frac{g\omega C}{g^2\varepsilon-\omega Cg\phi+\omega^2C^2\varepsilon}$$

where $\phi = \phi_1+\phi_2$ and only first-order terms in ε and ϕ are retained. As C varies, this Q-factor has a maximum value

$$Q^{\max} = \frac{1}{2\varepsilon-\phi}$$

which occurs when $\omega C = g$. Since ε and ϕ vary with frequency, $Q^{\max}$ varies with frequency. For the gyrator of Fig. 1, equation (18) predicts that $Q^{\max}$ becomes infinite at about 140 kHz.

It has been suggested[11] that when a gyrator is used for inductance simulation the frequency at which its Q-factor becomes infinite can be increased by compensating in its terminations for phase shift in its transfer admittances. In the case of the gyrator of Fig. 1, it is noted that assuming y_{12} and y_{21} each to have a single pole at −10 MHz would give approximately the same phase shift for $y_{12}y_{21}$ as the fuller expressions in equation (18). Assuming that the gyrator is terminated with capacitance C at each port (simulating a parallel tuned circuit), the compensation is carried out by inserting a resistance R in series with each terminating capacitance so that

$$\frac{1}{2\pi CR} = 10^7$$

Assuming that ε does not increase significantly with frequency, this compensation may be shown[11] to increase the frequency of infinite Q-factor approximately by the factor $(1+C/C_p)$, where C_p is the parasitic capacitance at each port. The increase of ε with frequency increases this factor and may result in the compensation preventing the Q-factor from becoming infinite at any frequency (see Sect. 6).

When the gyrator is terminated in capacitance C at port '2' its input conductance at port '1' may be shown to be given by

$$G_1 = \frac{(b_{22}+\omega C)^2g_{11}-(b_{22}+\omega C)N+g_{22}(g_{11}g_{22}-M)}{g_{22}^2+(b_{22}+\omega C)^2} \qquad (21)$$

where $y_{11} = g_{11}+\mathrm{j}b_{11}$, $y_{22} = g_{22}+\mathrm{j}b_{22}$ and $y_{12}y_{21} = M+\mathrm{j}N$. For the gyrator of Fig. 1, equation (18) predicts and measurements confirm that g_{11} and g_{22} are positive at all frequencies, so that the numerator and denominator of G_1 are both positive if C is large and positive. If the numerator of G_1 in equation (21) has complex zeros in C, G_1 is in fact positive for all values of C; but if the numerator of G_1 has real zeros, the larger zero is a critical value of C and G_1 is negative for a range of values of C immediately below this critical value. The critical value for a capacitive termination at port '1' may be obtained by interchanging suffixes '1' and '2' in this calculation. Critical values predicted by the y-parameters obtained by g.c.a.p. are listed in the second and third columns of Table 2. Hence it appears that a terminating capacitance of not less than 1100 pF at

Table 2. Quantities derived from g.c.a.p. predictions for y-parameters of gyrator.

Frequency (MHz)	Critical terminating capacitance at port '1' (pF)	Critical terminating capacitance at port '2' (pF)	$2\mathrm{Re}(y_{11})\mathrm{Re}(y_{22})$ (mS)^2	$\lvert y_{12}y_{21}\rvert + \mathrm{Re}(y_{12}y_{21})$ (mS)^2	$2\mathrm{Re}(y'_{11})\mathrm{Re}(y'_{22})$ (including correcting networks) (mS)^2	Margin of unconditional stability (mS)^2	$\frac{1}{2}\lvert y_{21}+y^*_{12}\rvert^2$ (mS)^2
0·1	non-existent	non-existent	0·000065	0·000058	0·000303	0·00025	0·000062
0·2	1050	1050	0·000070	0·00023	0·00160	0·0014	0·00024
0·3	1060	1040	0·000077	0·00053	0·00505	0·0045	0·00054
0·6	850	810	0·000126	0·00211	0·02568	0·0236	0·0021
1	560	510	0·00029	0·00584	0·0522	0·046	0·0059
2	200	170	0·00192	0·0232	0·0908	0·068	0·023
3	79	64	0·0076	0·0517	0·1264	0·075	0·051
4	31	22	0·0214	0·0905	0·1752	0·085	0·091
5	7·5	2·1	0·0479	0·1386	0·2432	0·105	0·14
6	negative	negative	0·0919	0·1946	0·334	0·14	0·20
8	negative	negative	0·2486	0·3241	0·600	0·28	0·33
10	non-existent	non-existent	0·5154	0·4651	0·989	0·52	0·47

either port should ensure a positive input conductance at all frequencies at the other port. Experiments show that about 1150 pF at port '1' or about 1250 pF at port '2' is required. A terminating capacitance of 1100 pF corresponds to a simulated inductance of about 6 mH, which is thus approximately the critical value of inductance referred to in Section 2.

6 Unconditional Stability

It has been predicted in Section 3.2 and confirmed in Sections 3.3 and 4.2 that the gyrator of Fig. 1 is not unconditionally stable. If unconditional stability is required, it can conveniently be ensured by loading the ports with RC correcting networks which increase $\mathrm{Re}(y_{11})$ and $\mathrm{Re}(y_{22})$ sufficiently for Llewellyn's third condition (eqn. (7)) to be satisfied at all frequencies. The choice of the gyrator ports for the connexion of correcting networks has the advantages firstly of straightforward calculation, since neither correcting network affects y_{12} or y_{21}, and secondly of requiring small capacitors in view of the high impedances at the ports.

The two sides of equation (7) are calculated from the estimates of the gyrator y-parameters obtained by g.c.a.p. and are listed in the fourth and fifth columns of Table 2. It may be seen that the addition of 0·2 mS to each of $\mathrm{Re}(y_{11})$ and $\mathrm{Re}(y_{22})$ is sufficient to satisfy equation (7) by a small margin at each frequency. To load each port with a 5-kΩ resistor would degrade the performance of the gyrator unacceptably at low frequencies (ε of eqn. (1) would be about 0·5), and hence a capacitor is required in series with each 5-kΩ resistor. The value of these capacitors should be as low as is consistent with unconditional stability to avoid unnecessary increase in the parasitic capacitances at low frequencies. Correcting networks each comprising a 56-pF capacitor in series with a 5·1-kΩ resistor would increase $2\mathrm{Re}(y_{11})\mathrm{Re}(y_{22})$ to $2\mathrm{Re}(y'_{11})\mathrm{Re}(y'_{22})$ listed in the sixth column of Table 2 and would hence satisfy equation (7) by the margins listed in the seventh column of Table 2.

It may be shown that the margins in the seventh column of Table 2 are sufficient to ensure that equation (7) remains satisfied when its two sides suffer adverse changes of up to 26%, which is probably sufficient to allow for changes of component parameters likely to be encountered in practice. If statistics of the component parameters were available, probable changes in the two sides of equation (7) would normally be calculated by computer.

The correcting networks increase each parasitic capacitance of the gyrator to about 90 pF at low frequencies, making the gyrator unable to simulate inductance at frequencies above about 700 kHz.

If an unconditionally stable gyrator is used as an inductance simulator its Q-factor remains finite at all frequencies. In terms of the discussion in Section 5 the correcting networks cause ε to increase more rapidly with frequency so that 2ε always exceeds ϕ. In fact the correcting networks cause $Q^{\max}$ for the gyrator of Fig. 1 to fall rapidly at frequencies above 100 kHz, falling to approximately 10 at 200 kHz and to less than 2 at 1 MHz.

It may be shown that when resistors are inserted in series with terminating capacitors to extend the bandwidth of the gyrator of Fig. 1 by phase compensation (as outlined in Sect. 5), the terminations serve as correcting networks for obtaining unconditional stability provided that the terminating capacitances exceed about 500 pF.

7 Passivity

Sufficient conditions for the passivity of a two-port network are given as equations (2), (3) and (4). Equations (2) and (3) are identical to equations (5) and (6) and have been shown to be satisfied at all frequencies by the gyrator of Fig. 1 without applying any correction to the gyrator. In order to predict whether the correction for unconditional stability has also ensured passivity, the quantity $\frac{1}{2}\lvert y_{21}+y^*_{12}\rvert^2$ has been calculated from the g.c.a.p. estimates of the y-parameters and is tabulated in the eighth column of Table 2 for comparison with $2\mathrm{Re}(y'_{11})\mathrm{Re}(y'_{22})$. It is seen that $2\mathrm{Re}(y'_{11})\mathrm{Re}(y'_{22})$ exceeds $\frac{1}{2}|y_{21}+y_{12}^*|^2$ at all the frequencies of Table 2, predicting that with the correcting networks the gyrator is passive.

8 Conclusion

In this paper the Rao–Newcomb circuit is analysed and its performance discussed. The analysis presented is based on the hybrid-π equivalent circuit and leads to approximate expressions for the gyrator y-parameters. The poles and zeros of these expressions are independent of the emitter-base capacitances of the transistors.

The accuracy of the analysis is demonstrated by comparing its predictions for the gyrator y-parameters with values obtained by computer analysis of the equivalent circuit and with values derived from admittance measurements.

When constructed with transistors with low-frequency common-emitter current gains of about 125, the gyrator is capable of simulating an inductor with Q-factor 37 at low frequencies. If the collector-base capacitance of each transistor is about 7 pF, simulated inductances less than 6 mH may have negative Q-factors over a range of frequency above 140 kHz (unless phase compensation is used) and the performance of the gyrator is limited by a parasitic capacitance of about 30 pF at each port.

It is shown to be straightforward to render the gyrator unconditionally stable and passive by connecting simple RC correcting networks at its ports. This increases the parasitic capacitance at its ports to about 90 pF. When an inductance is simulated by the gyrator corrected so as to be unconditionally stable, the maximum Q-factor obtainable falls off rapidly above 100 kHz.

9 Acknowledgments

The Author wishes to thank Mr R. H. Grigg of the Electronics Branch, Royal Military College of Science, for assistance in carrying out the circuit analysis computations and Prof. P. C. J. Hill, Head of the Electronics Branch, Royal Military College of Science, for encouragement and helpful criticism.

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The Author

Dr. H. G. Brierley read mathematics and physics at King's College, Cambridge, obtaining the B.A. degree in 1954 and a Certificate in Education in 1955. Following National Service he taught mathematics for a year and in 1958 he joined G.E.C. at Telephone Works Coventry as a laboratory engineer in the group responsible for the design and development of passive networks used in carrier telephony. In 1960 he moved to development of quartz-crystal oscillators, repeater amplifiers and other active networks required in carrier telephony, being appointed a Group Leader in 1963.

During the period 1963-7, Dr. Brierley carried out part-time research at the Lanchester College of Technology, Coventry on the design of negative-feedback transistor amplifiers and was awarded a Ph.D. degree by the C.N.A.A. in 1968 for this work.

In 1969 he took his present post as Senior Lecturer in the Electronics Branch at the Royal Military College of Science at Shrivenham, where he is mainly engaged on teaching telecommunications and electronics to undergraduates.

IERE News and Commentary

Applications for Transfer and Election

The introduction of new regulations for membership and particularly those which have altered the academic qualifications required from 1st January 1974, has meant that the Membership Committee is currently dealing with a very considerable number of applications based on the existing regulations. It is obviously essential that every priority should be given to such applications to enable them to be considered by the end of 1973. Accordingly the Committee hopes that applicants whose proposals are not affected by the new Bye-laws will appreciate that progressing their proposals now take rather longer than would normally be expected. Their patience, if delays occur, is therefore requested.

Reduced Rates for IEE Publications

Under the mutual arrangements existing between the two Institutions, members of the IERE may subscribe to IEE periodicals at the following reduced rates which apply for 1974. (The subscriptions to non-member customers are shown in brackets.)

Electronics and Power	£12·50	(£16·50)
Single copy	1·25	(1·65)
Proceedings IEE (p or m*)	38·25	(51·00)
Combined p or m	57·50	(76·50)
Single copy	3·75	(5·00)
Electronics Record	13·50	(18·00)
Single copy	3·75	(5·00)
Power Record	13·50	(18·00)
Single copy	3·75	(5·00)
Control and Science Record	13·50	(18·00)
Single copy	3·75	(5·00)
Electronics Letters (p or m)	23·25	(31·00)
Combined p and m	35·00	(46·50)
Single copy p or m	2·25	(3·00)

* Paper or microfiche.

Rates for airmail supplement will be supplied on request. Members are asked to place their orders through the IERE Publications Department, 9 Bedford Square, London WC1B 3RG, to ensure that they are granted the special rate.

The Graham Clark Lecture

The 1974 Graham Clark Lecture by Sir Kingsley Dunham, F.R.S., on 'Natural Resources, The Engineer and the Environment' will now take place on Tuesday 8th January, at 6 p.m. at the Institution of Electrical Engineers, Savoy Place, London WC2 (*not* at the I.C.E. on Wednesday, 9th January, as previously announced).

Membership Subscriptions for 1974

The impossibility of delaying the revision of members subscription rates was referred to in the Annual Report of the Council, published in the September 1973 Journal. At its meeting in October the Council approved the Executive and Finance Committees' recommendations for increases in subscriptions under Bye-Law 26 (old Bye-Law 24), to take effect from 1st April 1974 as follows:

Fellows	£18·50
Companions	18·50
Members	16·00
Associates	14·00
Graduates*:	
over 35 years	14·00
25–34 years	10·00
25 years	7·50
Associate Members*:	
over 35 years	14·00
25–34 years	10·00
up to 25 years	7·50
Students:	
over 25 years	6·00
under 25 years	3·50
undergoing an approved course and not being in receipt of salary	1·00

* A reduced annual subscription of £3·50 is payable by Graduates and Associate Members undergoing an approved course and not in receipt of salary.

The cooperation of members of all grades in arranging to remit subscriptions by the Direct Debit System is especially urged as this method enables the costs of collection and accounting to be kept to a minimum.

Appleton Commemorated in Renaming of SRC Laboratory

The Secretary of State for Education and Science, Mrs Margaret Thatcher, recently visited the Radio and Space Research Station near Slough in Buckinghamshire, to attend a small ceremony at which she unveiled a plaque commemorating the change of name of the Station to The Appleton Laboratory. The new name has been chosen in view of the very close connexion over a long period with the work of the Station of the distinguished physicist, the late Sir Edward Appleton, C.B.E., K.C.B., F.R.S., who was also Secretary of the former Department of Scientific and Industrial Research from 1939 to 1949.

The Science Research Council has also agreed that the Rutherford High Energy Laboratory and the Daresbury Nuclear Physics Laboratory should in future be known as The Rutherford Laboratory and The Daresbury Laboratory, respectively. These name changes take effect immediately and are in line with the Council's policy to make use as widely as possible of the expertise which is available in the laboratories. No changes in name are contemplated for the Royal Greenwich Observatory, the Royal Observatory, Edinburgh, or the Atlas Computer Laboratory.

Correction

The following addition should be made in the paper 'Fault diagnosis using time domain measurements' published in the September 1973 issue of *The Radio and Electronic Engineer* (Vol. 43, No. 9):

Page 532, Section 14: *Add*

16. Hayes, K. A., 'The necessity for test points in servomechanisms', Report No. REMC/28/FR, Ministry of Technology, May 1969.

Members' Appointments

CORPORATE MEMBERS

Mr. Frank Cox (Fellow 1958) has been awarded the degree of Master of Philosophy from Brunel University; his thesis was a study of a system for controlling urban taxi cabs. Mr. Cox was previously Chief Engineer of J. Langham Thompson Ltd. and subsequently Technical Director of Ether Engineering Ltd.

Mr. A. A. Kay (Fellow 1959, Member 1953) has been appointed Director of Engineering of Rediffusion Consumer Electronics Ltd., a new company formed from Rediffusion Vision and Rediffusion Vision Service; Mr. Kay had been Chief Engineer of the former company since 1965. He has served on the Communications Group Committee (formerly the Television Group) since its formation.

Dr. Konstantin Apel (Member 1965) is now Director of Engineering of Eaton GmbH, Meersburg, Germany. He was formerly Director of W. Holzer & Co., also of Meersburg. While with Elesta A G of Switzerland, Dr. Apel contributed a paper on a gas-filled counting tube to the 1963 conference on Cold Cathode Tubes which was subsequently published in the Journal.

Mr. K. F. Baker (Member 1969) has been awarded the degree of M.Sc. in the physics of semiconductor devices by CNAA following study at Lanchester Polytechnic. Mr. Baker was previously an engineer in the Advanced Laboratory of Painton & Co. Ltd.

Mr. A. M. Bell (Member 1962, Graduate 1960) has joined Aviation Electronics Ltd. of Montreal as Supervisor of Communications and Avionics systems. His previous appointment was with the Canadian Marconi Co. which he joined in 1970 following his retirement from the RAF with the rank of Squadron Leader.

Mr. A. M. A. Bowman (Member 1971, Graduate 1962) has taken up the post of Assistant Teacher of mathematics at Easthampton Park Education Centre, Bracknell. Mr. Bowman has been with the BBC since 1959, latterly as a member of the Technical Management and Training Section of the Film Services Department.

Mr. P. L. Brown, B.Sc., M.Eng. (Member 1963), formerly Head of the Electronic Engineering Department at Letchworth College of Technology, has been appointed Principal of the George Stevenson College, Watford.

Flt. Lt. B. V. Canton RAF (Member 1973) who has been at the RAF College, Cranwell, as a lecturer on computers since 1971, has been posted to RAF Medmenham.

Mr. M. J. Clements (Member 1969, Graduate 1966) has been appointed Group Engineering Manager with RS Components Ltd. He was previously with Hugin Cash Registers Ltd. as Manager of the Electronic Applications Organisation since 1966.

Capt. P. C. V. Dolan (Member 1970, Graduate 1964) who relinquished his appointment in the then Ministry of Aviation Supply on retirement from the Royal Signals, is now with AEG-Telefunken, Backnang, West Germany, as a member of the project management team concerned with the ESRO modular repeater and the communication orbital test satellite programmes.

Mr. D. M. Embrey (Member 1973) has been appointed Technical Director of AB Electronic Components Ltd., Abercynon, Glamorgan. Mr. Embrey was previously Chief Engineer of Abergas Ltd., Caerphilly, Glamorgan.

Mr. I. M. Findlay (Member 1973, Graduate 1969), formerly an Electronic Design Engineer with the British Aircraft Corporation, Bristol, has joined McMichael Ltd., Slough, as a Development Engineer.

Mr. A. G. Fitzgerald (Member 1972, Graduate 1969) has been appointed Project Quality Manager with Marconi Space & Defence Systems Ltd., Stanmore. He was previously Deputy Quality Manager with Microwave Associates Ltd.

Lt. Cdr. J. Grady, RN (Member 1973, Graduate 1968) has been posted to the Air Engineering School (HMS *Daedalus*) as Training Design Technology Officer. He was for the previous seven years in HMS *Neptune*, latterly as Head of the Navigational Aids Section.

Sqn. Ldr. A. Griffiths DUS, BSc, RAF (Member 1970) has been appointed Lecturer in Electrical Engineering at the Royal Naval College, Manadon. He was previously lecturing in electronics at the RAF College, Cranwell.

Mr. D. A. Hall (Member 1973, Graduate 1968), formerly Senior Engineer with Reditune Ltd., is now Audio Engineer with Megawatt Engineering Ltd., Edenbridge.

Mr. T. Hughes (Member 1966, Graduate 1961) is now with Systems Designers Ltd., Frimley, as Marketing Executive (Communications).

Mr. D. T. Law (Member 1970, Graduate 1965) has joined Software Sciences Ltd., Farnborough, Hampshire, as Senior Consultant. He was previously with EASAMS Ltd.

Capt. R. J. Loader, REME (Member 1972, Graduate 1966) has been posted to the School of Electronic Engineering, Arborfield, as Officer in Charge of the Light Air Defence Radar Section. Capt. Loader was Project Officer with the Joint Services Reliability and Maintainability Advisory Team at Decca Radar Ltd.

Inst. Lt. Cdr. K. A. Lowen, RN (Member 1971, Graduate 1963) is now Base Instructor Officer, Gibraltar, and Assistant Principal of Gibraltar & Dockyard Technical College. He was previously in the Control Weapons Group of HMS *Collingwood* as Senior Instructor Officer.

Mr. W. F. Maddox, BSc. (Member 1968) has joined KGM Betriebsgesellschaft für electronische Anlagen mbH, Munich, as Control Systems Engineer. Mr. Maddox was previously an Electrical Controls Engineer with the British Aircraft Corporation, Bristol.

Mr. E. F. Munroe (Member 1966, Graduate 1962) has joined British Electronic Controls Ltd. as Group Marketing and Sales Manager. He was previously with ICFC-Numas Ltd.

Mr. C. W. W. Read (Member 1971) formerly Chief Engineer at Computing Techniques Ltd., has been appointed General Manager with Amplicon Electronics Ltd. of Hove, Sussex.

Major B. Reavill, REME (Member 1965) has relinquished his position as Officer Commanding, 19 Field Workshop (Electronic), REME, and has now taken up the appointment of Officer in Charge of No. 10 Maintenance Advisory Group, Technical Group REME, at SRDE, Christchurch.

Mr. T. P. Reid (Member 1960, Associate 1959) has joined Ultra Electronics Ltd. as an Executive Engineer in their Development Laboratory at Greenford. Mr. Reid was previously with Redifon Ltd. for two years following five years working in the United States and Canada.

Sqn. Ldr. F. B. Sansom, RAF (Member 1972, Graduate 1970) has been posted to HQ Strike Command, RAF High Wycombe (Electrical Engineering (Comms 3)). He previously held a Signals appointment at the Ministry of Defence.

Mr. J. A. Selby (Member 1972, Graduate 1967) has moved from Mullard Ltd. where he was a Senior Electronic Design Engineer, to Newmark Triplite Ltd., Croydon as a Senior Engineer.

Mr. S. A. Wales (Member 1963, Graduate 1957) is now General Manager and Managing Director of Semtech Ltd. Mr. Wales was previously with Bourns (Trimpot) Ltd.

Mr. R. B. Ward (Member 1970) has been appointed Group Leader – Security Systems with Marconi-Elliott Avionic Systems Ltd. Borehamwood, Herts. He was previously a Project Engineer.

Mr. N. W. White (Member 1971) who has been with Marconi Instruments Ltd. since 1958, has been promoted to Manager of the Communications Division; previously he was Chief Design Engineer for Television Applications.

Mr. L. W. B. Worrall (Member 1971, Graduate 1968) is now Manager, Special Projects with Foxbord Yoxall, Redhill. For the past year he has been Project Administrator with the Company.

NON-CORPORATE MEMBERS

Mr. E. O. Awala (Graduate 1969) has been promoted to Acting Area Engineer in the Department of Posts & Telecommunications, Calabar, Nigeria.

Mr. R. Chapman, B.Sc. (Graduate 1967) who has been a Research Student at the University of Newcastle-upon-Tyne for the past three years, has been appointed a Lecturer in Electrical Engineering at Paisley College of Technology.

Mr. A. E. Christoforou (Graduate 1972) is now Chief Engineer in charge of the Electronics Department with A. G. Andronicou Co., Ltd., electrical and electronics engineers in Limassol, Cyprus.

Mr. W. Finnie (Graduate 1968) has been appointed Area Manager with Labhire Ltd., Whitburn, West Lothian.

Mr. S. O. Foye (Graduate 1971) has been appointed Head of Electrical and Instrument Department of the Nigerian Paper Mill Ltd., Jebba, Kwara State, Nigeria. Mr. Foye obtained the Higher National Diploma in Electrical and Electronic Engineering and the College Diploma of South East London Technical College in 1970.

Mr. P. R. Hayns (Graduate 1964) is now a Senior Development Engineer with Trend Communications Ltd., High Wycombe.

Mr. J. Henshaw (Graduate 1970) has been promoted to Principal Engineer in the Plessey Company and is working as a Quality Planning Engineer on Transmission and Electronic Exchanges.

Mr. P. G. Howard (Graduate 1968) who was with Fielden Research Ltd. from 1971 to the summer of this year, has joined Hewlett Packard Inc., Colorado Springs, USA, as an Electronics Engineer.

Mr. E. V. Jansz, B.Sc. (Graduate 1971) who has been working for the past two years in Ceylon, has been appointed an Engineer with L. M. Ericsson (Pty) Ltd., in Australia.

Mr. P. T. Khoo, B.Sc, M.Phil., Ph.D. (Graduate 1963) has been appointed to a Lectureship at Singapore Polytechnic.

Mr. R. D. Laurence (Graduate 1964) has been appointed Senior Tutor at the Post Office Telecommunications Management College.

Mr. A. Ludditt (Graduate 1971) is now with the British Steel Corporation as a Designer in the Electrical Design Department of the Special Steels Division, Rotherham.

Mr. S. F. Monk (Graduate 1971) who was with Marconi Radar Systems is now with Stibbe Electronics as an Electronics Test Engineer.

Mr. M. W. F. Owen (Graduate 1970) has joined Rank Precision Industries as a Technical Training Instructor in the Analytical Division at Margate.

Mr. B. H. Quah (Graduate 1970) is with Texas Instruments Malaysia working as a Repair and Maintenance Engineer.

Mr. A. Smith (Graduate 1969) who has been with Panavia A/c GmbH since 1971 as a Quality Engineer, has returned to the British Aircraft Corporation, Preston, to work in the same capacity.

Sqn. Ldr. A. A. P. Thorpe (Graduate 1968) who has been attached to the Royal Malaysian Air Force for the past two years, has returned to the United Kingdom and has been posted to 1 ACC, RAF Wattisham.

Mr. M. L. Walsh, B.Sc. (Graduate 1973) has joined the Radio and Television Division of Decca Ltd. at Bridgnorth, as a Development Engineer. He was previously with GEC (Radio & Television) Ltd. in South Wales.

Mr. R. I. Watson (Graduate 1971) who joined the Post Office in 1965 and was promoted to Assistant Executive Engineer in 1970, has now joined the General Electric Company, Coventry, as Senior Design Engineer.

Mr. O. Hlaing, B.Sc.(Eng.) (Associate 1970) is now working as an Exchange Engineer with the Rangoon Telephone System.

Mr. M. Jamil (Associate 1971) is now with Flight Inspection Group of Saudi Arabian Airlines, in Jeddah.

Mr. T. L. B. Scott (Associate 1961), formerly Marketing Manager with Motorola, has been appointed Sales Manager of Systron-Donner Ltd., Leamington Spa.

Obituary

The Council of the Institution has learned with regret of the deaths of the following members.

Noel Arthur Lockley (Member 1951, Associate 1950, Graduate 1949) died on 27th October 1973 aged 50 years. He leaves a widow and three sons.

Born in Bridgnorth, Shropshire, Mr. Lockley spent the whole of his professional life in that district. He obtained his technical education through attendance at the Wolverhampton and Staffordshire Technical College from 1941 to 1945 whilst working as a Laboratory Assistant with the Radio Gramophone Development Company, with whom he continued under its various ownerships, initially AT & E, then Plessey and for the past five years the Decca Group. During the 1950s he was closely involved in the design and engineering of radio equipment for telephony links, especially for operation in remote areas where power supply economy is vital.

By 1968 Mr. Lockley had risen to the position of a Senior Engineer and in 1971 he was appointed Chief Engineer, Technical Services, of the Decca Radio and Television Spares and Services Division on the merger and transfer to Bridgnorth of the previously separate groups. He took a leading part in setting up the new Division and earned the respect and friendship of many companies throughout the industry during his sadly short tenure of this post.

For many years Mr. Lockley was a member of the West Midland Section Committee and he always made a point of encouraging qualified engineers to seek Institution membership.

Michael Stillitano (Graduate 1970) died on 2nd August 1973, at the age of 30 years. He leaves a widow and young son.

Mr. Stillitano joined the RAF in 1959 and completed his apprenticeship at Halton in 1962. He was promoted to Sergeant in 1967 and served the last four years of his Service career at RAF Leconfield working as a team leader on airborne radio radar and instruments. He obtained his H.N.C. in 1969 with two Distinctions followed by three endorsements, all at Distinction level, in 1970. After completion of his service in 1972 he undertook a one-year Graduate course at the Teachers' Training College in Hull and had accepted an appointment to teach at a local Comprehensive School. A few days after his fatal heart attack news was received by his widow that Mr. Stillitano had fully qualified to teach Mathematics, Physics and Electronics.

Forthcoming Conferences

Signal Processing in Feedback Control Systems

The Automation and Control Systems Group Committee of the IERE is organizing a symposium on the above topic to be held in London in April 1974. It is intended that the symposium will appeal to system designers, system and component manufacturers and testers, as well as to system users. Papers are invited covering such topics as filter specification, analogue filtering, digital filtering, hybrid compensation, tunable filters, adaptive signal processing, including application in computer-controlled systems. Abstracts, which should be sent to Mr. R. C. Slater, Deputy Secretary, IERE, 9 Bedford Square, London WC1B 3RG, as soon as possible, will be reviewed by the symposium committee and authors notified by 1st January 1974. Equipment demonstrations as an integral part of, or sequel to, lecture presentations will be welcomed.

A symposium digest will be provided for all registered attendees, and a report on the symposium will appear in *The Radio and Electronic Engineer*. Suitable papers will also be considered for publication by the Institution in full. Programme details will be announced a few months before the symposium.

The Engineer in Society

Organized by the Institute of Electrical and Electronics Engineers, Region 8, and the Convention of National Societies of Electrical Engineers of Western Europe, Eurocon '74, will have as its theme 'The Engineer in Society'. The conference is being held in Amsterdam from 22nd–26th April 1974 and the organizers are the Royal Institution of Engineers in the Netherlands (KIvI) and the Benelux Section of IEEE.

The conference will:

pay special attention to the social relevance of engineering activities;

provide a forum for the announcement of important new advances in research, technology, development, design and manufacture;

complement existing specialist conferences by providing facilities for discussions on the state-of-the-art, as well as on trends and interactions of technologies;

provide an environment conducive to informal discussion between students, engineers, scientists and technical management.

Papers are invited on the following topics:

Controlling the Future: Energy resources, transport and storage of energy, new power conversion systems for electricity production, the importance of environmental and resource restrictions, increased use of electricity and its consequences.

Instrumentation electronics: Measuring techniques serving the extension of the border of human knowledge, new techniques in instrumentation for environmental control, the impact of microelectronics on instrumentation, new electronic techniques in testing, check-out and control systems.

Communications for the 1980s: the local cable network, long-distance transmission, terrestrial and satellite radio communication, spectrum engineering, electromagnetic compatibility, mobile communications, processors in communications, new services and new technologies.

The computer in society: computers and engineering, computers and education, computers and management, data banks, computers and leisure, large systems and computers (modelling and control).

Biomedical engineering: electronic aids for the handicapped and revalidation, the bed-side computer, biomedical instrumentation and telemetry, electronic image processing; models, electrophysiology and electrostimulation, pattern recognition, processing of bioelectric signals, ultrasonics.

Education: contributed papers and panel discussion on the theme 'Is our engineering education adequate for the future' (viewpoints of professors, students, industry, third world).

The closing date for receiving offers of papers has been extended to 15th December. Offers, accompanied by a 300–500 word abstract appropriate to a 20-minute presentation, should be sent to the Organizer, Eurocon '74, c/o KIvI, 23 Prinsessegracht, The Hague, The Netherlands, from whom all the particulars may be obtained.

Evaluation of the Action of the Space Environment on Materials

The Centre National d'Etudes Spatiales and the Centre d'Etudes et de Recherches de Toulouse are organizing, with the collaboration of the European Space Research Organization (ESRO), an international Conference on the above theme, to be held from 17th to 21st June, in Toulouse (France).

The aim of this Conference is to promote the exchange of information on how materials are affected by the space environment, with particular emphasis how such effects can best be simulated and studied.

The following subjects will be discussed:

Ageing phenomena in simulated space environments

The validity of various simulation techniques

Methods of material testing:

measurements in situ

synergistic effects of particulate and/or electromagnetic radiation

comparison of test results obtained by different methods (standardization of procedures)

Recent space environmental data

Orbital and interplanetary experiments on materials

An introductory plenary session will deal with methods of assessing the environmental conditions and emphasizing radiation fluxes experienced by materials in various parts of spacecraft, taking into consideration spacecraft geometry and mission requirements. The conference will then split into two parallel sessions: the first part will deal with all non-semiconducting materials including coatings, optical materials and structural materials, which are sensitive to the space environment, the second will deal with semiconducting materials and components including solar cells. The selection of papers will be by a Scientific Committee, the Secretary of which is M. H. Arciszewski (Member) of CNES and its UK members are Messrs H. J. H. Sketch and F. C. Treble of RAE.

Requests for further information and offers of papers should be sent to Centre National d'Etudes Spatiales, Départment des Affaires Universitaires, 129 rue de l'Université, 75327 PARIS CEDEX 07.

Fourth European Microwave Conference and Microwave 74

The Management Committee of the European Microwave Conference and the Directors of Microwave Exhibitions and Publishers Ltd. have agreed to combine their respective resources of the organization and sponsorship of next year's Microwave Convention, which is to be held in Switzerland. The combined four-day Conference and Exhibition, to be called the Fourth European Microwave Conference and Microwave 74, will take place in Montreux from 10th–13th September 1974, at the Maison des Congres. The exhibition centre has space for 185 stands, while the two conference halls each have a 500-seat capacity.

The Fourth European Microwave Conference and Microwave 74 will enjoy the full co-operation of the Convention of the National Societies of Electrical Engineers of Western Europe. The combined activity will also be supported by the sponsors of the European Microwave series of conferences, which include the IEE (Electronics Division), the IEEE (Region 8 and three of its relevant professional groups or societies), the Swiss National Committees of URSI and the IERE, which was a sponsor of Microwave 73.

The conference is to be chaired by Professor Frederic Gardiol, Professor of the Department of Electricity at the École Polytechnique in Lausanne, and will have as its main theme both microwave research and applications. The scope of the conference will therefore embrace both fundamental principles and hardware developments, which will be covered in parallel sessions.

Further information about the Conference and Exhibition may be obtained from Microwave Exhibitions and Publishers Ltd., 21 Victoria Road, Surbiton, Surrey.

Advances in Airborne Equipment for Navigation and Flight Control

The present rate of advance in electronics technology, not least in the field of integrated circuits, affects the interplay between the performance, integrity, reliability, logistics and cost of the equipment carried in aircraft for flight control and navigation. A joint IERE–Royal Institute of Navigation meeting is being planned for November 1974 at which it is intended to provide a review of recent advances and describe trends in airborne equipment design.

IERE members are invited to contribute papers to this meeting, the object of which is to present to navigators and others on the operational side of aviation a comprehensive review of present trends in equipment design; the emphasis is on equipment of the present and the near future, rather than on broader issues such as (for example) future satellite or other navigational systems.

Offers of papers should preferably be accompanied by a synopsis and should be sent to the Secretary, Aerospace, Maritime and Military Systems Group Committee, IERE, 9 Bedford Square, London WC1B 3RG.

Underwater Communication Systems

Improved Ship to Submarine Sonar Telephone

Sea trials have been completed on a new ship-to-submarine underwater telephone system, being developed by Marconi Communication Systems Limited, to give the Royal Navy an improved underwater voice communication system. Better transmission quality at longer ranges was the primary aim defined by the specification. The equipment will eventually replace current underwater telephones in all Royal Navy ships and submarines.

Based on the building block principle for maximum installation flexibility, the equipment consists of four basic electrical versions, ranging from a simple low-power transmit/receive system with a single transducer, to a high-power system with directional capability and several transducers. Application of the latest solid-state techniques has ensured higher efficiency and greater reliability.

The system is engineered to withstand the most extreme environmental conditions and a further requirement of the specification was that the system should operate with all existing transducers and on the standard NATO carrier frequency employing a single-sideband suppressed carrier mode.

Deep Sea Helium Speech Processors for Divers

The United States Navy is buying British systems capable of overcoming the deadly 'Donald Duck' effect which oxy-helium has on deep-sea diver's speech. The systems, worth, with spares, a total of £23,000, were developed for the Royal Navy by Marconi Space and Defence Systems Limited, from Admiralty Research Laboratories designs.

The 'Donald Duck' effect results from divers having to breathe an oxy-helium mixture in depths of greater than 600 ft, where air cannot be used safely. The mixture, being much less dense than air, produces changes in the speed of sound, and therefore in the pitch of a speaker's voice. This rises to an extent where it becomes completely unintelligible to the listener.

The Marconi system, designated the Type 023, was developed from ARL designs started in late 1968.* It has already seen service in the Admiralty Experimental Diving Unit and the Royal Naval Physiological Laboratory, and is currently being evaluated, with favourable results, in a series of medical research dives of up to 1000 ft by the Smithsonian Institute in the USA. It operates on a 'time stretching' principle, where each sound is digitally analysed, and the significant portion, typically about one third, is reconstructed at a slower rate, while the rest is rejected. This has the effect of lowering the frequency to about a third of its transmitted value, and thus creating full intelligibility.

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* Gill, J. S., Morris, R. J. and Edwards, M. G., 'A helium speech processor operating in the time domain', Conference on Electronic Engineering in Ocean Technology, Swansea, 1970. IERE Conference Proceedings No. 19, pp. 523–8.

INSTITUTION OF ELECTRONIC AND RADIO ENGINEERS

Applicants for Election and Transfer

The Membership Committee at its meetings on 3rd, 16th and 24th October 1973 recommended to the Council the election and transfer of 117 candidates to Corporate Membership of the Institution and the election and transfer of 15 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: 3rd October 1973 (Membership Approval List No. 166)

GREAT BRITAIN AND IRELAND

CORPORATE MEMBERS

Transfer from Member to Fellow

BROWN, John Millard, Squadron Leader. *Filey, Yorkshire.*
PARKS, Raymond, B.Sc., Ph.D. *Edinburgh.*
TUCKER, Kenneth William, Lieutenant Colonel. *Halstead, Kent.*

Direct Election to Fellow

ANSTEAD, John William. *Prestbury, Gloucester.*

Transfer from Graduate to Member

AHLUWALIA, Daljit Singh. *Allestree, Derby.*
BETTS, David Charles. *Burgess Hill, Sussex.*
BRETT, Donald Austin David. *Bridgend, Glamorgan.*
CRAPP, Alan Roy. *Kinghorn, Fife.*
FISHENDEN, Keith David, M.Sc. *Cwmbran, Monmouthshire.*
FROGLEY, Ian. *Buckhurst Hill, Essex.*
FURNISS, Michael John. *Littleover, Derby.*
GRAHAM, Christopher Bentley. *Birkenhead, Cheshire.*
GREEN, Anthony John. *Chorleywood, Hertfordshire.*
HALE, David Patmore, B.Sc. *Tilehurst, Berkshire.*
HORTON, Michael John. *Horndean, Hampshire.*
HUNT, Anthony Martin. *London, W.3.*
LENNON, Hugh Patrick. *London, S.W.6.*
MOWBRAY, Kenneth John. *Bracknell, Berkshire.*
OLDHAM, Eamonn John. *London, N.W.7.*
PALMER, Alan. *Rushden, Northants.*
REDDISH, Colin Sydney. *Thatcham, Berkshire.*
SCOTLAND, John Archibald. *Evesham, Worcestershire.*
SILVERSIDE, Walter Norman. *Kingston, Surrey.*
SODEN, Brian Oakley. *Send, Surrey.*
TAYLOR, John Chisholm. *Garforth, Leeds.*
WILLIAMS, William Fraser. *Horley, Surrey.*
YATES, David Claydon. *Evesham, Worcestershire.*

Direct Election to Member

PARROTT, John Haden. *Desford, Leicester.*
SHORROCK, Michael Alan. *Blackpool, Lancashire.*
SMITH, Alan. *Frimley Green, Surrey.*
WILKINSON, Anthony John, B.Sc. *Stevenage, Hertfordshire.*

NON-CORPORATE MEMBERS

Transfer from Student to Graduate

HENRY, William Michael Joseph. *Dublin 1, Eire.*
KELLY, Stephen William. *Chatham, Kent.*
STEED, Julian James. *Dover, Kent.*

Direct Election to Graduate

CABLE, Alan James, B.Sc. *Lowestoft, Suffolk.*

OVERSEAS

CORPORATE MEMBERS

Transfer from Graduate to Member

COLEMAN, Martin Reginald. *Reseda, California, U.S.A.*
COUSINS, Donald John. *Rockford, Illinois, U.S.A.*
MORTIMORE, Patrick Kenneth. *Durban, Natal, South Africa.*
YOUNG, Clarence Robert. *Auckland, New Zealand.*

Direct Election to Member

SOTUNDE, Jacob Sotayo. *Ikoyi, Lagos, Nigeria.*

Meeting: 16th October 1973 (Membership Approval List No. 167)

GREAT BRITAIN AND IRELAND

CORPORATE MEMBERS

Transfer from Member to Fellow

HOPKINS, William Bernard. *High Wycombe, Buckinghamshire.*

Direct Election to Fellow

ALLCOCK, Philip Robert. *Stafford.*
BURVILL, Ronald Frederick, Captain, R.N. *Camberley, Surrey.*

Transfer from Graduate to Member

CHADWICK, Derek. *Stretford, Manchester.*
CLARK, Roger William, Flight Lieutenant. *Medmenham, Buckinghamshire.*
CLUES, Beverley John. *Colchester, Essex.*
DAY, Kenneth John Coryton, Lieutenant, R.N. *Gosport, Hampshire.*
DUNTON, Keith. *Currie, Midlothian.*
ELLIS, William Henry. *Walsall Wood, Staffordshire.*
EVANS, Roger David. *Earley, Berkshire.*
FISHER, Granville. *Dalkeith, Midlothian.*
FLEMING, Alexander Mathieson. *Stevenage, Hertfordshire.*
HODGSON, Peter. *Hampton, Middlesex.*
HORTON, Raymond Stanley. *Ellesmere, Shropshire.*
JOHL, Dharm Singh. *Wallington, Surrey.*
KEEL, Peter Russell Godfrey. *Andover, Hampshire.*
KELLY, Joseph Brian. *Wigan. Lancashire.*
McGUIRE, Bernard Edward. *Wolverton, Buckinghamshire.*
MORRISON, James Ekron Little. *Penicuik, Midlothian.*
SPENCER, Anthony. *Birchwood, Lincoln.*
SUMMERHAYES, Peter Frank. *Henfield, Sussex.*
UPPINGTON, Kenneth Ralph. *Lochwinnoch, Renfrewshire.*
WILLIAMS, Kenneth. *Carluke, Lanarkshire.*

Transfer from Associate to Member

HOPKINS, David Louis. *Little Baddow, Essex.*

Direct Election to Member

FAIRCLIFFE, James Delanoy Saberton. *Burwell, Cambriage.*
FLOWER, Christopher William. *Feltham, Middlesex.*
JENKINS, Peter, B.Sc. *Alderley Edge, Cheshire.*
LEVY, Brian Martin. *Croxley Green, Hertfordshire.*
LEWCOCK, John, Lieutenant Commander. *Bordon, Hampshire.*
MacDONALD, Norman Stephen, B.Sc. *Barrhead, Glasgow.*
PORTER, Cyril Arthur. *Kenton, Middlesex.*

NON-CORPORATE MEMBERS

Transfer from Student to Graduate

MUDHAR, Narinder Singh, B.Sc. *Wembley, Middlesex.*

Direct Election to Graduate

KUMAR, Krishna, B.Tech. *Bradford, Yorkshire.*
MANSOOR, Gabriel. *London, E.5.*

Direct Election to Associate

ALPE, John Michael Elvin. *London, S.E.23.*

STUDENTS REGISTERED

COLES, Simon John. *Warminster, Wiltshire.*
CULLEN, Michael John. *Banstead, Surrey.*
McGUINNESS, Patrick Gerard. *Carshalton, Surrey.*

OVERSEAS

CORPORATE MEMBERS

Transfer from Graduate to Member

KFIR, Moshe. *Nathanya, Israel.*

Direct Election to Member

BRETTON, Kenneth Martin. *Burbank, California, U.S.A.*
LAMB, Charles Andrew, Flight Lieutenant. *Butterworth, Penang, Malaysia.*
WOYTHAL, John Michael, B.Sc. *Maryland Heights, Missouri, U.S.A.*

NON-CORPORATE MEMBERS

Direct Election to Graduate

MONDAL, Kalyan, B.Tech. *Calcutta 31, West Bengal, India.*
NNAMA, Emmanuel Chukwuma, B.Sc.(Eng). *Lagos, Nigeria.*

STUDENTS REGISTERED

LAU, Yeuk Shan. *North Point, Hong Kong.*
LIM, Kok Kwang. *Singapore 9*
OH, Ai Leng Miss. *Singapore 12.*
PHUA, Hoar Meng. *Singapore 26.*
POEN, Wan Loe. *Kowloon, Hong Kong.*
WAN, Kim Ping. *Kowloon, Hong Kong*
YOUNG, Wai Kwan. *Wanchai, Hong Kong.*

Meeting: 24th October 1970 (Membership Approval List No. 168)

GREAT BRITAIN AND IRELAND

CORPORATE MEMBERS

Transfer from Graduate to Member

BAILEY, Alan John. *Goring, Sussex.*
BARRETT, John Flannigan. *Bathgate, West Lothian.*
BENSON, Francis William. *High Wycombe, Buckinghamshire.*
BOCKING, Brian John, B.Sc., M.Sc. *Frimley, Surrey.*
BONNEY, Trevor James. *Harlow, Essex.*
BREINGAN, William John Malcolm. *North Harrow, Middlesex.*
BROWN, Frederick Richard Lloyd. *Wolverhampton, Staffordshire.*
CARTER, Leslie William. *Little Sutton, Cheshire.*
CHESHIRE, Anthony Ernest. *Solihull, Warwickshire.*
COLE, Royston Douglas. *Welland, Worcestershire.*
COOKE, Peter Alfred. *Steeple Claydon. Buckinghamshire.*
DAVIDSON, Robert Samuel. *Royston, Hertfordshire.*
DAVIES, David John. *Hayes, Middlesex.*
DODGSON, Peter. *Chalfont St. Giles, Buckinghamshire.*
DOWDEN, Michael John. *Ampthill, Bedfordshire.*
DRAYCOTT, Michael John. *Hitchin, Hertfordshire.*
DUTHIE, Ian Fraser. *Eaglesham, Glasgow.*
EDWARDS, Robert Ernest. *Chapelfields, Coventry.*
EKE, Kenneth Ian. *Sanderstead, Surrey.*
FILLION, Jacques Douglas. *Crawley, Sussex.*
GENENDER, Alan Jack. *Farnborough, Hampshire.*
GILTROW, Michael John. *Bishop's Stortford Hertfordshire.*
GREEN, Duncan Graham. *Wells, Somerset.*
HAMPSON, Christopher Gerald. *Wallasey, Cheshire.*
HORAN, David George. *Morden, Surrey.*
HOWES, John Frederick. *Hildenborough, Kent.*
INSKIP, David John. *Huncote, Leicester.*
ISTED, Alan John. *Lymm, Cheshire.*
JONES, Maurice Richard. *Biggleswade, Bedfordshire.*
KONDROLLOCHIS, Michael. *Woodley, Berkshire.*
PAYNE, Alan Norman. *Draycott, Somerset.*
KEBLE, John Mervyn. *Wembley Park, Middlesex.*
SHIPLEY, John Terence, B.Sc. *Hazlemere, Buckinghamshire.*
TERRY, John Beevor. *East Grinstead, Sussex.*
WESSON, Anthony Ernest. *Winford, Isle of Wight.*
WILKINSON, Harry. *Chinley, Cheshire.*
WILLARD, Gerald Arthur. *Carshalton, Surrey.*
WILLIAMS, Alan. *Brentwood, Essex.*

Direct Election to Member

BLETCHER, Michael Ian. *Stoneleigh, Surrey.*
BREEZE, Michael Ernest. *Woodley, Berkshire.*
CLARE, Colin Peter, B.Sc. *Iver Heath, Buckinghamshire.*

PUNTIS, John Arthur, B.Sc.(Eng). *Southampton, Hampshire.*

NON-CORPORATE MEMBERS

Transfer from Student to Graduate

SIMPSON, Malcolm Laurence, B.Sc. *South Harrow, Middlesex.*

Direct Election to Graduate

ELLIOTT, Donald. *Portsmouth, Hampshire.*
FARRAR, David Samuel. *Blackrock, County Dublin.*
WHELAN, David Lawrence. *Ipswich, Suffolk.*

Direct Election to Associate

HENTHORNE, Thomas Clifford. *Shevington, Lancashire.*

STUDENTS REGISTERED

BACON, Richard Ernest. *Bracknell, Berkshire.*

OVERSEAS

CORPORATE MEMBERS

Transfer from Graduate to Member

ARULIAH, Christy Gunapalan. *Dhahran, Saudi Arabia.*
BROOKES, David. *St. Peters, Missouri, U.S.A.*
JONES, Lynn Williams. *Tehran, Iran.*

Direct Election to Member

ARORA, Narain Das. *New Delhi 27, India.*

STUDENTS REGISTERED

SHA, Dilip Kumar. *Dist. Dacca. Bangladesh.*
SIT, Ping Keung. *Kowloon, Hong Kong.*
THAM, Chin Kim. *Singapore 19.*

Notice is hereby given that the elections and transfers shown on Lists 162, 163, 164 and 165 have now been confirmed by the Council.

STANDARD FREQUENCY TRANSMISSIONS— September 1973

(Communication from the National Physical Laboratory)

Sept. 1973	Deviation from nominal frequency in parts in 10^{10} (24-hour mean centred on 0300 UT)			Relative phase readings in microseconds NPL—Station (Readings at 1500 UT)	
	GBR 16 kHz	MSF 60 kHz	Droitwich 200 kHz	*GBR 16 kHz	†MSF 60 kHz
1	0	0	0	699	610·6
2	+0·1	+0·1	0	698	610·1
3	0	0	0	698	609·7
4	+0·1	+0·1	0	697	609·0
5	+0·1	+0·1	+0·1	696	608·3
6	+0·1	+0·1	0	695	607·2
7	+0·1	+0·1	+0·1	694	606·6
8	+0·1	+0·1	+0·1	693	605·7
9	+0·1	+0·1	+0·1	692	604·9
10	0	+0·1	0	692	604·3
11	0	0	+0·1	692	605·2
12	0	0	+0·1	692	603·9
13	0	0	0	692	603·9
14	0	+0·1	0	692	603.3
15	0	0	0	692	603·7
16	0	0	0	692	604·1
17	0	0	0	692	603·9
18	0	+0·1	0	692	603·3
19	0	0	−0·1	692	603·1
20	+0·1	+0·1	0	691	602·4
21	+0·1	+0.1	0	690	601·7
22	+0·1	+0·1	−0·1	689	601·1
23	0	+0·1	0	689	600·3
24	+0·2	+0·1	0	687	599·4
25	+0·1	+0·1	0	686	598·7
26	−0·1	0	0	687	598·7
27	0	0	0	687	598·3
28	0	0	0	687	598·3
29	0	0	−0·1	687	598·7
30	0	0	0	687	598·3
31					

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

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

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

Forthcoming Institution Meetings

London Meetings

Wednesday, 5th December

AEROSPACE, MARITIME AND MILITARY SYSTEMS GROUP

Use of Split P.P.I. Techniques in Clutter and other Investigations

By P. D. L. Williams (*Decca Radar*)

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

The accurate comparison between two radars having only one or two chosen differences in parameters is often upset by having to view two independent displays driven from independents scanners. Croney and White have demonstrated split image displays, but the two pictures are separated in time by half an aerial rotation period.

A selection of p.p.i. photographs are to be presented delayed only by a progressive 1 millisecond delay. The chosen parameters investigated are wavelength, horizontal aerial aperture and receiver type (log or linear).

The coastal siting of the equipment enabled land and sea targets to be examined in the presence of varying amounts of sea and rain clutter as well as thermal noise.

Thursday, 6th December

EDUCATION AND TRAINING GROUP

TEC, ERB and the Technician Engineer

By A. J. Kenward (*SERT*)
London School of Hygiene and Tropical Medicine, 6 p.m. (Tea 5.30 p.m.)

The identification, definition, education and training of Technician Engineers and Technicians have been important topics for debate over the last few years. The success of the work of the Engineers' Registration Board in forming the composite register and producing and obtaining acceptance of common standards for registration among forty autonomous engineering bodies has helped to clarify ideas on identification and definition.

The education, and to some extent the training, of these two categories of Technicians and Technician Engineers is now under consideration in depth by the Technician Education Council. The first consultative report of the Council will be published immediately before this meeting and will provide an opportunity for discussion of the aims and objectives and of proposals for technician education.

Wednesday, 12th December

COMPONENTS AND CIRCUITS GROUP

Colloquium on IMPACT OF MICRO-ELECTRONICS ON INSTRUMENT DESIGN

IERE Lecture Room, 2 p.m. Please note change of time.
Further details and registration forms from Meetings Secretary, IERE.
The application of hybrid microelectronics to instruments
By R. Jones (*Hewlett Packard*)

The influence of modern storage devices on frequency response analyser design
By Dr. J. Izatt (*Solartron*)

The argument for o.e.m. in-house design of l.s.i microcircuits
By J. G. L. Rhodes (*Pye TMC*)

The use of integrated circuits in quantization distortion measurements—a case history.
By D. Waddington (*Marconi Instruments*)

Design of a modern counter timer
By N. L. Ayres (*AMF Venner*)

Wednesday, 9th January

COMMUNICATIONS GROUP

Adaptive Delta Modulation Systems

By R. Steele (*Loughborough University of Technology*)

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

After describing the linear delta modulator and its limitations, the lecture will review the various types of adaptive delta modulation systems which have been conceived to overcome some of these limitations.

These adaptive systems can be divided into two basic groups, syllabically companded and instantaneously companded systems. The former are suitable for speech transmission while the latter have particular characteristics which make them suitable for television encoding. Many different adaptive systems have been conceived and an attempt will be made to show underlying similarities between them. Lastly, the relationship between delta modulation systems and p.c.m. systems will be made for both linear and companded systems.

Wednesday, 16th January

AEROSPACE, MARITIME AND MILITARY SYSTEMS GROUP

Colloquium on DOPPLER VOR SYSTEMS

IERE Lecture Room, 2.30 p.m.

Further details and registration forms from Meetings Secretary, IERE.

Wednesday, 23rd January

COMPONENTS AND CIRCUITS GROUP

Colloquium on RECENT DEVELOPMENTS IN TURNTABLE DESIGN

IERE Lecture Room, 2.30 p.m.

Further details and registration forms from Meetings Secretary, IERE

Thursday, 24th January

EDUCATION AND TRAINING GROUP

Training the Euro-Engineer

By F. R. J. Langridge (*Engineering Employers Federation*)

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

Monday, 28th January

JOINT IEE/IERE COMPUTER GROUP

Designing Machines for People (or Human Factors in Computer Design)

By Dr. C. R. Evans (NPL)

IEE, Savoy Place, W.C.2, 5.30 p.m. (Tea 5 p.m.).

Wednesday, 30th January

COMMUNICATIONS GROUP

Colloquium on ACTIVE AERIALS

IERE Lecture Room, 2.30 p.m.

Further details and registration forms from Meetings Secretary, IERE

Kent Section

Wednesday, 5th December

Electronic Systems for the Space Environment

By A. J. Price (*Marconi Space and Defence Systems*)

Medway and Maidstone College of Technology, Chatham, Kent, 7 p.m.

The paper considers the influence of the space environment on the design of electronic circuits. Specific systems are described and their operation in weather satellites and sounding rockets.

Wednesday, 30th January

Medical Electronics

By W. J. Perkins (*National Institute for Medical Research*)

Lecture Theatre 18, Medway and Maidstone College of Technology, Chatham, Kent, 7 p.m.

Thames Valley Section

Thursday 24th January

New Sources of Power

MEETING CANCELLED

East Anglian Section

Wednesday, 9th January

JOINT MEETING WITH IEE

Tomorrow's World in Radio Communications

By T. R. Rowbotham (*Post Office*)

King Edward VI Grammar School, Broomfield Road, Chelmsford, 6.30 p.m. (Tea 6 p.m.).

Thursday, 24th January

JOINT MEETING WITH IEE

Planning for UHF Television Coverage

By A. L. Witham (*IBA*)

University Engineering Laboratories, Trumpington Street, Cambridge, 6.30 p.m. (Tea 6 p.m.).

Southern Section

Tuesday, 4th December

Future Telecommunications Projects in Space

By W. M. Lovell (*Marconi Space and Defence Systems*)

Brighton Technical College, 6.30 p.m. (Refreshments available from 5.45 p.m. in Refectory)

Wednesday, 5th December

Inertial Navigation

By G. U. Rands (*Marconi-Elliott Avionic Systems*)

H.M.S. *Daedalus*, 6.30 p.m.

The ability to determine the present position of an aircraft to an accuracy of 1 mile or better without using any ground aids or airborne radar is provided by inertial navigation techniques. The lecture deals with principles, techniques and hardware, and briefly traces the history of inertial navigation, describes present day systems and takes a look into the future.

Wednesday, 12th December

Stored Program Control of Telephone Exchanges

By B. L. Nuttal (*Post Office*)

University of Southampton, 6.30 p.m. (Tea served in Senior Common Room from 5.45 p.m.)

Within the last few decades we have seen enormous advances in the technologies available for logic and circuit design. In particular the advent of transistors and integrated circuits has enabled computers to be manufactured that are larger, faster and more powerful than could ever have been conceived 30 years ago. These computers have had many applications within industry but it is only comparatively recently that the use of a computer in the control of a telephone exchange has been considered. Nowadays, many telephone administrations are turning to stored program control of exchanges as a possible alternative to existing equipment. However, the introduction of stored program control poses probably as many new problems as it solves old ones so is this faith in computers for telephone purposes justified?

Wednesday, 30th January

JOINT MEETING WITH IEE

Colloquium on COMPACT MOBILE TRANSMITTERS—IMPROVEMENTS FOR HIGH STRESS ENVIRONMENTS

The Cinema, Plessey Company, West Leigh, Havant, 4 p.m.

Thursday, 31st January

Meteorological Telecommunications Systems Engineering

By C. E. Goodison (*Meteorological Office*)

Farnborough Technical College, 7 p.m. (Refreshments available in College Refectory from 6.30 p.m.).

Meteorological telecommunications systems consist of different scales of size and sophistication ranging from world encompassing systems which are computer controlled to simple 50-baud teleprinter links. Following the pattern of present day developing communication systems the present picture of meteorological telecommunications is one of rapid change. The lecture will describe some of these systems and the subjects will include the Global Telecommunication System of the World Weather Watch, its Main Trunk Circuit and particularly the Regional Telecommunication Hub at Bracknell.

Yorkshire Section

Tuesday, 4th December

JOINT MEETING WITH IEE

Faraday Lecture

By Dr. A. J. Churchman
City Hall, Sheffield, 7 p.m.

Thursday, 17th January

JOINT MEETING WITH IEE

The Problems of Radio Systems above 10 GHz

By Dr. P. A. Watson (*University of Bradford*)

University of Leeds, 7 p.m. (Refreshments 6.30 p.m.).

The effects of rainfall, multipath propagation and other lower atmospheric phenomena on the propagation of radio waves above 10 GHz will be discussed and related to the design of systems at these frequencies. Some possible future satellite and terrestrial radio relay systems using these frequencies will be described.

East Midland Section

Tuesday, 4th December

The Impact of Advances in Electronics in Electrical Heating Processes

By J. E. Harry (*Loughborough University*)

Edward Herbert Building, Loughborough University of Technology, 7 p.m. (Tea 6.30 p.m.)

Recent advances in electronics are finding extensive applications in electroheat. Examples of their application are for temperature and power control and in the generation of medium and high frequency power. Some of these developments will be described together with their applications.

Thursday, 17th January

Interfacing Strategy on Real Time Computer Systems with particular reference to CAMAC

By H. Bisby (*Atomic Energy Research Establishment, Harwell*)

Lecture Theatre 'A', Physics Block, Leicester University, 7 p.m. (Tea 6.30 p.m.).

CAMAC is a definitive style for implementing the interface conditions which exist when many channels of input/output information share a common data-processor/controller. The features of CAMAC are described to indicate the applicability of CAMAC-compatible equipment and programming to real-time situations. Some of these are illustrated by typical systems which use either a computer or a simpler device as the central processor/controller.

West Midland Section

Thursday, 24th January

JOINT MEETING WITH IEE

Electronic Ignition Systems

By R. L. Rivers (*Mobelec*)

Lanchester Polytechnic, Coventry, 7 p.m.

South Midland Section

Monday, 14th January

JOINT MEETING WITH IEE

The Use of Radar in Meteorology

By Dr. R. Starr (*RRE, Malvern*)

Winter Gardens, Malvern, 7.30 p.m.

South Western Section

Tuesday, 4th December

JOINT MEETING WITH IEE

Space Technology and the Future

By G. K. C. Pardoe (*General Technology Systems Ltd*)

The College, Swindon, 6.15 p.m. (Tea 5.30 p.m.)

The lecture will establish the current situation in space technology and examine how this will evolve in both short term and long term (to year 2000) periods. Technical and cost benefit aspects will be discussed, so too will be the fundamental problem of how to organize space projects.

Wednesday, 5th December

Liquid Crystals

By G. Elliott (*Marconi*)

No. 4 Lecture Theatre, School of Chemistry, University of Bristol, 7 p.m. (Tea 6.45 p.m.)

The speaker will discuss the following: the nature of liquid crystals and their special and optical properties; the behaviour of thin layers and their response to electrical fields; the construction of an electro-optic device and examples of electro-optic effects; and their development and applications.

Wednesday, 23rd January

New Electronic Components using Amorphous Semiconductors

By Dr. J. Allison (*University of Sheffield*)

Lecture Room 4E3.10, University of Bath, 7 p.m. (Tea 6.45 p.m.).

Thursday, 31st January

JOINT MEETING WITH IEE

High Fidelity Sound Reproduction

By R. L. West (*Polytechnic of North London*)

Main Hall, Plymouth Polytechnic, 7 p.m. (Tea 6.45 p.m.).

The lecture and demonstrations will be largely historical in nature, not only tracing the gradual improvement in sound quality over the years, but also showing how and why these developments came about. The lecture will cover the whole field of reproduced sound and highlight the most important landmarks on this journey.

North Eastern Section

Wednesday, 12th December

Computer Controlled Telephone Exchanges

By Dr. M. T. Hills (*University of Essex*)

Main Lecture Theatre, Ellison Building, Newcastle upon Tyne Polytechnic, Ellison Place, 6 p.m. (Refreshments in Staff Refectory from 5 30 p.m.)

Wednesday, 19th December

JOINT MEETING WITH IEE

Colloquium on COMPUTERS IN MARINE AUTOMATION

Henderson Hall, University of Newcastle-upon-Tyne, 10 a.m.

Data highways in ship automation

By L. R. Thompson (*Hawker Siddeley Dynamics*)

Cargo instrumentation for LNG tankers
By I. W. F. Paterson (*Whessoe Systems and Control*)

Application of control techniques to marine problems
By Dr. R. V. Thompson (*Marine Industries Centre*)

Automation and control in the fishing industry
By M. Hatfield

Help monitoring of marine equipment
By M. Langballe (*Det Norske Beritas*)

Use of computers in the evaluation of radar data
By Dr. R. J. Tunnicliffe

Use of computers for automation in marine systems
By Dr. D. Strong (*Ministry of Defence*)

Marine autopilot position fixing system interfacing
By W. H. P. Canner (*UWIST*)

Adaptive series compensation for large 2-axis stable platform
By Lieutenant Commander R. Wilson (*ASWE*)

Wednesday, 9th January

Electronics in Radio Astronomy

By Professor J. G. Davies (*Jodrell Bank*)

Main Lecture Theatre, Ellison Building, Newcastle-upon-Tyne Polytechnic, Ellison Place. 6 p.m. (Refreshments in Staff Refectory from 5.30 p.m.).

North Western Section

Thursday, 13th December

The Application of Electronics in Telephone Exchange Switching

By F. W. Croft (*Post Office Telecommunications*)

Lecture Theatre R/HIO, Renold Building, UMIST, 6.15 p.m. (Tea 5.45 p.m.).

The lecturer will give an outline of the Post Office electronic telephone exchange widely used in public service, and touch briefly on a new system for large electronic exchanges. Also covered will be electronic equipment systems used to steer calls over the electro-mechanically switched network, including reference to Stored Program Control processors.

Thursday, 17th January

The Texan Stereo Amplifier

By R. Mann (*Texas Instruments*)

Lecture Theatre R/HIO, Renold Building, UMIST, 6.15 p.m. (Tea 5.45 p.m.).

Merseyside Section

Wednesday, 12th December

R.F. Sputtering of Thin Films

By E. F. Lever (*Liverpool Polytechnic*)

Department of Electrical Engineering and Electronics, University of Liverpool, 7 p.m. (Tea 6.30 p.m.)

The introduction will outline some applications of dielectric thin films, and will describe the problems of their fabrication. R.f. sputtering will be introduced as a solution to this problem, and the general idea of the process will be described.

The characteristics of the r.f. glow discharge will be described, with emphasis on ion sheath formation and the effects of magnetic fields. Practical sputtering systems will then be considered, with detailed attention to the target system, the substrate, the vacuum plant, and the problem of impedance matching. R.f. power requirements will be discussed, with mention of the relative merits of free-running and fixed frequency operation, and the problems of power losses. Finally, the performance of practical systems will be summarized.

Northern Ireland Section

Saturday, 8th December

Dinner Dance

CANCELLED

Wednesday, 12th December

Wine and Cheese Party

90 Belmont Road, Belfast

Wednesday, 9th January

Please note change of date

Optoelectronics

By Speaker from *Texas Instruments*

Cregagh Technical College, Montgomery Road, Belfast 5, 7 p.m.

South Wales Section

Wednesday, 12th December

JOINT MEETING WITH IEE

Developments in Data Communications

By M. B. Williams (*PO Telecommunications HQ*)

Department of Applied Physics, UWIST, Cardiff, 6.30 p.m.

(Tea in College Refectory from 5.30–6 p.m.)

Most data terminals in the forseeable future are expected to operate at speeds within the capacity of telephone circuits. Development will continue therefore, of modems and other techniques of adapting the telephone network to handle data. Many countries and organizations are studying the prospects for providing new forms of data communication services. These include more versatile switching systems and the evolution of digital networks capable of carrying a range of new services.

Wednesday, 9th January

Electronic Instrumentation for Respiratory Monitoring

By C. L. Smith (*Glamorgan Polytechnic*)
Department of Applied Physics, UWIST, Cardiff, 6.30 p.m.

(Tea in College Refectory from 5.30—6 p.m.)

The lecture will firstly consider flow transducing problems encountered in monitoring respiratory information both in anaesthesiology and lung function applications. A recently developed respiratory transducer, which provides a digital output signal will then be described, together with circuit techniques adopted for evaluating the required flow parameters. The use of a computing technique which evaluates lung function data will also be discussed. Owing to the availability of a digital flow signal the computer can plot graphical information which is not 'smoothed out' as in other transducing techniques.

Scottish Section

JOINT MEETINGS WITH IEE

Electronic Aids for Medical and Biological Studies

By Dr. E. T. Powner (*UMIST*)

Monday, 10th December

Room 406, James Weir Buildings, University of Strathclyde, Glasgow, 6 p.m.

Tuesday, 11th December

South of Scotland Electricity Board Showrooms, 130 George Street, Edinburgh, 6 p.m.

JOINT MEETINGS WITH IEE

Electronics in Ocean Technology

By Dr. V. G. Welsby (*University of Birmingham*)

Monday, 7th January

Room 406, James Weir Buildings, University of Strathclyde, Glasgow, 6 p.m.

Tuesday, 8th January

South of Scotland Electricity Board Showrooms, 130 George Street, Edinburgh, 6 p.m.

JOINT MEETINGS WITH BRITISH COMPUTER SOCIETY (PUBLIC LECTURE)

From Morse Code to Data Networks

By E. B. Stuttard (*Racal Milgo*)

Wednesday, 16th January

Napier College of Science and Technology, Edinburgh, 7 p.m.

Thursday, 17th January

Boyd Orr Building, Glasgow University, 7 p.m.