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Patterns of Research in the Seventies

THE interdependence of pure and applied research was commented upon with some vigour by the President of the Institute of Physics, Sir Brian Flowers, when he spoke at the opening of the Physics Exhibition in London on April 9th. Sir Brian hoped that the Minister for Aerospace and Shipping, Mr. Michael Heseltine, who had just opened the Exhibition, would point out to some of his colleagues that there was not one set of rules for pure physics and another for applied physics. The 57th Exhibition was clearly an appropriate occasion for such an observation because the traditional 'mix' of pure research apparatus and applied research equipment shows this continuity of natural laws. Both the Department of Trade and Industry and the Procurement Executive of the Ministry of Defence mounted extensive and extremely interesting demonstrations at the Exhibition which clearly underlined the point.

The Exhibition also presented the corollary that pure research must receive support even though an immediate field of application cannot clearly be seen. Government-financed research associated with the DTI now applies, or soon will apply, the customer/contractor principle under which the requirement for any particular piece of research has to be defined by the recently established Requirements Boards and similar policies have long guided research for Defence. A fairly large proportion of the research carried out in the Universities comes under the Science Research Council of the Department of Education and Science, which also operates several research laboratories: here one may assume that Sir Brian's tenure of the chairmanship of the SRC for the past six years will have left these organizations well aware of the interdependence of pure and applied research.

Nevertheless, while interdependence may be widely accepted, rigid independence may make the application of research difficult, a point of view expressed recently by Professor Kurt Hoselitz, Director of the Mullard Research Laboratories. Introducing the work of his Laboratories during the past year, he suggested that the ready availability *within one organization* of experts from several disciplines made the application of research easier. Thus, if physicists, electronic engineers, chemists and metallurgists can collaborate readily and closely, the combined expertise has a great advantage over, for instance, that of a university department in which there was a far greater degree of organizational and physical separation from departments concerned with other disciplines. Certain implications of such a comment must react on the concept of a university and the place of research within it.

Most universities appreciate the problems of undertaking technological research which can have a meaning beyond the granting of doctorates and they have their own solutions. The Department of Electronic and Electrical Engineering at the University of Birmingham has greatly developed its postgraduate activities and a tenfold expansion of research in the '60s made its move in 1970 to its new Gisbert Kapp Building imperative. During Open Days held in March to mark the naming of the Building it was apparent that some of the problems of interdependence can be resolved partly by close links with industry and government establishments who sponsor research, and partly by covering a wide range of projects in several research divisions.

Whatever the means by which individual areas within universities, government and industry carry out their research, progress of all projects will benefit by a proper appreciation of the interdependence both of pure and applied research *and* of disciplines.

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Mr. J. D. Parsons (Member 1967) graduated from University College of South Wales, Cardiff, in 1959, and joined the scientific staff of G.E.C. Applied Electronics Laboratories, Stanmore. From 1962 to 1966 he was lecturer and subsequently senior lecturer at The Polytechnic, Regent Street, London, during which time he obtained the degree of M.Sc. in electronics at King's College, London. In 1966 he became Principal Lecturer at South Birmingham Technical College (now City of Birmingham Polytechnic) and since 1969 he has been with the Department of Electronic and Electrical Engineering at the University of Birmingham. His research interests in radio communication are at present directed towards mobile radio diversity reception and a study of electromagnetic propagation in urban environments. Mr. Parsons is currently chairman of the West Midlands Section of the Institution.

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Professor W. Gosling (Fellow 1968) has occupied the Chair of Electrical Engineering at the University College of Swansea since 1966 and he has recently been appointed for a three-year term as Vice Principal of the College. A Vice-President of the Institution, fuller notes on Professor Gosling's career appeared in the September 1972 and January/February 1973 issues of the *Journal*.

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Multiple channel u.h.f. reception on naval ships

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Based on a paper presented at the IERE Conference on Radio Receivers and Associated Systems held at Swansea from 4th to 6th July 1972.

SUMMARY

The environmental conditions which are present on a naval ship pose a number of problems in designing an efficient u.h.f. receiving system not normally encountered in a typical ground station. These conditions and the methods which are adopted to ensure reliable and efficient communication are described in the paper.

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1 Introduction

To remain an efficient and viable weapon, a modern naval ship must contain a communication system which is capable of passing information with a high degree of reliability and with the minimum of delay. The high-frequency band between 3 and 30 MHz is intensively used world wide and suffers from variable propagation conditions, so that for paths within the radio horizon other frequency bands are more effective, for example 225 to 400 MHz which is specifically allocated for use by mobile services. This band has many advantages, such as a large total frequency range allowing a less crowded use of the spectrum, low atmospheric noise level, more reliable propagation conditions, minimal bandwidth constraint on modulation method and smaller aerial system structures. Nevertheless, a number of the more general problems associated with communications together with those special to ship installations have to be resolved.

2 The Naval Ship Environment

The radio communication requirements of a naval ship are such that it may be necessary to operate some tens of channels simultaneously, most of which fall within the u.h.f. band. Constraints set by the size of the ship and the demands on the deck and superstructure for weapons and their associated control systems preclude the use of individual aerials for each communication channel and methods have to be employed to effect a reduction by using a common aerial for a number of channels.

Because signals are required to be received simultaneously from different bearings and different elevations on differing frequencies, the aerial must present wideband terminal impedance and radiation characteristics. The radiation characteristics must be sensibly circular in the horizontal plane and be of maximum amplitude at low angles of elevation in the vertical plane. Since these characteristics must be maintained in the ship environment it is necessary for the aerial to be sited high on the ship's structure and relatively free of obstruction from the superstructure. It must therefore be of simple design, physically small and lightweight in construction.

Consideration must also be given to propagation characteristics and the effects of different phenomena, such as multi-path reflexions from the surface of the sea, in selecting the most suitable plane of polarization. Taking all the factors into account, both electrical and mechanical, distinct advantages are shown by adopting a vertical plane of polarization.

The simultaneous use of one aerial for a number of circuits can best be achieved by using a technique known as common aerial working (c.a.w.), a means whereby a number of transmitters or receivers may be coupled to one aerial with the minimum of interaction and degradation in the performance of each circuit. This technique enables the number of aerials for a given installation to be drastically reduced, thus minimizing the siting problem, and also allows the remaining aerials to radiate to the best advantage by considerably reducing mutual interaction.

There is also the need for two-way communications. This postulates the co-siting of transmitters and receivers in the same ship. Receiving aerials are, if they have the the characteristics specified above, placed in an environment of high electromagnetic fields not only from the aerials associated with the transmitters in the u.h.f. band but from those of powerful radar circuits. The receivers must therefore be sufficiently robust to withstand high r.f. potentials without damage and be sufficiently linear so as not to present a degraded sensitivity performance.

The presence of high fields brings the added problem of mutual interference generated by non-linear effects in the ship's metal structure or parts of the communications system. High potentials induced into receivers could be minimized by siting the receiver aerials away from the main radiation lobes of directional aerials and in a mutual null of the more omnidirectional aerials. Since however both transmitter and receiver aerials are required to provide their maximum field sensitivity in the horizontal plane a vertical co-linear arrangement of transmitting and receiving aerials around the supporting mast structure offers the best solution. Whilst the use of a common transmitter and receiver aerial marginally reduces siting problems, the problem of protecting the receivers from the high potentials of the transmitters becomes almost impossible to solve especially in the multiple installation.

The effects of out-of-band radiation can best be minimized by the use of highly selective filters which may be located between the aerial and the receiving equipment. The technique of common aerial working makes use of such filters and these are used to advantage in further minimizing any high potentials presented to the receivers.

Whilst the siting of the receiving aerials in a relatively radiation free environment in an elevated position on the ship's mast assists in reducing non-linear interference from the ship's structure, the magnitude of the interference demands the same care is taken with the transmitting circuits. Further, because it is only practicable to obtain comparatively limited degrees of isolation between the transmitting and receiving aerials extreme care must be taken in the construction of both aerials and feeder arrangements, such as the use of like materials in the construction of conductors and connexions and the use of solid rather than stranded conductors in the cabling. Even the use of these precautions can result in poor performance if the mechanical construction is unsound.

Finally, it is necessary for the equipment to operate reliably and efficiently in the general mechanical and atmospheric environments of the ship with particular regard for withstanding shock, vibration and extremes of humidity and temperature.

It will therefore be seen that the performance of a radio receiver in the shipborne environment relies to a marked extent on the performance of other equipments in the communications system to obtain a realistic performance particularly in reducing the high induced voltages in the aerial system. The equipments integrated into the system are shown in schematic form in Fig. 1 with a list of the relevant factors affecting the system

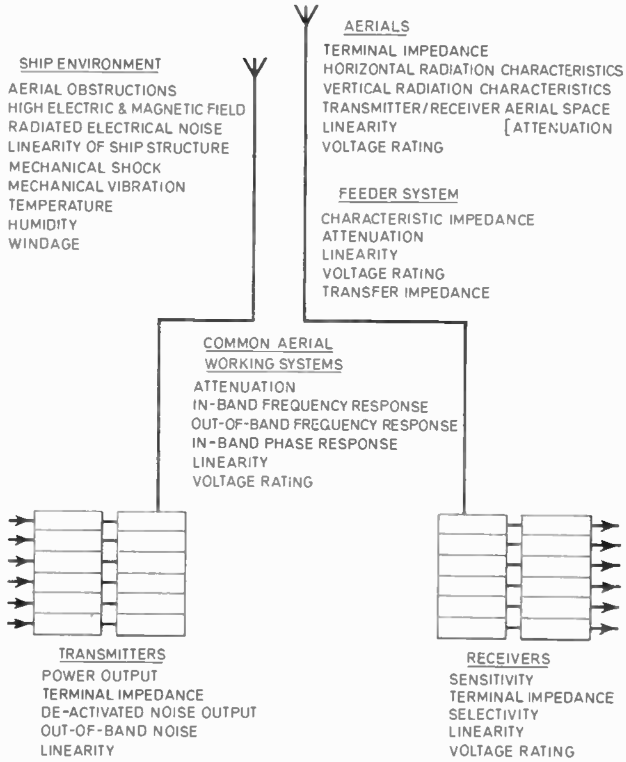

Fig. 1. System schematic.

performance. Typical values for the characteristics are shown in Table 1. These values are justified later in the paper.

Table 1

Equipment performance

AERIAL SYSTEM (TRANSMITTING AND RECEIVING)
- Frequency band: 225 to 400 MHz (broad band).
- Terminal impedance: sensibly 50 Ω.
- Degree of mismatch: production of a voltage standing wave ratio of not less than 0·5 with respect to 50 Ω.
- Horizontal radiation characteristic: circular within ±1 dB.
- Vertical radiation characteristic: cardioid at the first-order resonance with maximum radiation in the horizontal plane throughout the band.
- Gain: sensibly unity, with respect to a dipole.
- Space attenuation between the transmitting and receiving elements: not less than 40 dB.
- Product linearity: at least −130 dB.
- Voltage rating (transmitter): minimum peak inception voltage 165 V per channel.

FEEDER SYSTEMS
- Characteristic impedance: 50 Ω.
- Attenuation: 2 dB.
- Product linearity: at least −130 dB.
- Voltage rating (transmitter): minimum peak inception voltage 165 V per channel.
- Transfer impedance: 10^{10} Ω.

COMMON AERIAL WORKING SYSTEM
- Frequency band: 225 to 400 MHz.
- Tuning characteristic: infinitely variable across the band.
- Characteristic impedance: 50 Ω.
- In-band attenuation per channel: 3 dB maximum.
- Selectivity: 30 dB or greater on the adjacent channels, say 1 MHz operation.
- Product linearity: at least −130 dB.
- Voltage rating (transmitter): minimum peak inception voltage 165 V per channel.

3 Receiver Environmental Performance

3.1 *Sensitivity*

In the higher v.h.f. and the u.h.f. bands atmospheric noise is of a sufficiently low level that at normal temperatures the receiver input sensitivity is limited solely by the noise generated by its own characteristics.

Considering the case of an analogue voice signal from a double sideband amplitude modulation (d.s.b.a.m.) mode of transmission modulated to a depth of 30%, utilizing a characteristic band of 300 Hz to 3300 Hz and fed into a receiver of a noise factor of 10 dB, then for a minimum receiver output signal plus noise to noise ratio, $(S+N)/N$, of 10 dB the input signal from a source of an equivalent resistance of 50 Ω will be required to be of at least 2·2 μV e.m.f.

This may be determined from the equation,[1]

$$e^2 = 8KTBRnF/m^2$$

where K = Boltzmann's constant, $(1{\cdot}38 \times 10^{-23}$ J/K)
T = absolute temperature, typically 293 K
B = effective post-detection characteristic bandwidth in Hz
R = equivalent noise resistance of source in ohms
n = signal/noise power ratio at the receiver output
m = degree of modulation
F = receiver noise factor (power ratio).

Assuming a matched generator and receiver source resistance, the minimum carrier power which will be absorbed by the receiver will be $2{\cdot}4 \times 10^{-14}$ W.

In the case of a frequency modulated transmission, the minimum carrier e.m.f. which must be present is given by,[1]

$$e^2 = 8KTB^3RnF/3D^2$$

where D = normal deviation from mid-frequency.

In the ship environment, the presence of the high radiated fields from the ship's own transmitters, as described in Section 3.3.2, which may be working on adjacent channels preclude the use of higher sensitivities than can be achieved with amplitude modulated transmissions. Hence, assuming K, T, R, n and F have the same values it is necessary for B^3/D^2 to be not greater than 1×10^5. To cater for the different requirements the receiver has been designed to have a maximum characteristic band of 15 kHz and a maximum deviation from the carrier of 20 kHz.

3.2 *Transmitter Power*

The transmitter carrier power requirements can be readily obtained for ship–ship and ship–air communications circuits since the circuit is free from obstructions which could modify the field strength in the vicinity of the receiving aerial. Similarly, atmospheric noise is sufficiently low that it may be neglected. With regard to anomalous propagation conditions, which may enhance or degrade the far field strength, it is usual to compute the required transmitter power on the basis of normal propagation and accept the slight reduction in range which is produced under the degraded conditions. A further source of communication failure can be attributed to multi-path reflexion. These effects may be minimized by siting the ship's aerials at an optimum height above the sea level. At a height of 100 ft, failure from this effect will not occur on ship–ship circuits out to the radio horizon and can only occur for a near negligible part of the range on ship–air circuits.

Kitchen and Redhouse[2] have applied and extended the work of Norton[3] and shown that the path attenuation between ships at the radio horizon is approaching 139 dB and between ship and aircraft will be between 134 and 150 dB. Therefore, for ship–ship communication to be maintained out to the radio horizon and, allowing for feeder and c.a.w. system losses in both ships of 10 dB, the transmitter carrier power must be 149 dB above the minimum carrier input power required by the receiver $(2{\cdot}4 \times 10^{14}$ W), that is, 20 W. In the case of ship–air circuits it is necessary to increase the transmitter power by a further 6 dB, to cater for the worst condition, asuming negligible loss in the aircraft receiver aerial and feeder, to maintain communication out to the radio horizon. However, because of the additional problems produced in the ship installation by an increase in transmitter power coupled with the marginal reduction in performance of some extreme ranges on some frequencies a transmitter power of 20 W is accepted.

3.3 *The High Field Environment Requirements*

3.3.1 In-band frequencies

It is common practice for mobile radio communication circuits to operate in the simplex mode for two-way communication, therefore, if at a terminal the transmitting and receiving systems are sited in close proximity precautions must be taken to prevent damage to and, if there is a requirement for radio sidetone, distortion in the receiver by the potentials induced in its aerial from the local transmitter. First, the potentials must be reduced to a minimum by providing a reasonable path attenuation between the transmitting and receiving aerials. Second, sufficient protection must be provided within the receiver to prevent damage either by automatic gain control or muting or a combination of both.

The degree of isolation which may be reasonably produced is of the order of 50 dB, made up of aerial to aerial space attenuation of 40 dB and feeder and common aerial working system losses of 10 dB. Therefore, for a d.s.b. a.m. transmission modulated to 100% a peak e.m.f. of 600 mV will be presented to the receiver.

3.3.2 Out-of-band frequencies

Special care must also be taken in suppressing transmissions from the other local transmitters, particularly those which are close to a wanted transmitter frequency. These can cause interference to or complete suppression of a wanted weak signal by breakthrough, resulting from inferior screening, cross modulation and reciprocal mixing in the receiver mixer stage. In the presence of a multiplicity of local transmitters, intermodulation products may be generated by electrical non-linearity of joints or materials in the ship's structure (rusty bolt effect)[4, 5] and in parts of the communication system including the receiver.

The levels of the potentials which can be produced at the receiver input by the local transmitters may be as high as 20 mV peak e.m.f. from a d.s.b. a.m. transmission, some 76 dB above the minimum peak signal level which must be received. The only isolation given to the receivers is that offered by the aerial space attenuation, some 40 dB, feeder and common aerial working system losses, 10 dB, and the selectivity of the common aerial working system, 30 dB. Within this environment the receiver must retain its performance with the minimum of degradation down to a wanted signal level of 2·2 μV e.m.f.

Interference from non-linearity, intermodulation products, must be suppressed to a level of −115 dB with respect to the carrier power of any one transmitter at the transmitting aerial since the isolation in this case is only given by the aerial space attenuation, 40 dB, and the receiving feeder and c.a.w. system losses of 5 dB, resulting in a signal level of 10 dB lower than the weakest signal to be received. Similar precautions against non-linearity must also be taken in the receiving aerial, feeder and c.a.w. system. However, in this instance the levels of signals producing the interference are 45 dB lower than those generated in the transmitting system.

The effects of non-linearity in the ship's structure are the most difficult to overcome since the most economic materials and methods of jointing themselves, coupled with the corrosive environment, naturally produce the effect. Interference may only be minimized by siting the aerials clear of the structure in such a way that the transmitted fields within the structure are comparatively weak and the radiation characteristics of the receiving aerials are not predominantly in the direction of the structure.

3.4 *Alternative Aerial Sites*

So far, consideration has only been given to the best conditions which may be achieved in the ship environment. In a number of installations, such as those which require more channels than can be economically fed into a single aerial in a clear site or where it is impracticable to site the aerials clear of the ship's superstructure, other aerial arrangements must be used to supplement or used in place of the clear sited aerials.[6] Typical sites are on the extremes of mast yard arms, and result in a degradation in the radiated field characteristics. Isolation between transmitting and receiving aerials may also be reduced, typically by about 10 dB, increasing the in- and out-of-band peak potentials applied to the receiver to about 2 V and 66 mV respectively. The linearity of the system will also be degraded. An increase in the transmitter power to overcome the effects of the nulls in the radiation characteristics results in further aggravating the increased potential induced in the receiving aerials and ship's structure and hence the overall degree of non-linearity in the overall system. In these circumstances, the selection of specific yard arm aerials to obtain communication on a given bearing may be necessary and must be accepted.

No mention is made of the absolute performance requirements of the transmitters since these aspects fall outside the scope of the paper, nevertheless, for efficient communication it is necessary for their performance to match similar high standards. Characteristics which must receive particular attention are in-band

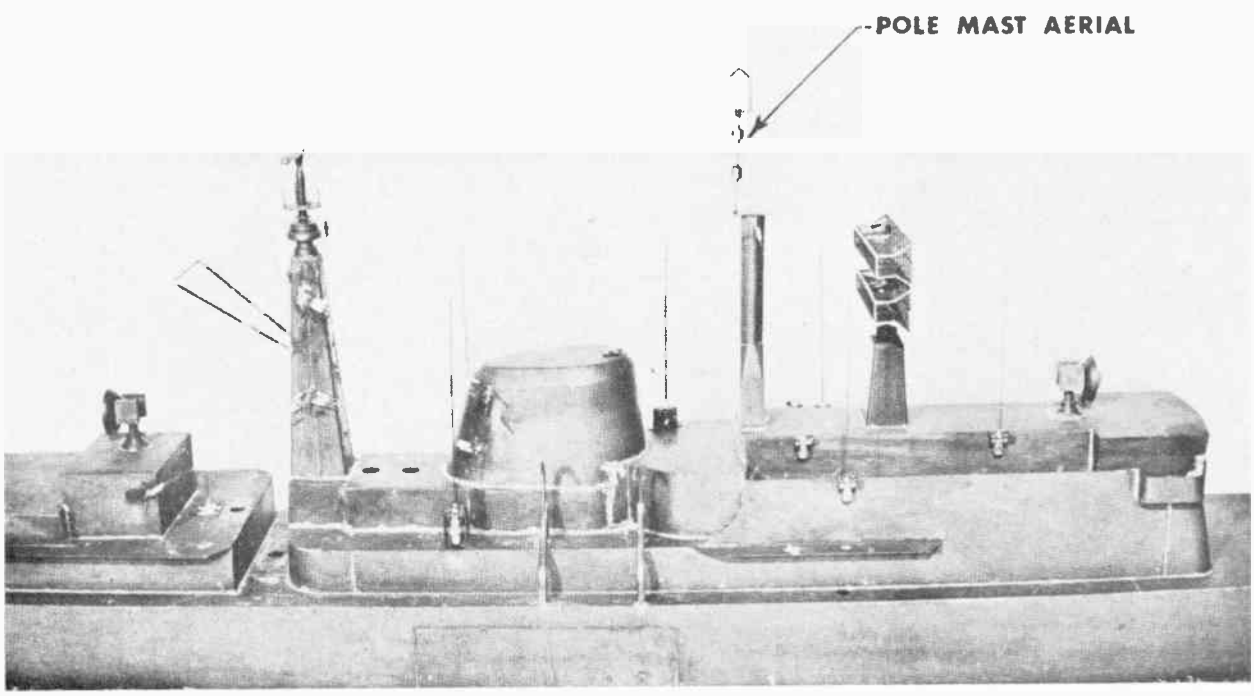

Fig. 2. Model mast structure.

noise in the de-activated condition, out-of-band noise and output stage linearity.

With regard to transmissions in other bands, they too must fulfil stringent requirements and reflect the same care in system design so that mutual interference is minimized. To cater for the variable conditions which may arise, the u.h.f. receiver must be capable of withstanding an e.m.f. of 20 V without damage.

4 Aerials

So that the aerials may be efficiently utilized an unobstructed site, clear of the ship's superstructure, is necessary such as high on the ship's mast. The need to use separate aerials for the transmitter and receiver circuits requires that the aerials be co-linearly sited to minimize mutual screening effects and to provide maximum isolation. These conditions can be most readily achieved by mounting each aerial around part of the mast structure such as a pole some 0·25 m in diameter which is fitted as an extension to an existing mast (Fig. 2). The need to site other aerials in the vicinity with minimum interaction is best achieved by producing the u.h.f. communication aerials in a cylindrical form, leaving the pole section free for running feeders to any aerials which may be sited on the mast head.

The need for each aerial to produce a sensibly circular radiation characteristic in the horizontal plane requires the use of more than one element such as dipoles, to each aerial and as the array of elements is required to produce a relatively constant impedance over a wide band of frequencies it is necessary to minimize the high terminal impedance presented on the even-order modes of resonance. It has been shown[7, 8] and used in a large number of aerial designs (References 6 and 9 for example) that by adopting a reduced aspect ratio the terminal impedance on the even-order modes is reduced without materially affecting the impedance on the odd-order modes. In this instance it has been found convenient to adopt this principle and develop four dipole aerials into two plane cylinders each 0·21 m long and 0·55 m in diameter. To minimize the effects of the supporting mast a cylindrical reflector, 0·92 m long and 0·32 m diameter is fitted behind the radiator cylinder and around the mast. Improvements to the terminal impedance characteristics can be made by inserting line choke sections across the aerial elements between each of the feed points (Fig. 3).

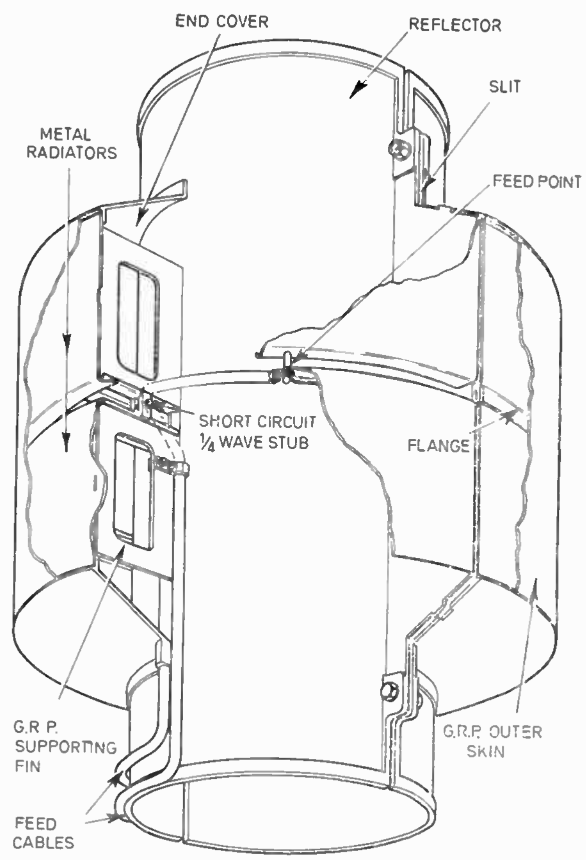

Fig. 3. Aerial arrangement.

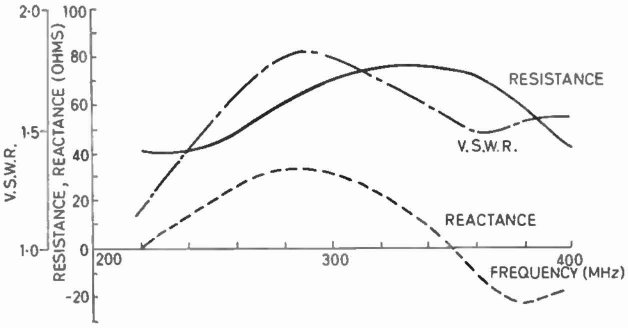

Fig. 4. Aerial impedance characteristics.

Each dipole of the above aerial arrangement is found to produce a v.s.w.r. which does not exceed 1·8 with respect to 50 Ω over the 225 to 400 MHz frequency band.

As it is more convenient to connect the four dipole feeders in parallel across the main aerial feeder the resulting impedance is reduced to 12·5 Ω and impedance correction is necessary. This is achieved by inserting a three-stage Chebyshev line transformer between the junction and the main aerial feeder. The transformer characteristics only produce an additional degradation in the impedance of 6%. The characteristics presented to the feeder are shown in Fig. 4. The radiation characteristics in the vertical plane with the aerial mounted in free space are found to be very close to a simple dipole aerial. However, the presence of the mast produces a small degree of distortion due to small circulating currents within the mast and modifies the characteristic between the limits of $-0{\cdot}4$ dB and 1·4 dB on any frequency with respect to a simple dipole. The characteristics in the horizontal plane are found to be within ± 1 dB with respect to a dipole.

Circulating currents induced in the pole mast are reduced and hence the isolation between two aerials mounted on a common mast improved by fitting a disk type r.f. choke around the mast between the two aerials. The reflecting cylinder mounted behind each aerial also provides additional isolation and these two factors enable a minimum isolation of 40 dB to be achieved when the aerial centres are displaced by about 2·5 m (Fig. 5).

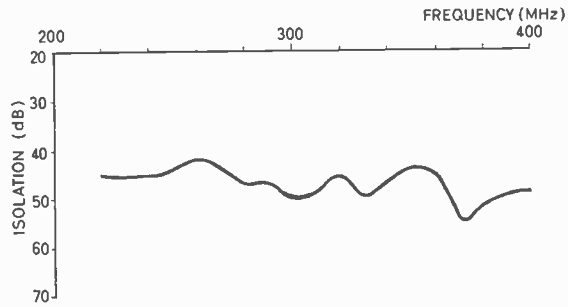

Fig. 5. Aerial isolation characteristics.

Non-linearity of the aerial and transformer is minimized by using similar metals, copper or brass, thoughout in the construction of the conducting materials. Special attention has also been paid to jointing, hard solder or compression connexions are used wherever possible, and solid as opposed to braided conductors have been used in the feeders. Extensive measurements have shown that these precautions have enabled the third-order product levels to be kept as low as 135 dB below the carrier power of the transmitters.

In order to minimize the overall weight, the aerials have been constructed with glass reinforced plastic (GRP) to support the radiator elements. The weight has been further reduced by forming the radiating surfaces from copper spinnings and embedding them in the GRP, suitably ribbed to obtain the required strength and rigidity. The reflector is also constructed from GRP and, to minimize the weight, a zinc spray has been employed to provide the reflecting surface.

As an aid to assembly around the mast, each aerial is constructed as two semi-cylinders which are bolted together and to the mast. The effects of the discontinuity of the radiating surface, approximately 1 mm wide, is to produce a minor degradation in the horizontal radiation characteristic and can be partially corrected by the adjustment of the line chokes. However, since their adjustment also affects the terminal impedance it is necessary for a compromise to be accepted.

The overall weight of each aerial is 21 kg.

5 Aerial Feeder Systems

Feeder systems between the aerial and the terminal equipment, sited in the communication offices, may in some instances be as long as 60 m (200 ft). Their runs may pass through numerous offices which contain other electrical equipment or machinery and be contained in cable forms containing many hundreds of conductors which may carry ship's power supplies, control circuits, both digital and analogue, telephone and communication circuits. It is therefore imperative that the aerial feeder cables present low loss, possess a high degree of isolation both to and from other cables, adequate thermal and voltage rating, withstand a high degree of shock and vibration and extremes of temperature particularly above deck. It is also important for the feeders to have a reasonable degree of mechanical flexibility to ensure they will not suffer damage from flexing during installation and will give reliable service after installation. Experience has shown that cable installation and maintenance costs can be considerably higher than for any other part of an electronic system and because of the complexity of many of the cable runs would have to be abandoned in the event of a failure. To recover a faulty cable results in disturbance of the other cables in the run and must be avoided at all costs.

Modern coaxial aerial feeders which are suited to the ship environment are now commercially available under a number of trade names. These consist of a convoluted solid outer conductor which has a high resistance to the ingress of the elements and offers a high degree of electrical isolation. Their dielectric is formed from a helix of low loss insulant such as polythene or, where there is a high ambient temperature requirement such as on the mast of a ship which is subjected to funnel gases, polytetrafluorethylene. The centre conductor, supported by the helix is either made from solid copper wire or copper tube. Experience has shown that the use of plastic serving on this type of cable, particularly when it is fitted on the weather decks, is inadvisable because of the ingress of moisture and funnel effluent through voids or faults in the serving and can quickly lead to corrosion of the outer conductor and an ultimate failure. Care must be taken at points where the feeder is supported on mast structures since electrolytic action rapidly develops unless a suitable insulant is placed between the conductor and the mast. Since the feeder contains a sensibly air-spaced dielectric the possibility of the ingress of moisture at joints must be eliminated. This is most easily prevented by pressurizing the feeder with dry air at a comparatively low pressure of about 138 000 newtons/m^2. Constant monitoring of the feeder by a pressure gauge gives early warning of a pending failure.

Commercial connectors for this type of feeder are also available. Whilst they are manufactured to suit a specific manufacturer's cable their mating faces are made to internationally agreed dimensions. A typical example is the E.I.A. flange connector. Provided a reasonable degree of skill and care is used during installation, the performance of these cables and connectors matches to requirements of the system. However, similar metals, such as brass, must be used for the connectors to minimize electrolytic action which will result in the generation of interference either as broad band noise or as non-linearity or both.

6 Common Aerial Working

Common aerial working (c.a.w.) is a technique which may be employed to combine a number of transmitter or a number of receiver circuits so that they may operate simultaneously from a single aerial. Numerous methods have been developed over the years such as ring couplers, non-reciprocal junctions and resonant line filters. Whilst ring couplers are comparatively small and simple, their isolation over frequency band ratios of 1·78 : 1 is unacceptable. Non-reciprocal junctions on the other hand present insufficient linearity and it is therefore necessary to resort to the use of the relatively large and expensive resonant line filter systems.

The resonant line filter systems utilize the filter characteristic to give the isolation between the channels connected to the common aerial thus minimizing interaction and maintaining a good impedance match between the aerial and the transmitters or receivers. Two basic methods of coupling have been developed, one whereby the transmitters or receivers are connected in parallel to the common aerial feeder and the second where they are connected in series.

The parallel system, primarily developed in this country has been used on naval ships for the last three decades. (Reference 10 and unpublished work of P. E. Trier, R. E. Fischbacher and H. P. Mason). However, because of the need to cluster the line filters around the common junction with the aerial feeder, the maximum number of circuits which may be combined is limited to about six. Because of this constraint, modern systems now have to employ the series system which allows up to twelve or even more circuits to be connected to the one aerial feeder.

The series system consists of a single aerial feeder which is terminated into a short circuit At discrete distances from the short circuit, sections of strip line which are at least a half wavelength on the lowest band frequency, one for each transmitter or receiver circuit, are connected into the aerial feeder. Connexion to each transmitter or receiver feeder is made by means of a sliding contact on each strip line. At a distance of a quarter wavelength of the geometric mean of the minimum and maximum band frequencies from the connexion, a narrowband line filter is fitted into the equipment feeders. The filter is designed to present low loss to the wanted frequency which has been allocated to that circuit and sensibly a short circuited impedance to all other frequencies.

In a receiving system, an incoming wanted signal on passing along the aerial feeder to a given receiver will be rejected by the short-circuited feeder termination provided the distance from the short circuit to the strip line connexion is adjusted to be $(2n+1)$ quarter wavelengths of the wanted frequency. The line filters to all other receiver circuits will also reject the wanted frequency provided they are detuned since their short-circuited input impedance is reflected as a very high impedance at their connexions in the strip lines. By similar reasoning, the system can also be shown to operate on transmitter circuits, however, the isolation presented by the system is insufficient to allow transmitters and receivers to operate from a common aerial with an acceptable frequency separation. It should also be remembered that there may be a need to retune a receiving circuit whilst the transmitter of another channel is in operation and this could result in the full signal power of the transmitter feeding into the receiver if only momentarily during the tuning process.

6.1 *Line Filter Design*

The most efficient type of u.h.f. resonant filter, capable of tuning between the extremes of the band, is the quarter wavelength resonant coaxial line.[12] Although comparatively large, it permits very high selectivities to be achieved with minimum of loss and be fitted with a relatively simple tuning mechanism. Typically, a copper resonant line of 0·127 m diameter will present an unloaded Q factor of $8{\cdot}15\times10^3$ at 225 MHz and $10{\cdot}9\times10^3$ at 400 MHz. Nevertheless to achieve the desired selectivity which will present 30 dB rejection at say 1 MHz off tune demands that the filter still requires to present a Q factor, loaded Q, when coupled into the feeder of $3{\cdot}75\times10^3$ at 225 MHz and $6{\cdot}7\times10^3$ at 400 MHz.

The loss factor, α, presented by the filter can be shown to be related to the loaded and unloaded Q factors (Q_L and Q_U respectively) by the equation

$$\alpha=(Q_U/Q_U-Q_L)^2$$

and by substitution of the above figures gives 5 dB loss at 225 MHz and 7·5 dB loss at 400 MHz, which are unacceptable.

It will also be found that the selectivity is so sharp that unacceptable attenuation is presented at the extremes of the pass band of $\pm$ 30 kHz, in addition, accuracy of tuning and the provision of thermal compensation is

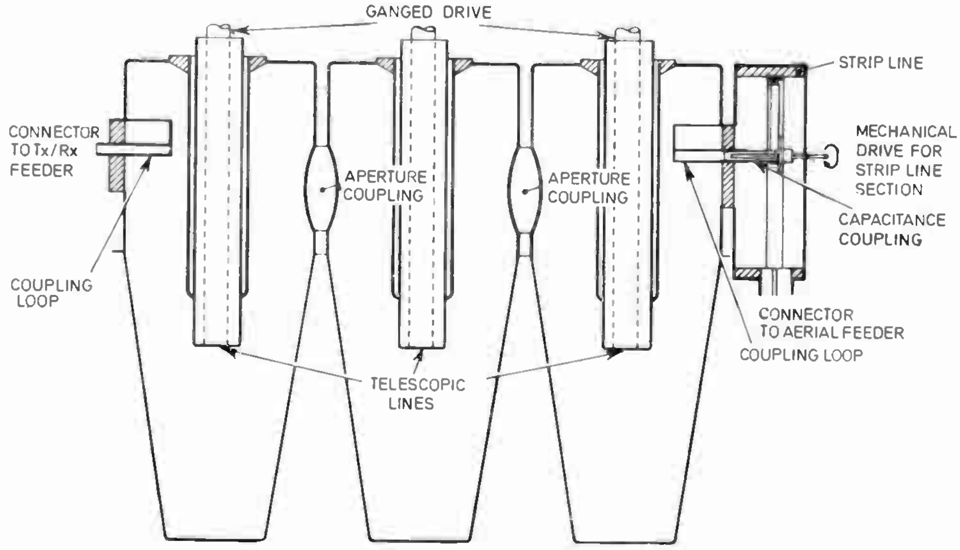

Fig. 6. Section of line filter and strip line.

difficult. However, if the number of resonant lines constituting the filter is increased and magnetically coupled, the required degree of selectivity may be obtained with an acceptable loss and passband (Fig. 6). Further, the problems of tuning accuracy and thermal compensation are considerably eased.

The loss may be shown from the equation

$$\alpha \simeq 1/[1-\sqrt[2n]{\alpha_1}/Q_U(f_0/f_1-f_1/f_0)]^{2n}$$

where n = number of resonant lines

f_0 = resonant frequency

f_1 = frequency at given rejection factor

α_1 = given rejection factor.

By substitution, when $n = 3$, $\alpha_1 = 10^3$, $f_0 - f_1 = 10^6$, the loss will be found to be 1·2 to 1·5 dB over the band of 225 MHz to 400 MHz.

An increase in the number of lines constituting the filter will be found to give no further reduction in loss, whereas a reduction to say 2, will produce a slight increase.

The most suitable method of coupling the feeders into the line filters is by inductive loops and between each line, by apertures. By positioning the loops and apertures along the lines with respect to the short circuited base a selectivity characteristic which approximates to the required linear increase of loaded Q with frequency may be achieved. However, since the law of coupling with band frequency is not linear, a higher selectivity at the extremes of the band, and consequently a higher loss, must be accepted. For this reason it is found better to approximate to a law of $Q_L \propto \sqrt{\omega}$, that is constant loss at the extremes of the band, rather than $Q_L \propto \omega$.

Assuming a constant input and output impedance to the lines and an increase of loaded Q which is proportional to the square root of ω, the most suitable position for coupling may be found from the expression

$$\cos \beta l_1 = (f_2/f_1)^{1\cdot 25} \cos \beta l_2$$

where βl_1 = position of coupling from the base at the the lowest band frequency (f_1)

βl_2 = position of coupling from the base at the highest frequency (f_2)

and if the line is homogeneous

$$\beta l_1/\beta l_2 = f_1/f_2$$

therefore

$$\cos \beta l_1 = (f_2/f_1)^{1\cdot 25} \cos (f_2/f_1)\beta l_1$$

If f_1 = 225 MHz and f_2 = 400 MHz, βl_1 is 0·65 radians. it can also be shown that

$$Q_L \propto 1/f^2 \cos^2 \beta l$$

from which the characteristic of Q_L to a base of frequency may be determined.

Applying the above characteristic to meet the selectivity requirements produces losses of 1·5 dB at 225 MHz and 400 MHz and 1·3 dB at 300 MHz (Figs. 7, 8 and 9). In practice the coupling loops and apertures have finite dimensions, and the loop presents a capacitance to the resonant centre conductor, resulting in modified characteristics. Also because of the capacitive effects of the coupling it is necessary to reduce the value of βl from that predicted.

Two other factors worthy of consideration in the design are the minimizing of the length of travel of the

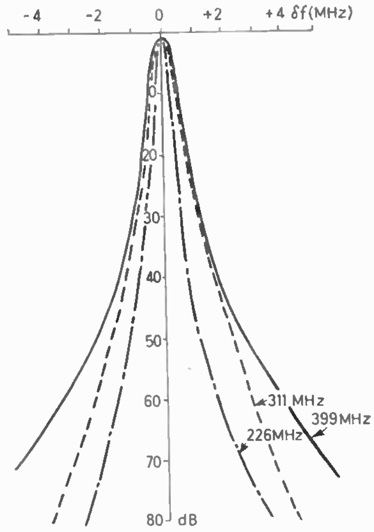

Fig. 7. Overall selectivity.

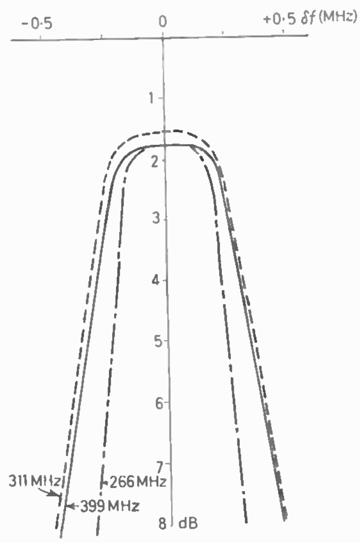

Fig. 8. Narrow band selectivity.

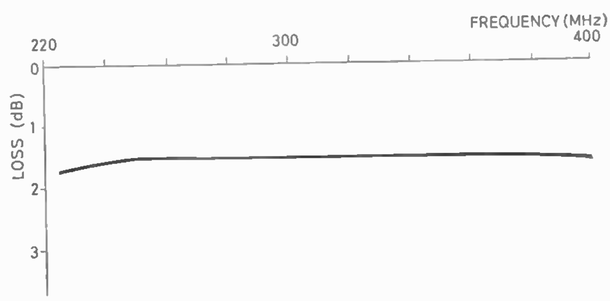

Fig. 9. Loss characteristic.

resonant centre conductor, thus reducing the overall length of the filter, and the reliability in maintaining the high selectivity during use. Neglecting the capacitive loading produced by the discontinuity at the open end of the resonant centre conductors, the rate of change of length of the conductors with frequency is 1·78 times as great at 400 MHz as at 225 MHz. Since the tuning mechanism must be designed to cater for the higher rate of change advantage of the accuracy is also taken throughout the band. This is achieved by producing a progressive increase in capacitance with a reduction of frequency by tapering the outer conductor. Taking into account the end capacitance[12] it has been found that a linear taper of the outer conductor produces a near constant rate of change of length with frequency. The use of helical lines to effect a reduction in overall length has not been employed on account of the difficulties of tuning over wide bands of frequency.[13]

In the ship environment it is necessary to ensure that there is no degradation in the very high performance required of the line filters, in particular, an increase in the r.f. resistance of the coaxial line sections. Should this happen it would result in a reduction of the unloaded Q factor and a consequent increase in loss. The short-circuited end of each line is the most vulnerable on account of the high circulating currents and for this reason the lines are constructed so that a permanent continuous surface is formed as far up the line as is practicable from the short circuit. The practical limit on the centre conductor occurs at the junction of the telescopic section and, to give access during construction, a flanged joint is made in the outer conductor at a similar distance, about 0·7 radians at 225 MHz. Whilst the base metal forming the lines is plated, its presence is as much for environmental protection as to improve or maintain the electrical performance. Experience with so-called low conductivity plated finishes, such as silver, over a number of years has shown that a degraded performance will invariably result in high Q circuits if the plating is more than a fraction of a micron thick.[14] This primarily results from the relatively poor finish and the practice of including additives in the electrolyte to obtain reliable plating.

6.2 *Strip Line Section*

To cater for all reflected reactances from the short-circuited aerial feeder termination and improvements in the matching of the reflected aerial impedance it is necessary for each of the strip line coupling sections to present an electrical length of a little greater than a half wavelength at the minimum band frequency. By introducing a low-loss dielectric base to the line its physical length may be reduced to reasonable dimensions as well as providing support to the conductor. Because of the need to make the connexion to the line filter adjustable, the line is developed into an annular form. Both the strip line and the contact are finished with noble metals, to minimize the risk of wear, the generation of interference either as broad band noise or from non-linearity, and to maintain an extremely low r.f. contact resistance. Both the input and output connexions to the line section are terminated in high grade connectors to allow the whole assembly for each channel to be withdrawn.

6.3 *Tuning*

Tuning of the line filters is achieved by means of anti-backlash screw drives mechanically coupled through gear trains. Thermal compensation is provided by connecting the telescopic resonant centre conductors to the drive mechanism through a shaft of low thermal expansion, connexion to the conductor is made at the open-circuited end and within the conductor.

Manual, fully automatic and preset automatic tuning facilities are provided to tune the line filters and strip line sections. Preset automatic tuning is achieved by means of an optical encoder which mechanically sets the line filter conductor and position of strip line connector to predetermined positions by comparison with a diode matrix store. The store has sufficient capacity for setting up twenty different frequencies on each channel. Selection of the wanted frequency may be made from the associated receiver or in conjunction with the receiver from a remote control position.

6.4 *Construction*

Each line filter with its strip line section and controls are mounted in withdrawable drawers. Up to six drawers with a power supply unit for the automatic tuning facilities together with the tuning meters may be housed in a standard type of service cabinet. Up to twelve drawers, provision for twelve channels, may be connected to a single aerial.

7 Receiver

The receiver is of single superheterodyne design and contains the following stages:

A pre-mixer r.f. amplifier with optional selectivity.

A balanced mixer.

Pre-set variable bandwidth r.f. amplifier centred on 50 MHz.

Detectors for double side band and frequency modulation.

Narrow and wideband output stages (300 Hz–3·5 kHz and 50 Hz–30 kHz).

Digital frequency synthesizer type of local oscillator with a crystal reference standard.

Solid-state devices are used throughout for amplification and where practicable for switching circuits. Because of the use of these devices amplifier stages are sensibly wideband and selectivity is provided by additional filtering circuits. Further, in the construction of the receiver the use of solid-state devices has allowed the advantages of the modular concept to be employed and variants of the basic receiver may be selected depending on the needs of a given installation simply by removing redundant modules. Figure 10 shows the basic design of the receiver, the pecked lines indicate the boundaries of the circuits contained in each module. Selection of the contents of each module are set by the requirement of the module in terms of redundancy for the various installations, the cost/reliability penalty, circuit function, overall size and heat dissipation.

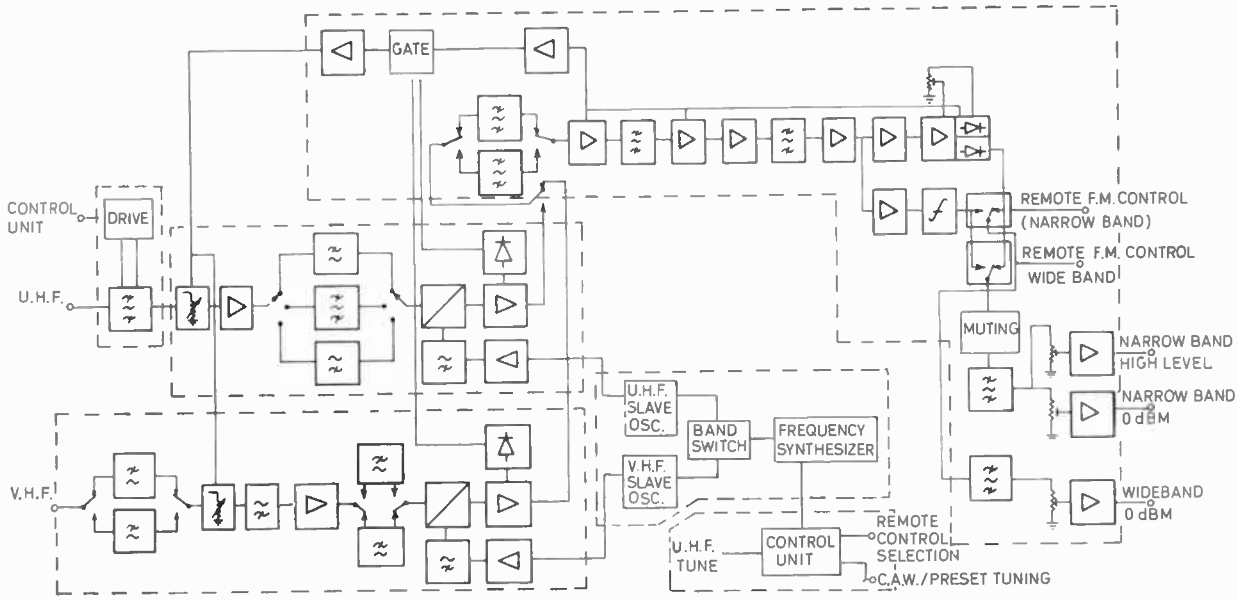

Fig. 10. Schematic of receiver.

7.1 *R.F. Amplifier and Mixer Stages*

The r.f. amplifier and mixer stages are contained in three modules, two of which, the variable selective filter for the u.h.f. band and the r.f. and mixer stages for very occasional use in the service v.h.f. band (100–156 MHz), are optional.

In a conventional installation it will have been noted that considerable selectivity is offered to the receiver by the c.a.w. system and to provide any worthwhile improvement within the receiver itself would result in a very inefficient circuit. Therefore it is not used in such installations. However, small ships which are fitted with the minimum of equipments do not justify the use of c.a.w. systems and to give some measure of protection in their absence the filter module would be fitted.

The filter consists of triple resonant line, each of about 5 cm × 5 cm section and 12·5 cm long. Each section of the filter is inductively coupled to produce a bandpass characteristic. Tuning is achieved by capacitance loading of the open-circuited end of the lines in 10° increments and is switched to a given increment by means of a small motor. The motor is controlled by means of digital code information passed from frequency synthesizer and is fully automatic. The loss of the filter is typically 0·5 dB and the bandwidth at the 1 dB point is 6 MHz. Whilst the circuit is mechanically complex the alternative use of varactor diodes is unacceptable on grounds of non-linearity in high electric fields.

The use of complex filtering in the v.h.f. circuits is unwarranted because of the limited use of the band.

Since the r.f. selectivity is provided either by the c.a.w. system or the optional r.f. filter no further attempt has been made to improve the selectivity in the r.f. amplifier or mixer. This decision has influenced the choice of converter used in the receiver. Until recently receivers generally incorporated r.f. amplification to improve the sensitivity and counteract the noise generated in the mixer stage. Whilst semiconductor mixer devices are less noisy compared with their thermionic counterparts it is still advantageous in the low atmospheric noise v.h.f. and u.h.f. bands to employ r.f. amplifiers to give added improvement in sensitivity although it could be argued that the low insertion loss and noise temperature characteristic of the Schottky barrier diodes as mixers under controlled operating conditions eliminated the need for r.f. amplification. Such arguments include the lower potentials which will be applied to the mixer for a given input signal level and achieving greater linearity. However, the device must be presented with the correct terminating impedance, either open or short circuited, at the image ports and this precludes its use as a wideband mixer unless input filters of greater complexity are used, particularly in the v.h.f. band. A further disadvantage of the Schottky barrier diode is relatively poor local oscillator rejection and re-radiation from the receiver could only be improved again by the use of more complex r.f. filtering in the absence of c.a.w. systems.

The use of the field effect transistor (f.e.t.) has been shown to offer greater advantages as a mixer for this application. Since its V_g/I_d characteristic more nearly approaches a true square law the magnitude of interference from odd-order intermodulation products, particularly the third, will be lower. A balanced circuit using a common base gives additional protection against local oscillator breakthrough and presents a relatively low input impedance which again assists in maintaining linearity. However, if input line impedance of 50 Ω is employed the image transfer is increased to approximately 200 Ω and the resulting increase reduces its dynamic range.

The f.e.t. mixer developed for this receiver has a noise figure of approximately 12 dB over the band of 225–400 MHz and a conversion gain of between 0 dB and +2 dB. It was found that the application of two 500 mV e.m.f. signals produces a third-order product at the output port which is 35 dB below the output level of the product producing signals.

The noise figure of the mixer dictates the necessity for low-noise r.f. amplification which again reduces the useful range of the mixer. To minimize the demands on the r.f. amplifiers as regards noise and linearity, the inclusion of an i.f. pre-amplifier became necessary and as this circuit has little protection from strong interfering signals it must possess a high degree of linearity over a large dynamic range. The choice of a suitable device fell on a multi-emitter transistor operating at a relatively high collector current which typically has a gain of 10 dB and a noise factor of 8 dB.

If it is assumed at this stage that the noise factor of the main i.f. amplifier can be maintained at a figure of about 17 dB including the effects of filter losses at the amplifier input, the gain of the r.f. amplifier to be 9 dB and the overall noise factor of the receiver to be 8 dB, and allowing 2 dB tolerance for production, then from the equation

$$F = F_1 + \frac{F_2-1}{G_1} + \frac{F_3-1}{G_1G_2} + \frac{F_4-1}{G_1G_2G_3}$$

where F = overall receiver noise factor,

F_1, G_1 = r.f. amplifier noise factor and gain respectively,

F_2, G_2 = mixer noise factor and gain respectively,

F_3, G_3 = i.f. pre-amplifier noise factor and gain respectively,

F_4 = i.f. amplifier noise factor.

The r.f. amplifier noise factor must not exceed 5 dB. In addition the r.f. amplifier must have a high degree of linearity over a large dynamic range and a high constant gain-bandwidth product. Multi-emitter devices with the emitters connected in parallel using a common collector are found to fulfil the requirements.

Since the r.f. amplifier and mixer are both broadband designs there is a need to restrict the noise on the image frequencies which is generated in the r.f. amplifier. To provide attenuation on these frequencies filters are fitted between the r.f. amplifier and mixer. In the u.h.f. band they are a combination of low- and high-pass filters and their attenuation together with that provided by the variable input provides an image rejection in excess of 80 dB. In the v.h.f. band circuit image protection is provided by two sets of low-pass filters fitted in front of the r.f. amplifier and an identical set between the amplifier and mixer producing an image rejection of greater than 80 dB.

7.2 *Automatic Gain Control*

Protection against over-voltage to the r.f. amplifier, mixer and i.f. pre-amplifier and dynamic range limiting of the r.f. amplifier is provided by an automatic p-i-n diode attenuator which is fitted between the input filter and the r.f. amplifier. Under normal operating conditions the dynamic range of the receiver is 108 dB of which 75 dB can be provided by applying a.g.c. to the main i.f. amplifier. The remainder must therefore be provided by the p-i-n diode attenuator. The attenuator may be activated from two sources, either by comparatively large signals appearing within the pass band of the main i.f. amplifier or by large out-of-band signals appearing across a voltage divider at the output of the i.f. pre-amplifier. Comparatively large signals appearing in the main i.f. amplifier will not only operate the normal i.f. amplifier a.g.c. but will also provide a d.c. drive to a gated amplifier which operates the p-i-n diode attenuator. Activation occurs when the inband input signal level is in excess of approximately 200 μV e.m.f. Large out-of-band signals operate the gated amplifier directly from the potential appearing across the i.f. pre-amplifier voltage divider and activate the gated amplifier when the input signal level is in excess of approximately 200 mV e.m.f.

Attenuators are fitted to both the v.h.f. and u.h.f. r.f. amplifier circuits; however in the case of the u.h.f. circuit the pass band of the mixer output stage is roughly equal to the pass band of the input filter. Consequently the r.f. amplifier is already protected against large out-of-band signals. When the receiver is connected to a c.a.w. system additional protection is provided by the greater selectivity of the line filters. In the case of the v.h.f circuit, no input filter protection is provided and unwanted signals which are outside the mixer output pass band remain undetected at the r.f. pre-amplifier voltage divider. Since under certain conditions, unwanted signal levels at the receiver input can be as high as 20 V e.m.f., certain damage to the r.f. amplifier will occur unless other means of protection are provided. To prevent damage from this source, a monitoring point is provided between the attenuator and the r.f. amplifier which provides a potential to activate the attenuator. Because of the comparatively large power which is dissipated in one of the arms of the attenuator when the receiver is subjected to the 20 V e.m.f. signal, it is necessary to provide a suitable conduction heat sink for the diode. It should be appreciated that the receiver is not required to operate in the presence of such signals but should remain undamaged and operate with its normal sensitivity on the removal of the signal.

7.3 *Linearity*

7.3.1 Cross-modulation

One of the sources of interference produced by non-linearity in the presence of a strong interfering signal displaced from the wanted signal is cross-modulation. The performance of the receiver under test conditions shows that, with an interfering signal of 300 mV e.m.f. modulated to 30%, the cross-modulation is at least 20 dB below the wanted signal level.

7.3.2 Intermodulation

A second source of interference again produced by non-linearity is that of intermodulation caused by some of the odd-order products. Invariably the strongest are the third-order produced by two unwanted strong signals on, say, frequencies f_2 and f_3 giving a product $(2f_2-f_3)$ or $(2f_3-f_2)$, on some frequency close to the wanted signal f_1. Under test conditions the receiver has shown itself capable of giving at least 6 dB signal plus noise to noise ratio $(S+N)/N$ when the unwanted signal amplitudes, are at least 66 dB greater than the wanted signal level of 2·2 μV e.m.f. modulated 30% and displaced from the wanted signal by approximately 1 MHz and 2 MHz.

7.3.3 Reciprocal mixing

In the presence of strong interfering signals reciprocal mixing, mixing produced by the strong interfering signal with the local oscillator sideband noise, may produce a signal at the intermediate frequency. This can be particularly troublesome if the level of the interfering signal and the local oscillator sideband noise is not minimized at mixer. In this design of receiver the local oscillator frequency is derived from a frequency synthesizer which contains a multiplicity of high speed fixed and variable divider stages and slave oscillators. Special care must therefore be taken to provide adequate screening and filtering within the synthesizer module. It has been demonstrated that an interfering signal 80 dB above the wanted signal and displaced from the wanted signal by at least 1 MHz will produce no marked degradation in the $(S+N)/N$ ratio of the wanted signal.

7.4 *Intermediate Frequency Amplifier*

The intermediate frequency amplifier is based on a design developed for a service airborne equipment and operates on a mid-band frequency of 50 MHz. It was selected so that a common basic design of synthesizer could be used for the two equipments.

In common with other circuits, the i.f. amplifier uses wideband amplifier stages and selectivity is provided by two crystal band pass filter stages. Two i.f. bandwidths are provided, one at 70 kHz (3 dB points) and a second at 37 kHz (3 dB points) with provision for a third band.

The basic i.f. amplifier has a noise factor of 5 dB; however, since it is preceded by one of the filters with a maximum in-band loss of 12 dB the overall noise factor is degraded to 17 dB. So that the detectors may be protected from wideband noise, the second filter stage is situated adjacent to the detectors. The filters provide 90 dB rejection at ±100 kHz on the wideband characteristic and 55 dB rejection at ±55 dB on the narrow band characteristic.

7.5 *Synthesizer*

The synthesizer used with this receiver is identical with that developed for the airborne equipment mentioned above and is based on the principles developed for the military range of communication equipment described briefly elsewhere.[15] Provision is made to select any frequency between 100 and 155·975 MHz and 225 and 399·975 MHz in multiples of 25 kHz, of which 19 frequencies may be stored in a non-destructive magnetic memory store. This facility allows rapid frequency selection from either the control unit at either the local or remote position.

7.6 *Control Unit*

Full control of the receiver is provided on the front panel of the control unit which is 89 mm high and 482 mm wide (standard 19-inch panel). It contains the following control and monitoring facilities:

Analogue frequency setting (5 controls).
Preset frequency store control (19 frequencies and manual).
Modulation mode control combined with the local/remote control.
Headset connector.
Volume control for headset.
Supply switch and fuses.
Diagnostic test set connector. (Allows faulty modules to be located from a 'go/no-go' diagnostic test set.)
Frequency logging panel.

In addition, provision is made behind the control panel for the preset channel selector information to be fed to the c.a.w. system.

7.7 *Construction*

The receiver is constructed in modular form and each module is plugged into the chassis and retained by locating screws. Electrical connexion with the chassis wiring is made at the same instant as the module is plugged into the chassis through multi-way connectors. The chassis is totally enclosed to prevent the ingress of moisture and foreign matter and is fitted with a readily removable sealed cover. Provision is made to equalize the pressure within the chassis should it exceed -3450 newtons/m^2 or ± 6900 newtons/m^2 with respect to the atmosphere through a dessicator unit.

Conduction cooling from the modules to the chassis is employed and the chassis is cooled by natural convection. The control unit is also constructed in modular form and is attached to the front of the main chassis in a similar manner to the modules.

A number of receivers may be mounted in a cabinet of a capacity to suit the ship installation requirements.

7.8 *Brief Overall Electrical Specification*

Frequency coverage

V.h.f. 100 MHz to 155·975 MHz in 25 kHz increments.
U.h.f. 225 to 399·975 in 25 kHz increments.

Modulation

Double side band amplitude modulation.
Frequency modulation, maximum deviation ±20 kHz characteristic band.
Narrow 300 Hz to 3·5 kHz.
Wide 50 Hz to 30 kHz.

Frequency stability

Better than ±5 parts in 10^6 over a period of 12 months.

Sensitivity

Noise factor not worse than 10 dB.
2·2 µV input e.m.f. d.s.b.a.m. modulated to a depth of 30% results in a 10 dB $(S+N)/N$ ratio over a 3 kHz characteristic band.

Quieting signal sensitivity

2·2 µV input e.m.f. modulated 1 kHz, deviation 3 kHz results in a 25 dB $(S+N)/N$ ratio over a 3 kHz characteristic band.

Image rejection

Not less than 80 dB.

Overall selectivity

Rejection greater than 55 dB at 50 kHz.
Rejection greater than 90 dB at 100 kHz.

Linearity

Cross modulation -20 dB from a 300 mV e.m.f. interfering signal on 30% modulation.

A.f. output characteristics

Narrowband 0 dBm, 600 Ω, balanced, and 2·5 V r.m.s. into 100 Ω, balanced.
Wideband 0 dBm, 600 Ω, balanced.

8 Conclusions

The paper shows the need for system design of communications installations which are required to operate to an exacting specification. The use of arbitrary equipment in such installations invariably results in operational and performance limitations which fall well short of the requirements. Conversely, a number of the units of the system may well have been over-designed for the requirements which result in unnecessary complexity and wastage of valuable space, additional cooling problems and greater unreliability. No specific mention is made in the paper of reliability, maintenance and spares support. The factors are nevertheless extremely important and great attention has been paid to them in the system and equipment design having regard for the overall maintenance costs involved in the life of equipments in the naval ship environment.

9 Acknowledgments

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Receiver innovations for single-sideband broadcasting

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Based on a paper presented at the IERE Conference on Radio Receivers and Associated Systems held at Swansea from 4th to 6th July 1972.

SUMMARY

Recent developments in integrated circuit technology and proposals by the European Broadcasting Union for single-sideband broadcasting have renewed interest in the re-design of broadcast receiver structures for the long and medium wave bands. This paper outlines the requirements for an s.s.b. receiver and reviews some of the many designs for such receivers which have been published recently.

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1 Introduction

Domestic broadcasting has now been with us for half a century, during which amplitude modulation has remained as the means of medium-wave broadcasting. Within the last few years, however, interference between stations has increased, but fortunately electronic technology has improved to such an extent that serious attention can now be paid to alternative modulation techniques.

Although a variety of compatible single-sideband systems have been proposed, they possess an inherent sensitivity to propagation disturbances, and also give rise to a higher level of out-of-band radiation than the a.m. transmitters they replace. For this reason a more conventional s.s.b. system with some sort of pilot carrier was proposed about five years ago. The paper by Eden, published in 1967, may be regarded as the first milestone in this direction.[1] Single sideband has long been familiar in communications applications, but a radically different approach to receiver design will be necessary if s.s.b. is to make a successful debut in the home. The requirements of such a receiver may be listed as follows:

(i) be simple to operate;

(ii) give adequate quality of reproduction;

(iii) be economic to produce and service;

(iv) be reasonably sensitive and use an internal aerial;

(v) receive existing double-sideband a.m. as well as the new s.s.b. signals, and with minimal difference in characteristics;

(vi) be free from extraneous noises such as heterodynes.

It is the purpose of the paper to set out some of the recent domestic s.s.b. receiver ideas that have appeared both in this country and in Europe.

2 Receiver Structures

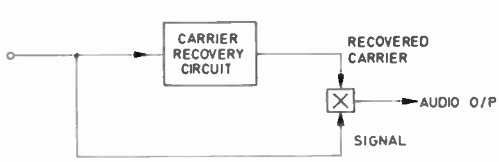
For decades the basic architecture of domestic radio receivers has remained essentially unchanged. The superheterodyne principle has been found to give adequate selectivity, sensitivity and stability, and the envelope detector has been a simple but effective method of recovering the modulation and providing a.g.c. The production of standard families of valves and transistors gave rise to little basic diversification in design; there was no need to change well-trusted principles.

The advent of the integrated circuit has brought with it the ability to design receivers of greater complexity and reliability for a given outlay, features that will be necessary if a new modulation system is to be used. The full extent of the changes in design philosophy that these developments will bring cannot yet be foreseen, but it seems probable that in 10 years time the envelope detector will not possess the monopoly that it does today. On the other hand, the superheterodyne principle will be assured of a place; although synchronous reception makes direct conversion receivers feasible in principle, stability and blocking problems threaten its widespread acceptance. The desire to integrate as much of the circuit as possible will provide additional impetus to change many widely accepted standards; the component designer is now a system designer as well. The part of the receiver which, more than any other, determines the end design is the detector itself, which is the subject of the next section of this paper.

3 Receiver Detectors

If a 'full' carrier (6 dB below p.e.p.) is transmitted together with a single sideband the programme may be received with an average of about 8% distortion using an envelope detector, so that a measure of compatibility with existing receivers is retained. A series of experimental broadcasts was made during March 1972 from a transmitter in Germany, and little difference was in fact discernible between s.s.b. with 'full' carrier and the corresponding a.m. signal. However, in order to realize the full advantages of s.s.b., especially a lowering in the programme distortion due to selective fading, a specially designed s.s.b. detector is really necessary. This conventionally consists of a 'product' detector, although another type of detector is possible in which the phase information present in an s.s.b. signal with 'full' carrier is detected and used to correct the envelope waveform of the signal.[2,3] Unfortunately, this principle is very sensitive to overmodulation which is very likely to occur on m.w. transmissions, and because of this shortcoming this method will not be discussed further here.

The principle of the 'product' detector is widely known: the received signal is multiplied by a carrier wave with the same frequency as that of the signal. Such detectors may be divided into two groups according to the manner in which they produce the demodulating waveforms.



(a) Basic homodyne.

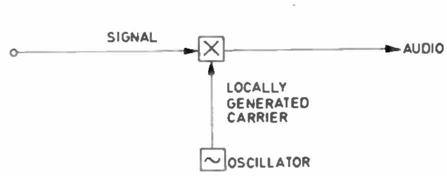
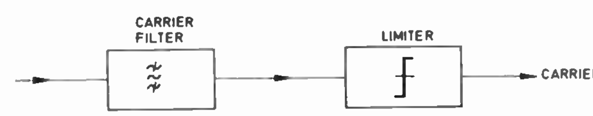


Fig. 1 (b) Basic synchrodyne.

The first of these groups may, using the terminology of Tucker, be termed 'homodyne' receivers,[4] which recover the carrier for the product detector from the incoming signal, as shown in Fig. 1(a). Adequate homodyne reception is considerably more difficult for single-sideband signal than for double-sideband a.m., because of the large variations in instantaneous signal phase that occur with the former. The pilot carrier is isolated by means of a narrow filter and an amplifier; limiting is frequently used to remove any remaining amplitude modulation of the demodulating wave. The bandwidth of the pilot filter is determined by the suppression of the transmitted carrier and the degree of quadrature distortion that would be acceptable, and in practice this leads to a pilot filter bandwidth of about 10 Hz and the use of a quartz or N-path[5] filter. This is essentially the 'exalted carrier' method of Crosby.[6] The requirements of the pilot filter have been recently examined by Bruch,[7] who has calculated the degree of second-harmonic distortion to be expected at various levels of carrier suppression. He has shown that, for suppressions of the order to be expected in practice, the bass end of the reproduced audio spectrum can be noticeably distorted. A further problem associated with the homodyne method of reception is the very narrow tuning range imposed by the high selectivity of the pilot filter. This makes some form of electronic tuning assistance mandatory; several schemes have been suggested and demonstrated, and these will be discussed in the next Section.



(a) Conventional carrier recovery.

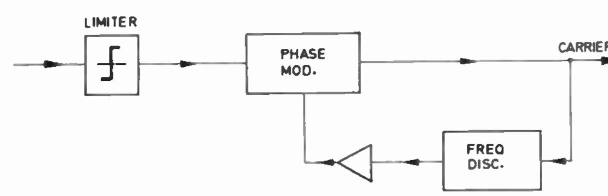


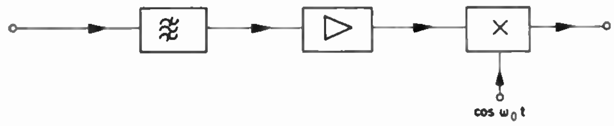
Fig. 2. (b) Feedback carrier recovery.

An alternative method of carrier recovery appears in a paper by Langecker.[8] Provided over-modulation does not occur, the average frequency of the amplitude limited signal is that of the carrier. Thus, by reducing the frequency (or phase) modulation of the limited signal the carrier may be recovered. This is achieved by means of a feedback loop where the output from a frequency discriminator is used to control a phase modulator, as shown in Fig. 2(b).

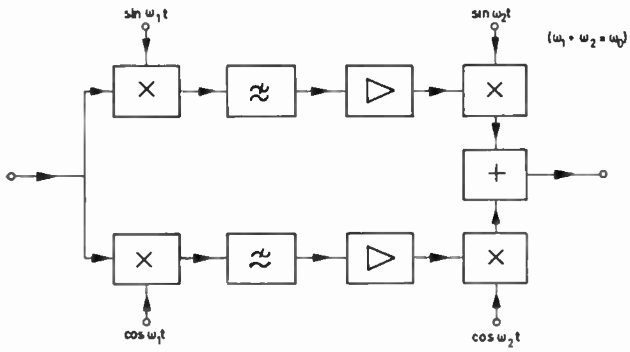
The other main group of detector types used for s.s.b. is the 'synchrodyne' group. Here the carrier wave for the product detector is generated locally by means of an oscillator, which may or may not be automatically synchronized to the received signal. The synchrodyne method is the one almost universally used for single-sideband reception in telecommunication circuits.

In conventional s.s.b. receivers the required sideband of the signal is usually selected by means of a piezoelectric or mechanical filter, but although this method gives good results, it remains obstinately expensive for domestic application. It is therefore fortunate that the use of synchronous demodulation enables us to carry out post-detection filtering to control the shape of the receiver passband. Figure 3 shows three basically different types of circuit; the pre-detection filter method, the two-path or Barber circuit, and the phasing method.

The Barber circuit consists of two matched intermediate frequency amplifiers with low-pass filters and



(a) Filter method.



(b) Two-path method.

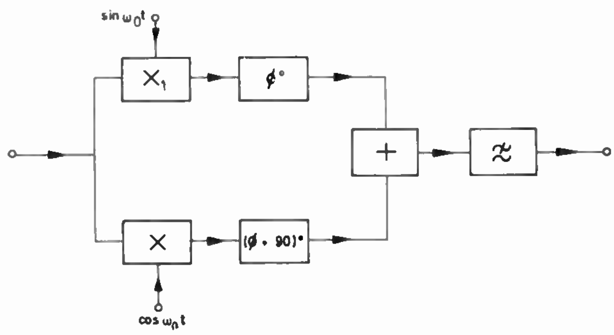


Fig. 3. (c) Phasing method.

product detectors. By operating them in phase quadrature and at an i.f. of about 2 kHz, it is possible to reject the image frequency and obtain a symmetrical passband corresponding in frequency to that of the derived sideband. The phasing method uses two product detectors in phase quadrature, thereby producing two a.f. outputs which are in turn shifted in phase by amounts differing by 90°. In this way the undesired sideband is again eliminated. One possible advantage of the phasing method in broadcast application is that it is possible to produce an asymmetrical passband with a very sharp effective slope at the carrier end, thus ensuring a good bass response.

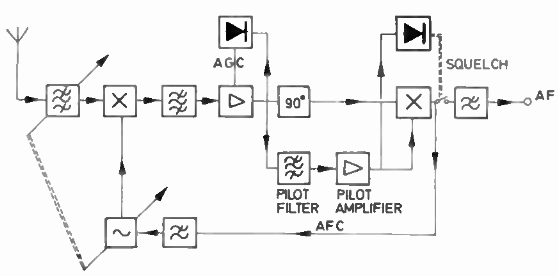
It is important that the demodulating waveform of the synchronous detector matches as closely as possible the frequency of the original carrier; it has been found that frequency errors in excess of about 10 Hz give rise to discernible deterioration in quality.[9] Thus both homodyne and synchrodyne methods of s.s.b. reception require a good degree of tuning accuracy, and many receiver designs have been published in recent years with the object of overcoming this problem.

4 Automatic Frequency Control Systems

If a large carrier (say 6 dB below peak envelope power) is transmitted with the signal, ease of tuning is fairly readily achieved by simple phase-lock systems. Philips at Eindhoven have demonstrated a homodyne receiver design in which the phase-shift associated with the carrier isolation filter is used indirectly to provide a control voltage.[10,11] Two such designs, taken from a paper by Hijdra, are shown in Fig. 4; consideration of the phase shifts in the two signal paths to the product detector shows that this control voltage will be zero only when the received carrier is exactly in the centre of the pilot filter passband. Unfortunately, if the capture range of the receiver is made large enough for tuning to be relatively foolproof, modulation frequencies become present in the loop, and quadrature distortion will result. This effect can be overcome by the use of a variable loop bandwidth controlled by the a.g.c. line.[12]

As the relative carrier power is reduced, the problem of finding the carrier in the incoming spectrum becomes more difficult, although direct phase-lock methods can still be applied with decreasing reliability. Hijdra's second receiver, shown in Fig. 4(b), is designed for carrier levels down to −20 dB. With partially or fully suppressed carrier, the use of a separate frequency reference in the receiver becomes desirable. This may be obtained from an external standard frequency receiver or a quartz crystal oscillator.[13] The former method presents numerous practical difficulties, and the alternative has been much more widely used in experimental designs.

Eden[1] suggested a design which has become the forerunner of a series of receiver designs using reference oscillators. In these receivers a pulse train is generated at the channel spacing frequency. Harmonics in the spectrum of this waveform will then appear at the carrier frequencies of received stations provided that transmitter frequencies are suitably allocated. Eden's design is for a double-conversion superheterodyne receiver in which the first local oscillator is locked to harmonics of



(a) S.s.b. receiver for carrier level of −6 dB (Hijdra)

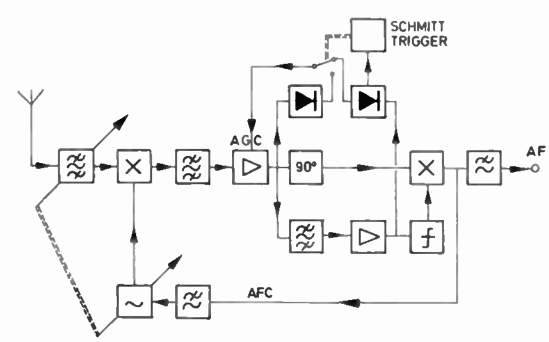


Fig. 4. (b) S.s.b. receiver for carrier level of −20 dB (Hijdra).

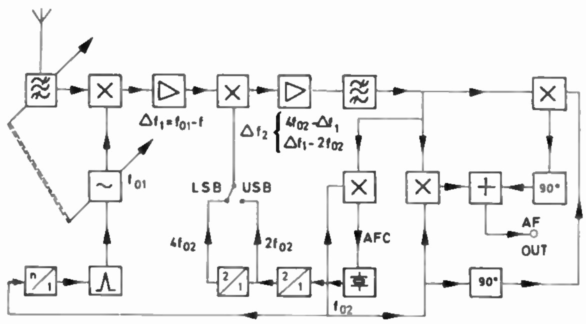
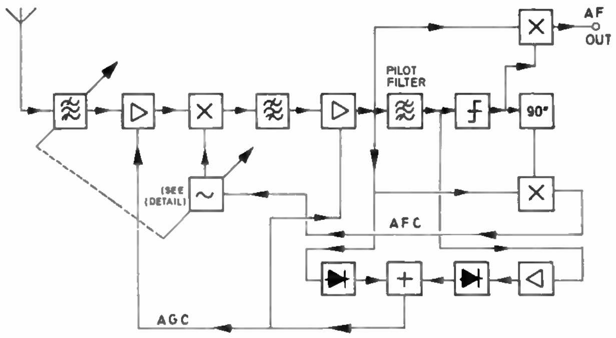


Fig. 5. Receiver using step-wise tuning (Eden).



(a) Block diagram of homodyne receiver (Bruch).

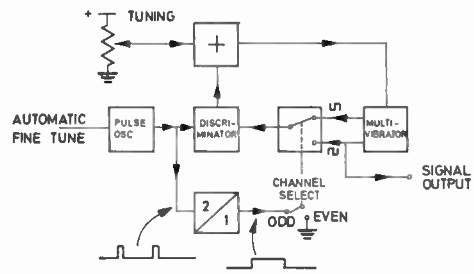


Fig. 7. (b) Detail showing operation of local oscillator tuning.

the pulse train (Fig. 5). In this way the local oscillator steps from frequency to frequency as the receiver is tuned. Selectivity is achieved by a bandpass r.f. filter and the undesired sideband is rejected using the phasing method.

A prototype receiver has been constructed by the authors[14] using many of the principles described by Eden (Fig. 6). The local oscillator is phase-locked to the pulse train by means of a sample-and-hold phase-lock loop, and the quadrature detector waveforms are obtained from the same logic circuit that produces the reference pulse train. Broad selectivity is obtained by means of a conventional i.f. transformer, and the shape of the overall passband is achieved using an audio low-pass filter. When this receiver is tuned, the output changes abruptly from station to station, unlike a conventional receiver where the transition is gradual.

A recent homodyne design by Bruch[7] also uses a phase-locked local oscillator, but here the mode of locking is somewhat different, and it is possible to arrange to lock either odd or even channels, the choice being made by a simple switch (Fig. 7). The reference frequency is derived from a precision delay line of the type used in colour television sets rather than a quartz crystal.

Another group of receivers that has been described also uses a pulse train at the channel spacing frequency, but these receivers use retroconversion to overcome oscillator drift rather than try to correct it. Two designs by the I.R.T. have been published; the first is due to Netzband.[13] In this receiver (Fig. 8) the incoming signal is converted to an intermediate frequency at which the desired station is selectively filtered, and is then converted back to its original frequency using the same local oscillator. In this way the effect of local oscillator drift is eliminated. The signal is detected using the reference spectrum, and the subjective effect of tuning is very similar to that of a conventional a.m. receiver.

The second receiver in this group is that described by Timmann,[15] the block diagram of which is shown in Fig. 9. This consists effectively of two parallel receivers; one converts the incoming signal to an intermediate frequency, while the other converts the reference spectrum using the same local oscillator. The converted reference spectrum is used to demodulate the converted signal, and in this way tuning whistles are eliminated.

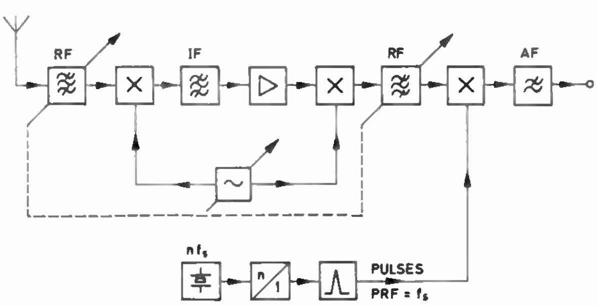


Fig. 8. Retroconversion receiver (Netzband).

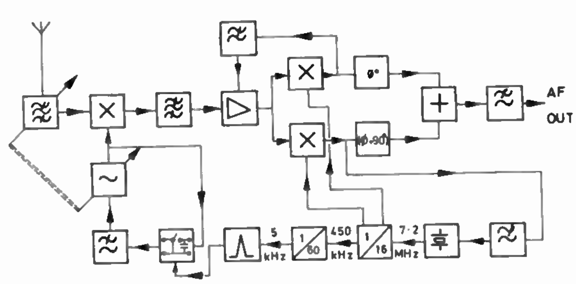


Fig. 6. Swansea receiver using step-wise tuning.

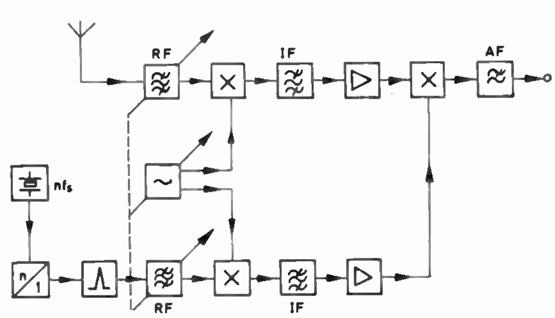


Fig. 9. Parallel conversion receiver (Timmann).

Before leaving the topic of tuning methods, it is worth commenting that with the coming of low-cost integrated circuits and display devices, the use of digital tuning methods in domestic receiving equipment may soon prove attractive. This would alleviate the problem of tuning in the congested high-frequency end of the m.w. band, for example, by giving a positive indication of received frequency or channel number. The use of complementary m.o.s. logic and liquid-crystal displays should reduce power consumption to the very low level necessary for portable use, and enable new cabinet styles to be adopted. The same approach to v.h.f. receivers could also be adopted if a proper manner of station frequency allocation was adopted.

5 Automatic Gain Control

Just as the use of s.s.b. would bring the need for an automatic frequency control system to aid unskilled operation, the use of an efficient automatic gain control is virtually obligatory in any domestic receiver. The transition to the new method of transmission will require a careful re-thinking of the problem of a.g.c. generation, especially with the introduction of new types of detector.

Conventional a.g.c. makes use of the steady voltage component appearing at the envelope detector, since for a.m. this is independent of modulation depth. With s.s.b., however, the mean signal level will change with modulation, even if a fairly large carrier component is radiated. Moreover, as we have seen, envelope detection will not be used for recovering the modulation, so that a special circuit will be necessary.

The final solution to be employed depends to a great extent upon exactly what level of carrier suppression is used. At a carrier level of -6 dB (with respect to peak envelope), the d.c. level in an envelope detector will vary by only about 1 dB, and the consequent distortion of dynamic range will not be apparent; hence, for example, Hijdra's receiver shown in Fig. 2(a) uses envelope derived a.g.c. Greater degrees of carrier suppression lead in turn to greater problems in deciding just how loud the output ought to sound. It is still possible to use the carrier level to control gain, but narrow-band filtering is necessary. The receiver designed by Hijdra for carrier suppressions of up to -20 dB (Fig. 4(b)) uses the envelope of the pilot signal as a source for the gain control signal; in a synchrodyne receiver the d.c. level appearing at the output of the in-phase detector provides the a.g.c. signal. When dealing with small carrier amplitudes such as this, however, it is necessary to be certain that a carrier is, in fact, being received, and Fig. 4(b) shows how the a.g.c. source can be switched by the presence of a carrier.

The carrier is not the only reference that could be used for a.g.c. but it is possibly the most convenient, since it raises no problems with compatibility. Thus Geluk, for example, has proposed the use of a sub-carrier system such as 'Lincompex',[16] and discussed the various advantages and disadvantages of such a system. If we restrict ourselves to the basic signal components, however, we may use the sideband power itself as a reference, which in practice means using audio-derived a.g.c. This is really much more suited to communications use, where constant audio output is not undesirable. The effect upon broadcast reception is greatly to compress the dynamic range, and to introduce some noise during programme intervals, although the latter effect could be minimized.

6 Conclusion

We have presented some, but by no means all, of the solutions to problems associated with the introduction of a new form of modulation into domestic use. In doing so, we have limited ourselves to the narrow context of receiver design and ignored some more wide-ranging technical points such as the optimum level of carrier suppression. The whole question of a change in modulation system must be answered not only from the engineer's point of view, but must also take into account the economics, politics and commercial pressures that would affect such a major transition. Whatever the outcome over the next few years, however, it is certain that these are times of change for domestic radio.

7 References

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Diversity reception for v.h.f. mobile radio

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SUMMARY

Multipath fading on v.h.f. mobile radio links is investigated, and aerial diversity reception considered for improved reception at the mobile terminal. A 3-channel adaptive array has been built using a conventional v.h.f.-a.m. receiver modified to measure signal amplitudes and relative phases. Experimental results show the effects of typical multipath environments, and the improvement obtained with an adaptive array.

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1 Introduction

Communications over v.h.f. land-based mobile radio nets, such as those used by police and other public services, frequently suffer 'break-up' of reception in urban, and to a lesser extent, wooded rural areas.[1,2] At v.h.f. and above, the direct wave from the transmitter is often not a major contributing component to the received field in these areas, and propagation is principally by way of scattering. Multiple reflexion and diffraction from buildings and other obstacles cause standing-wave patterns to be set up, because of the differing path lengths of the interacting waves. In moving through this environment the mobile receiver experiences a fading signal with amplitude variations of typically 20 dB. Minima in the standing wave pattern tend to occur at a separation of one half-wavelength, as would be expected if one considers the 'worst-case' situation for two interacting waves. This occurs when the sensing aerial is moved along the line of propagation of two waves travelling in opposite directions (as in ref. 3). The fading rate is thus dependent upon the transmission frequency and the speed of the vehicle, and at a vehicle speed of 15 m/s (34 mile/h) the fading rate is of the order of 0·1 Hz/MHz transmission frequency.

2 Effect of Multipath Fading

Amplitude modulated speech is the principal form of transmission on v.h.f. public service nets at the present time, and under fading conditions suffers multiplicative modulation by the fading waveform. The effect of rapid fading (hundreds fades/second upwards) is most easily visualized as a series of superimposed versions of the a.m. spectrum shifted up and down the frequency axis by the fading frequency and its harmonics. The latter may be quite numerous, owing to the steep 'drop-out' spikes characteristic of a fading waveform. Fading at a rate commensurate with the upper part of the speech band (above 1·5 kHz) is not particularly disastrous to intelligibility and the effect is something akin to tone interference at the fading frequency. As the fading rate falls into the 300 Hz to 1 kHz range (equivalent to the lower part of the speech band) severe distortion of the speech results, and if the fading depth is large, renders the information unintelligible. Sub-audio fading above about 50 Hz is moderately well tolerated by the human ear, as the variation of the speech spectrum becomes relatively small compared with its bandwidth. At slower fading rates, however, a secondary effect becomes dominant. In this range (a few fades/second), fading is more easily visualized as a 'keying' effect, and indeed sounds like it. At these rates the receiver a.g.c. system will respond, and for mild fading will attempt to provide a constant output. However, when the fades fall into the receiver noise level, e.g. in low signal areas, a.g.c. is unable to cope, and fading at word and syllabic rate occurs, giving poor results. It is found that fading rates of about 10 Hz are particularly bad, as the ear has an apparent capture effect for the fading frequency, the speech being totally unintelligible. As the fading rate reduces further, loss of one or more words occurs and repetitions of parts of the message are required.

At v.h.f. the fading rate falls in the 10 Hz range and hence produces 'worst-case' conditions of sub-audio fading. Under present operating conditions good and bad communication areas have to be learnt by the mobile users, so that they are able to move to a known 'good location' when they wish to communicate. This is an undesirable restriction as the base station can be out of contact with its mobile sub-stations at times, and the loss of command of the net becomes detrimental to the efficiency and effectiveness of the service.

3 Diversity Reception

It can be seen that the local field pattern varies very rapidly with displacement in the multipath environment, and at v.h.f. the distance between adjacent maxima and minima lies in the range 0·5 to 1 m. This is somewhat less than the length and width dimensions of the average vehicle, and hence a spatial array of aerials on the body of the radio vehicle might well provide uncorrelated samples of the field, enabling a significant improvement to be gained by suitable combination. Also for the simple standing-wave situation mentioned in Section 1, it was observed in ref. 3 that the standing-wave pattern of the magnetic field was spatially displaced by a quarter-wavelength from that of the electric field. Thus a combination of solely electric field-sensing and solely magnetic field-sensing aerials at one point on the vehicle body may also be of interest. Similarly a cross-polarized array of aerials might be used, exploiting the multiple scattering nature of the environment which produces incident waves with polarizations in many other directions as well as that initially transmitted. Since the standing-wave pattern is frequency dependent a third possible method of providing an improved signal at the mobile is the use of frequency diversity. However, spectrum space is already at a high premium, and the likelihood of more separate frequency channel allocations being provided to any mobile service is remote.

Diversity receiving systems of interest are thus limited to space, field and polarization diversity arrays, and can be broadly classed as aerial diversity schemes. Various signal processing techniques may be employed having obtained the diversity signals, and the three well-known combining techniques, selection, equal-gain, and maximal ratio, have been considered. Brennan[4] considered all three combining techniques in a Rayleigh fading environment, assuming uncorrelated inputs with noises of constant equal mean power, and showed that when the number of inputs is small, there is little to choose between the techniques as far as their ability to improve the average signal/noise ratio (s.n.r.) is concerned. Equal-gain is only marginally inferior to maximal ratio combining (< 1 dB different) for up to eight inputs, and selection is less than 3 dB different with up to four inputs. However, all diversity combiners provide an acceptable output for a much greater percentage of the time than a single channel and this is illustrated in Fig. 1 which shows cumulative probability distributions for a three-channel diversity system under Rayleigh fading conditions. This is the desirable feature of diversity as far as mobile radio is concerned, the small differences in average s.n.r. being much less important. Real input signals may not in fact correspond very well with Brennan's assumptions, although he points out that nearly ideal diversity improvement is still obtained with input correlation factors of 0·3, and indeed much of the improvement can be obtained with input correlations as high as 0·7. Considering the design aspects of the various systems it is apparent that the simplicity of the equal-gain combiner is hard to match, although the squaring technique of ref. 5 provides a simple maximum ratio combiner, providing Brennan's assumptions apply to the inputs.

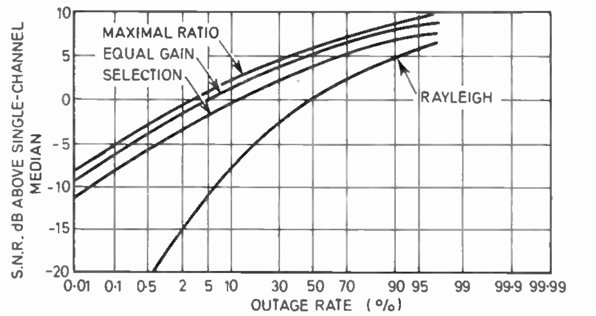


Fig. 1. Probability distribution functions for three-channel diversity systems.

4 Statistical Field Theory

Of the mathematical models proposed to represent the multipath field, the most complete is probably that of Clarke,[6] in which the field at any point is composed of N equal amplitude, vertically polarized plane waves, the nth wave arriving at an angle α_n to an arbitrary reference axis. The phase and angle of arrival of each component wave are assumed statistically independent and uniformly distributed through 2π radians.

The envelope of the incident field is then shown to be Rayleigh distributed, and its phase uniformly distributed through 2π radians. The normalized spatial auto-correlation function of the electric field is given by a zero-order Bessel function of the first kind

$$\bar{R}_E(L) = J_0(2\pi L/\lambda)$$

and Clarke also shows that the equivalent correlation function for the detected output of this received signal $\bar{R}_{\delta E}(L)$, is approximately equal to the square of the Bessel function defined above. Hence the normalized auto-correlation function of the envelope of the electric field and of its detected output are given by

$$\bar{R}_{\delta|E|}(L) \simeq J_0^2(2\pi L/\lambda)$$

In a similar manner the normalized auto-correlation function of the envelope of the signal received and that of the detected output for a pair of orthogonal magnetic field-sensing aerials is given by[7]

$$\bar{R}_{\delta|H|}(L) \simeq J_0^2(2\pi L/\lambda) + J_2^2(2\pi L/\lambda)$$

where $J_2(\ldots)$ is a second-order Bessel function of the first kind. Also the normalized cross-correlation function of the envelopes of the electric and magnetic fields and of their detected output is given by

$$\bar{R}_{\delta|E||H|}(L) \simeq J_1^2(2\pi L/\lambda)$$

where $J_1(\ldots)$ is a first-order Bessel function of the first kind. These output functions are plotted in Fig. 2.

It is observed that $\bar{R}_{\delta|E|}(L)$ decreases rapidly with displacement, reaching a value of 0·3 at approximately one quarter-wavelength displacement, and not exceeding a value of 0·15 at displacements greater than 0·3 wavelengths. Thus signals obtained from electric field-sensing aerials displaced by not less than a quarter-wavelength, should be substantially uncorrelated and enable the near optimum improvement expected of a diversity combining system to be achieved. Rustako[8] built and tested a diversity system and showed that little difference in performance resulted when using aerials spaced by $\frac{1}{4}$, $\frac{3}{4}$, or $\frac{5}{4}$ wavelengths. Inspection of $\bar{R}_{\delta|H|}(L)$ shows a slightly higher correlation than $\bar{R}_{\delta|E|}(L)$, but still falls within the Brennan criterion with displacements greater than about a quarter-wavelength, and thus a magnetic field-sensing aerial array should not give significantly different results from those of a comparable electric field-sensing array. At zero displacement $\bar{R}_{\delta|E||H|}(L)$ is equal to zero, and the electric and magnetic fields are independent (in a time sense). Thus a combination of electric and magnetic field-sensing aerials at a point, should provide ideal signals for a diversity combiner. The mathematics of polarization diversity in the scattering environment has not been considered by the authors, but an investigation of polarization diversity reception from a circularly polarized transmission has been reported by Lee and Yeh.[9] They assume negligible cross-talk between their orthogonally polarized transmissions, but this is not in accord with our experience.

5 Diversity Combining Techniques

In the past a number of diversity combining systems have been described for mobile radio use. The simplest of them in concept is the post-detection combining receiver, consisting of a separate r.f., i.f., and detector for each channel, the audio outputs from which are summed into a common final stage. Various i.f. combining receivers have been proposed in an attempt to improve performance and cut down on the amount of equipment required. These are generally applications of two well-known signal co-phasing techniques, one employing a

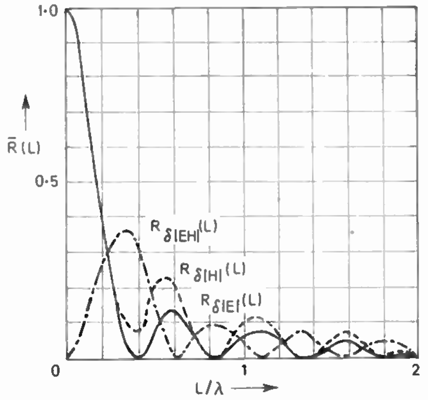


Fig. 2. Theoretical normalized correlation functions for the envelopes of isotropically scattered waves.

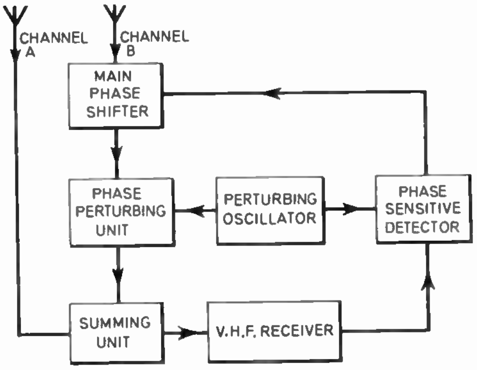


Fig. 3. Schematic diagram of the self-phasing aerial system.

multiple-heterodyning principle[10] and the other, a squaring technique requiring an additional pilot-tone carrier to be present just outside the modulation band.[5]

However, all the i.f. systems require an entirely new design of receiver, and are not readily adaptable to existing single-receiver systems. A large capital investment has already been made by the users of these present systems, and to be economically viable a relatively cheap 'add-on box' system using the existing receiver is required. In this paper an r.f. pre-receiver combining system will be described, which uses a standard mobile a.m. receiver. A similar system designed for use on f.m. links has been described previously by the authors.[11] In order to optimize the design parameters for such a system it has been adapted to measure and record the received signals, so that the nature of the multi-path field may be investigated in more detail.

6 A Self-phasing Aerial Array

This system is designed to co-phase a number of aerial signals and add them together, passing the sum into a conventional a.m. receiver. Operation of the array is based on a phase-perturbing technique devised by Lewin,[12] which permits detection of the phase error between each aerial signal and the sum signal. The information is extracted at the receiver detector and used to drive control loops designed to minimize these errors.

The principle of operation is demonstrated in Fig. 3 for a simple two-aerial system. One aerial signal (channel A) is considered the reference signal, and is connected directly to the summing unit. The other (channel B) is passed to a phase-shifter capable of changing the phase through 2π radians. The output is passed to a phase-perturbing unit, which is a small-deviation phase modulator and provides the phase-difference information to set the main phase-shifter. This signal then enters the summing unit, the output of which feeds the receiver. Figure 4 shows carrier vector diagrams illustrating how the control information is derived. It can be seen that amplitude modulation at the perturbing frequency is produced in the sum signal. This modulation is in phase with the perturbing frequency when signal B lags signal A and the sum signal (Fig. 4(a)), and in antiphase when signal B leads signal A and the sum signal (Fig. 4(b)). When signal B

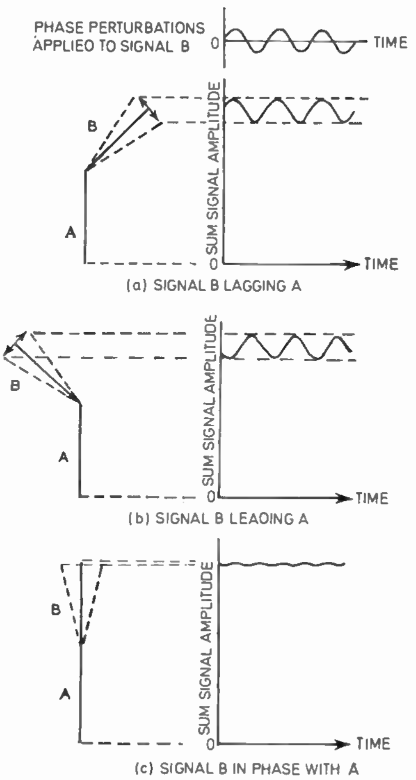


Fig. 4. Derivation of the control signal by phase perturbation.

is in phase with signal A (Fig. 4(c)) only a small amplitude second harmonic of the perturbing frequency is produced. The a.m. error signal is extracted from the detector of the receiver and phase detected with respect to the phase perturbing oscillator. The polarity of the filtered d.c. output thus obtained indicates the direction in which the main phase-shifter must be changed to co-phase the signals A and B. The method is readily extendable to any number of signals, if a separate control loop is provided for each additional aerial and a separate perturbing frequency is used. The control output from the detector of such an N-channel system is approximately proportional to

$$\sum_{r=1}^{N} \sin \theta_{er} \sin \omega_r t,$$

where θ_{er} is the phase error angle between the rth signal and the sum signal, and ω_r is the perturbing frequency used on that channel.[7] Multiple demodulation of this signal by product detection with the appropriate driving oscillator then yields N d.c. control signals. It is observed that a zero-value error signal is also obtained when a particular input signal is in anti-phase with the sum, but since this is a position of unstable equilibrium and any perturbation drives the system away from this condition, it is not of practical consequence. Location of the perturbing frequencies is an important factor, as they must be within the i.f. pass-band of the receiver, but outside the a.f. bandwidth. The control frequencies form a.m. sub-carriers of the transmitted a.m. and for 3 kHz speech bandwidth must be selected to be at least 6 kHz to avoid overlapping of side-bands. The lower limit of transmitted speech is normally restricted to 300 Hz and this is the bandwidth available for location of all the control frequencies if overlap with the side-bands of the other control frequencies is to be avoided. This factor is the fundamental limitation on the speed of phase correction of the system, as the d.c. control voltages must be filtered from the difference-frequency signals which also appear at the phase-detector outputs. The total i.f. bandwidth required for this configuration is somewhat large, and requires 25 kHz frequency-channel allocations to be operated. Increasing use of 12·5 kHz channel spacing allows a receiver i.f. pass-band of not more than ± 4 kHz, and a reduced perturbing frequency will have to be considered. The control frequency error signal is always less than 4% of carrier level (using 0·2 radian phase perturbation), and is even smaller in the near co-phased state, so that the control frequency side-bands are at a very low level compared with the main speech side-band. It is therefore thought that location of the control frequencies just above 3 kHz may not significantly affect the intelligibility of the received speech, providing sharp filtering of the audio output is employed.

The number of channels which will provide an economic diversity improvement is not likely to be greater than four, and probably only two or three. Figure 5 shows that the theoretical improvement in average s.n.r. above one channel for two aerials is 11 dB for 99% of the time (i.e. 1% outage rate), but only a further 4·5 dB and then 2·2 dB for three and four aerials respectively.[4] Also with restricted processing bandwidth, the more control signals used, the slower each channel can co-phase because of the reduced rise-time of the d.c. control signal filter, which is designed to exclude control frequency difference components. Thus for mobile radio, a fast-acting two-channel system may not give a performance which is significantly inferior to that of a slower multi-channel system.

Viewing diversity in this light illustrates the 'weak-signal' improvement that might be expected in an existing or proposed service and suggests an increased service-area capability or the possibility of reducing the transmitter power, maintaining the existing or proposed service-area. Alternatively the reduction in outage-rate obtainable may be considered, and this is important to designers of services required to handle data transmissions. These are usually far less redundant than speech transmissions and therefore require a system with a low outage-rate to provide a good service. Referring

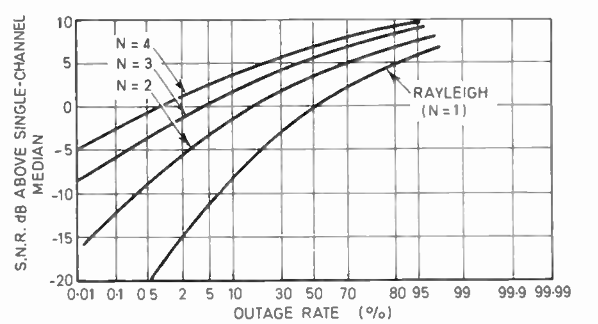


Fig. 5. Probability distribution functions for equal-gain diversity systems.

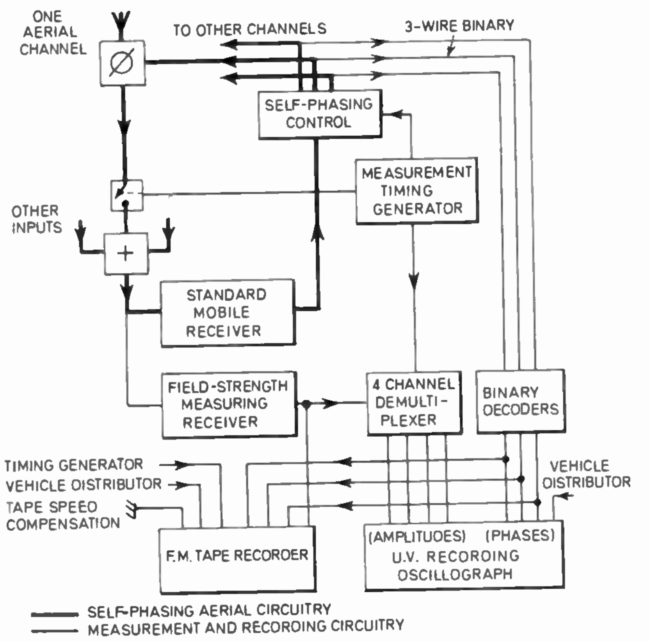


Fig. 6. Self-phasing aerial array with measurement and recording system.

to Fig. 5 again, it can be seen that whereas a conventional system which is acceptable with a s.n.r. down to 8 dB below the mean, has an outage rate of 10% in a Rayleigh fading environment, the use of two-channel diversity would reduce it to 0·8%, and three-channel diversity to only 0·03%.

The equipment built at the University of Birmingham consists of a three-channel self-phasing array with an additional non-phase-adjusting channel. The summing unit is active, enabling isolation of any number of channels for test purposes, although in an engineered system it would probably be wiser to use a passive summing method (e.g. hybrid junction). The main phase-shifters are quantized into eight $\pi/4$ radian steps, and are driven by a bi-directional ring-counter running from a fixed clock. The d.c. error signals, integrated over a clock-period for higher noise immunity, are used to determine the direction of count at each clock-pulse, so that the system 'hill-climbs' towards the optimum phase-shifter positions, and in the steady field state, each channel oscillates about the two-phase-shifter positions either side of the actual phase correction required.

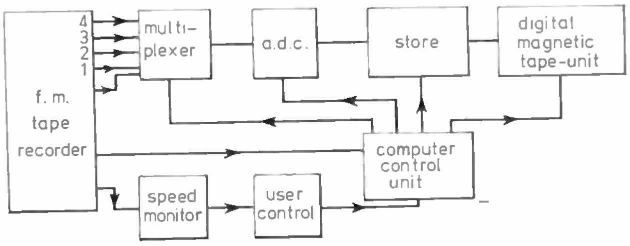
7 Measurement Scheme

The system described in the previous Section has been modified to facilitate measurement and recording of the signal strengths from each aerial element, their sum, and the relative phase between them. A diagram illustrating the modified system is given in Fig. 6.

As with all 'random' problems, results are most usefully analysed on a statistical basis, and hence a prime requirement is a signal record suitable for analogue-to-digital conversion, and subsequent analysis with the aid of a digital computer. However, it is also instructive to have a visual record of some results in order to pick out general trends and special peculiarities in areas of interest, and hence two multi-channel recording equipments are used. An f.m. tape recorder gives a complete instrumentation record for statistical analysis, and an ultra-violet (u.v.) recording oscillograph gives periodic visual output when required.

Signal amplitude measurements are made sequentially by switching the appropriate input channels at the summing unit, the latter being modified to feed both the existing receiver and a field-strength measuring receiver which has a logarithmic response. The measuring receiver output is passed direct to one channel of the tape recorder and also de-multiplexed for display on the u.v. recorder. Phase-shifter positions are monitored from the bi-directional counter, the 3-wire binary code (representing the number of $\pi/4$ radians inserted phase-shift steps) being converted to a single-wire eight-level code before display on the u.v. recorder, and continuous recording on the tape. Timing waveform and vehicle distributor pulses are also recorded, the former for computer synchronization of the multiplexer and analogue-to-digital converter to the tape-recorded signals and the latter for correlation of recorded time with distance travelled in the vehicle.

Analogue-to-digital conversion of the recorded data is achieved using a PDP–9 computer, incorporating a 12-bit a/d converter with a four-channel multiplexer, on an 'on-line' basis. Figure 7(a) shows the tape recorder coupled to the computer so that the four tracks containing the phase and amplitude information are fed to



(a) Block diagram of digitizing system.

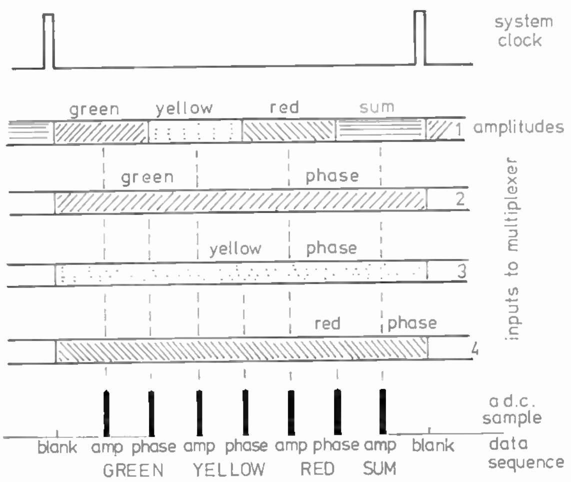


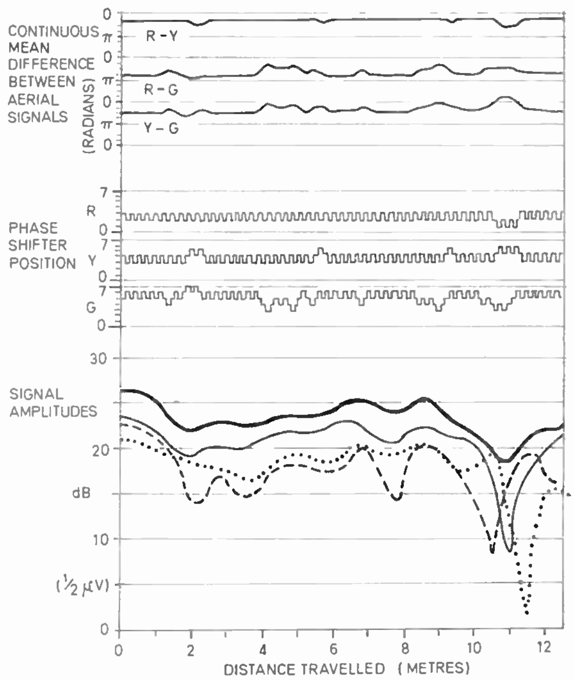
Fig. 7. (b) Digitized data sequence showing multiplexer linking tape-recorder tracks to a.d.c.

the multiplexer, with tape flutter compensation applied to the channel containing amplitude information. The computer central processor controls the multiplexer switching times and digitization timing from clock-pulses received from the timing track of the tape recorder. A speed monitor enables the operator to have overriding control of the digitization process if any vehicle speed fluctuations occurred during the original recording. A measuring sequence taking eight samples in every clock-period has been devised and the signal amplitudes are read off one tape channel, interleaved time-wise with the phase-shifter measurements from the other three channels as shown in Fig. 7(b). Since the amplitude information is already multiplexed and has to be read during specific sections of each clock-period, whilst the phase information on each channel remains static for the duration of a clock-period, this interleaving process provides the optimum settling time for the multiplexed measuring receiver signals, and the inclusion of a blank dummy reading permits a convenient sequence of pairs of amplitude and phase readings for each channel to be maintained. The dummy reading occurs at the time of a system clock pulse and is set as a zero-valued digital word, thereby maintaining individual word identification in the data sequence. The digitized data are collected from the a/d converter in the computer store, and periodically transferred to incremental digital magnetic tape for permanent storage. This is later transferred via a PDP-8 satellite computer to the University main KDF-9 computer. This machine provides adequate rapidly accessible bulk storage, and suitable processing facilities for the statistical analysis of data.

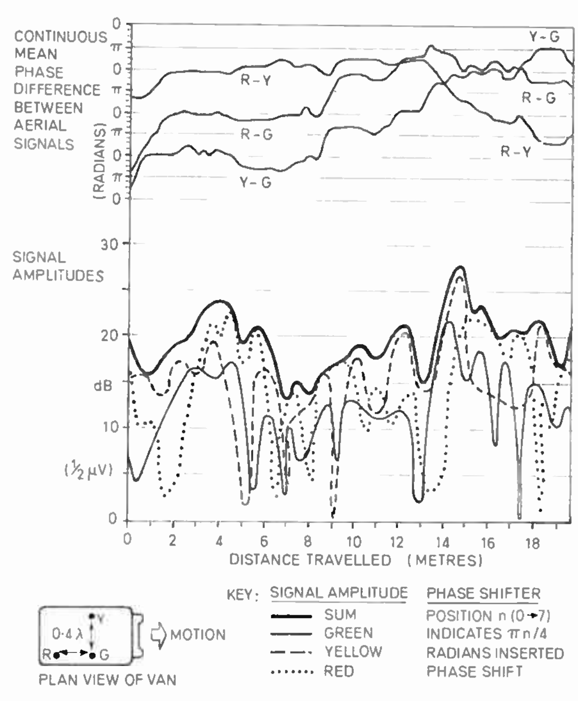
8 Results

A set of experiments has been conducted under differing propagation conditions in the Birmingham area, using the self-phasing array coupled to three quarter-wavelength monopole aerials. These were positioned at three corners of a square, of side 0·4 wavelengths, on the roof of a 15-cwt Bedford Utility van, in such a way that two aerials approximately straddled the width of the van halfway down its length, and the third was positioned at the rear off-side corner.

Some representative u.v. recorder output obtained during these tests appears in Fig. 8. Figure 8(a) illustrates fairly good reception conditions when a strong component of the direct wave from the transmitter is present. The sum signal is not appreciably greater than the best individual signal, although fades are marginally less deep. The change in relative phase between the signals is small, and the constancy of the *R*-channel phase-shifter (oscillating back and forth between the two phase-shifter positions nearest the correct phase-shift required) shows that a control loop is not necessary in every channel of the system. Removal of the adaptive circuitry associated with one channel would merely fix the optimum sum-signal in-phase with that channel, and the remaining channels would co-phase to satisfy this condition. However, the system clock-rate would then have to be increased to maintain the phase compensation rate, and observation of the one acute fading trough in the figure reveals circumstances in which all the phase-shifters have to change position to follow the rapid changes of relative phase between the aerial signals. This fade occurred on the approach to a very tall building, and each aerial is seen to pass through it in sequence,



(a) University of Birmingham Ring Road, grid reference SP048848 (Transmitter location 50° to direction of motion of vehicle, distance 450 m.)



(b) Lionel Street, Birmingham, grid reference SP066873. (Transmitter location 10° to direction of motion of vehicle, distance 14 km.)

Fig. 8. Samples of u.v. recording oscillograph output showing channel amplitudes, phase shifter positions and computed mean phase difference between aerial signals.

indicating a null channel formed obliquely across the road. At this point the sum signal also drops, but only by about 7 dB as opposed to the 12–18 dB of the individual aerial signals, and the rate of fall is much less than that of the individual signals. If the aerial spacing were increased, this particular fade would have been reduced still further in the sum signal, but it is not expected that a significant improvement would be provided on average.

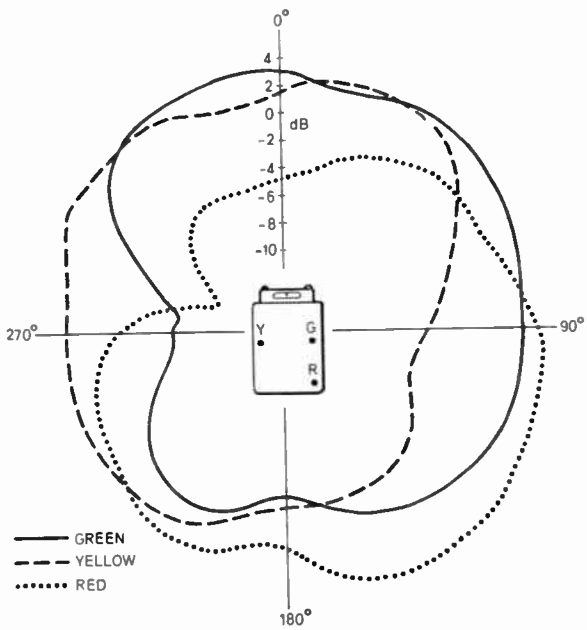


Fig. 9. Horizontal radiation pattern of each $\lambda/4$ aerial, in the presence of the others, on the van roof.

In contrast, Fig. 8(b) is taken from a severe multi-path area in the centre of Birmingham, and shows individual aerial signal strengths with many rapid and deep fades, which drop below receiver sensitivity into noise. Peak to trough signal variations are as high as 27 dB, but the maximum deviation of the sum signal is only 15 dB, and is well below half that of any individual channel on average. Also the fading rate is greatly reduced, and the sum signal never 'drops-out' into the receiver noise. Conventional receiver a.g.c. is thus able to smooth-out the fading effects in the sum signal to a much greater degree than that from a single aerial. It is also observed that the rate of change of phase at the aerial terminals is quite rapid (up to 4π radians/wavelength on average), and this is consistent with the expected maximum fading rate mentioned in Section 1. The importance of the maximum phase-gradient between elements is that it determines the rate at which the self-phasing array must adapt, in order to follow the fades at a given vehicle speed. The system used in these tests was operated at a clock-pulse rate of 50 Hz which allows operation at full efficiency under extreme multi-path conditions with vehicle speeds up to 18 m/s (40 miles/hour). Higher vehicle speeds can be tolerated in areas where the multi-path conditions are not so severe.

Assuming that the received fields are perfectly stationary, it would be expected that if the test vehicle followed a straight line at constant speed through the field, then the R-channel signal would correspond to a 0·4 wavelength shifted replica of that on the G-channel.

However, from Fig. 8(b) it can be seen that although the deep fades in these signals follow the expected pattern there are noticeable differences in the fine detail. Some variation might be expected as the course taken by the van deviated to avoid stationary vehicles, and the field was also disturbed by the passing of high-sided metal-clad vehicles which were numerous in the area.

However, an additional influencing factor is that the horizontal radiation patterns of the aerials are not perfectly omnidirectional, and measurements of these patterns has shown that a variation of up to 4 dB from the mean value can be caused by the vehicle body alone. The effect of the other elements in the array (which act as terminated parasites since the inputs are isolated at the summing unit), is to cause a bulk shift of the individual patterns away from the inserted parasitic element. This shift is about 2–3 dB and means that each element exhibits a slightly directional response favouring an outward direction from the vehicle, as shown in Fig. 9. This is not likely to degrade the diversity performance in an urban environment where waves can arrive from any direction, nor will it affect rural reception since in no direction is there a minimum response in all the aerial radiation patterns. Lee[13] has also shown that mutual coupling effects are insignificant in maximal ratio combiners with small numbers of inputs.

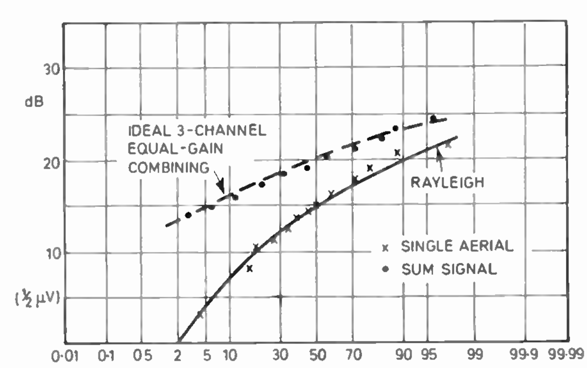


Fig. 10. Experimental probability distribution functions in an urban area.

To illustrate the effectiveness of the self-phasing array as a diversity combiner, the amplitude information from Fig. 8(b) has been presented in the form of the cumulative probability distributions shown in Fig. 10. It can be seen that the individual aerial signals (the R-channel signal has been plotted here) follow a close approximation to a Rayleigh distribution, whilst the sum signal is never more than 1 dB away from the ideal three-channel equal-gain combining curve given by Brennan. This is a very satisfactory result. The graphs show that the 50% cumulative distribution level of the sum signal is about 5 dB greater than that for one channel, and the 90% level is about 9 dB greater. These are almost exactly equal to the theoretical values.

9 Conclusions

Analysis has shown that Rayleigh fading conditions exist in practice over small areas, and where such conditions do exist the three-element self-phasing array performs in a near optimum manner, and provides a significant improvement in received signal. Although the mean signal will vary over longer runs, the array is expected to perform adequately in its present form under all normal conditions. The final choice of the number of diversity channels will depend upon the particular application, but will probably be best defined in terms of outage-rate. Subjective tests using speech transmissions have indicated that three-channel diversity is little better than two-channel; however, when the nature of the message is such that a very low outage-rate is necessary (e.g. data transmissions), more diversity branches may be desirable.

Field diversity has not yet been employed, but in principle the elements of an 'energy-density' aerial[3] will provide signals suitable for use with the self-phasing array.

10 Acknowledgments

The authors would like to thank the British Broadcasting Corporation Research Department for the loan of a field-strength measuring receiver, and for providing facilities for the measurement of the aerial radiation patterns. Thanks are also due to the staff of the U.K. Home Office Wireless Depot, Romsley, for providing test transmissions.

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Sensitivity of narrowband cascaded bandpass filters

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SUMMARY

The sensitivity of the magnitude transfer function of narrowband cascaded filters is evaluated in terms of pole sensitivities. It is shown that Q sensitivity is often negligible in determining performance and that the worst-case sensitivity of the magnitude transfer function exhibits maxima at particular frequencies and that Q-factor and ω_0 sensitivity are the dominant factors determining the sensitivity of the magnitude transfer function. The worst-case sensitivity of the cascade-realization is compared with that of the corresponding ladder realization.

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1 Introduction

The realization of high Q second-order inductorless transfer functions is a frequently encountered requirement. The primary purpose of this contribution is to develop a relationship between the often-quoted pole sensitivities and the magnitude transfer function sensitivity. Then, it will be shown that the worst-case sensitivity of a high Q all-pole section exhibits maxima at the half-power frequencies and that the worst-case sensitivity curve may be obtained for higher order cascaded bandpass structures. The worst-case sensitivity of the magnitude transfer function is then compared for cascade- and ladder-structure realizations of bandpass filters.

2 Q Factor and ω_0 Sensitivity

The transfer function

$$KH(s) = \frac{K\frac{\omega_n}{Q}s}{s^2+\frac{\omega_n}{Q}s+\omega_n^2}, \qquad K = \text{constant} \qquad (1)$$

may be realized by employing any one of a large number of available synthesis techniques and high order *narrowband* functions realized by cascading such sections. For each section, $Q \gg 1$. The literature contains many references to the sensitivities of Q and ω_0 to elements x_i; that is,

$$S_{x_i}^{Q} \equiv \frac{\mathrm{d}Q/Q}{\mathrm{d}x_i/x_i} \quad \text{and} \quad S_{x_i}^{\omega_0} \equiv \frac{\mathrm{d}\omega_0/\omega_0}{\mathrm{d}x_i/x_i} \qquad (2)$$

and it is not difficult for the network designer to obtain these sensitivities for Linvill-n.i.c.-, inductance-simulation-, positive-feedback- and Sallen-Key-methods of synthesis. In this contribution, we are concerned with the sensitivities of the narrowband transfer function $H(\mathrm{j}\omega)$ to the parameters Q and ω_0 because it is the *transfer function* that is of ultimate interest.

First, it will be shown that there exists an exact relationship between $S_{\omega_0}^{H(s)}$ and $S_{Q}^{H(s)}$ and then, for the narrowband case, that the latter sensitivity term is negligible over the frequency range of interest. From equation (1),

$$H^{-1}(s) = Q\left[\frac{s}{\omega_0}+\frac{\omega_0}{s}\right]+1 \qquad (3)$$

giving

$$\frac{Q\,\mathrm{d}H^{-1}(s)}{\mathrm{d}Q} = Q\left[\frac{s}{\omega_0}+\frac{\omega_0}{s}\right] \qquad (4)$$

and

$$\frac{\omega_0\,\mathrm{d}H^{-1}(s)}{\mathrm{d}\omega_0} = Q\left[\frac{\omega_0}{s}-\frac{s}{\omega_0}\right]. \qquad (5)$$

Now,

$$S_Q^{H(s)} = -S_Q^{H^{-1}(s)} = -\frac{\mathrm{d}H^{-1}(s)}{\mathrm{d}Q}\cdot\frac{Q}{H^{-1}(s)} \qquad (6)$$

and

$$S_{\omega_0}^{H(s)} = -S_{\omega_0}^{H^{-1}(s)} = -\frac{\mathrm{d}H^{-1}(s)}{\mathrm{d}\omega_0}\cdot\frac{\omega_0}{H^{-1}(s)} \qquad (7)$$

so that, from (4), (6) and (7)

$$\frac{S_Q^{H(s)}}{S_{\omega_0}^{H(s)}} = \frac{\dfrac{s}{\omega_0}+\dfrac{\omega_0}{s}}{\dfrac{\omega_0}{s}-\dfrac{s}{\omega_0}} = \left[\frac{s^2+\omega_0^2}{\omega_0^2-s^2}\right] \tag{8}$$

From equation (1), $H(s)$ is only a function of Q and ω_0 where it is understood that Q and ω_0 are in general functions of network parameters x_i. Thus,

$$\begin{aligned} S_{x_i}^{H(s)} &\equiv \frac{\mathrm{d}H(s)}{\mathrm{d}x_i}\cdot\frac{x_i}{H(s)} \\ &= \left[\frac{\partial H(s)}{\partial\omega_0}\cdot\frac{\mathrm{d}\omega_0}{\mathrm{d}x_i}+\frac{\partial H(s)}{\partial Q}\cdot\frac{\mathrm{d}Q}{\mathrm{d}x_i}\right]\frac{x_i}{H(s)} \\ &= \left[\frac{\partial H(s)}{\partial\omega_0}\cdot\frac{\omega_0}{H(s)}\right]\left[\frac{x_i}{\omega_0}\cdot\frac{\mathrm{d}\omega_0}{\mathrm{d}x_i}\right] \\ &\qquad + \left[\frac{\partial H(s)}{\partial Q}\cdot\frac{Q}{H(s)}\right]\left[\frac{x_i}{Q}\cdot\frac{\partial Q}{\partial x_i}\right] \end{aligned}$$

giving, directly from the definition of sensitivity,

$$S_{x_i}^{H(s)} = [S_{\omega_0}^{H(s)}.S_{x_i}^{\omega_0}]+[S_Q^{H(s)}.S_{x_i}^{Q}] \tag{9}$$

Then, substituting (8) into (9) gives

$$S_{x_i}^{H(s)} = S_{\omega_0}^{H(s)}\left[S_{x_i}^{\omega_0}-\left(\frac{s^2+\omega_0^2}{s^2-\omega_0^2}\right)\cdot S_{x_i}^{Q}\right] \tag{10}$$

and is an exact result for the bandpass function of equation (1). Substituting $s = \mathrm{j}\omega$,

$$S_{x_i}^{H(\mathrm{j}\omega)} = S_{\omega_0}^{H(\mathrm{j}\omega)}\left[S_{x_i}^{\omega_0}+\left(\frac{\omega_0-\omega^2}{\omega_0^2+\omega^2}\right)S_{x_i}^{Q}\right] \tag{11}$$

where the term inside the square-brackets is a function of the realization technique that is employed.

Consider now the realization of *narrowband* low-sensitivity transfer functions of the type that employ the biquad[1,2,3], the gyrator[4,5] or the g.i.c.[6,7] in which a transfer function of the form of equation (1) results in sensitivities $S_{x_i}^{\omega_0}$ and $S_{x_i}^{Q}$ that are either constants *of the same order of magnitude* or are equal to zero. This class of networks is now widely employed to realize high-Q bandpass functions. Now, for the narrowband case, the magnitude of the function $(\omega_0^2-\omega^2)/(\omega_0^2+\omega^2)$ in equation (11) is much less than unity over the region of ω that is generally of interest; that is, we may assume

$$|\omega_0^2-\omega^2| \ll \omega^2$$

so that, if $S_{x_i}^{\omega_0}$ and $S_{x_i}^{Q}$ are of the same order of magnitude, it follows from equation (11) that

$$\sum_{\text{all } x_i} |S_{x_i}^{|H(\mathrm{j}\omega)|}| \simeq |S_{\omega_0}^{|H(\mathrm{j}\omega)|}|\,.\sum_{\text{all } x_i}|S_{x_i}^{\omega_0}| \tag{12}$$

Now, from equation (5) it follows that

$$S_{\omega_0}^{H(s)} = \frac{Q\left[\dfrac{\omega_0}{s}-\dfrac{s}{\omega_0}\right]}{Q\left[\dfrac{s}{\omega_0}+\dfrac{\omega_0}{s}\right]+1} \tag{13}$$

and, since[8]

$$S_{\omega_0}^{|H(\mathrm{j}\omega)|} = \mathrm{Ev}\,[S_{\omega_0}^{H(\mathrm{j}\omega)}] \tag{14}$$

it follows from (13) and (14) that

$$S_{\omega_0}^{|H(\mathrm{j}\omega)|} = \frac{Q^2\left[\dfrac{\omega_0}{\omega}+\dfrac{\omega}{\omega_0}\right]\left[\dfrac{\omega}{\omega_0}-\dfrac{\omega_0}{\omega}\right]}{Q^2\left[\dfrac{\omega_0}{\omega}-\dfrac{\omega}{\omega_0}\right]^2+1} \tag{15}$$

which is an exact result. The narrowband assumption allows simplification because

$$\frac{\omega_0}{\omega}+\frac{\omega}{\omega_0} \simeq 2$$

and

$$\frac{\omega}{\omega_0}-\frac{\omega_0}{\omega} \simeq 2\frac{\Delta\omega}{\omega_0}, \qquad \text{where} \qquad \Delta\omega \equiv \omega-\omega_0$$

giving

$$S_{\omega_0}^{|H(\mathrm{j}\omega)|} \simeq \frac{(2Q)^2\left[\dfrac{\Delta\omega}{\omega_0}\right]}{1+(2Q)^2\left[\dfrac{\Delta\omega}{\omega_0}\right]^2} \tag{16}$$

Substituting equation (16), which is only a function of the required transfer function, into equation (12) gives

$$\sum_{\text{all } x_i}\left|S_{x_i}^{|H(\mathrm{j}\omega)|}\right| \simeq \frac{(2Q)^2\left[\dfrac{|\Delta\omega|}{\omega_0}\right]}{1+(2Q)^2\left[\dfrac{\Delta\omega}{\omega_0}\right]^2}\cdot\sum_{\text{all } x_i}|S_{x_i}^{\omega_0}| \tag{17}$$

This worst-case sensitivity function for a high-Q section of the form described in equation (1) is an important result because the term

$$\sum_{\text{all } x_i}|S_{x_i}^{\omega_0}|$$

is usually known and for the biquad, gyrator and g.i.c. realizations is equal to 3. Realization by one of these methods is assumed in the remainder of this contribution. Equation (17) is given in Fig. 1 and its usefulness in determining the worst-case performance of cascaded structures will be explained. Note that the worst-case sensitivity of the single section of Fig. 1 exhibits maxima at the half-power frequencies $\omega_0\pm\omega_0/2Q$ and the peak sensitivity is $3Q$. A computed *exact* curve, for $Q = 10$, is given in Fig. 1 to illustrate the validity of the approximations used in the derivation of equation (17). An important conclusion is that the narrow-band sensitivity performance is virtually independent of $S_{x_i}^{Q}$ and is

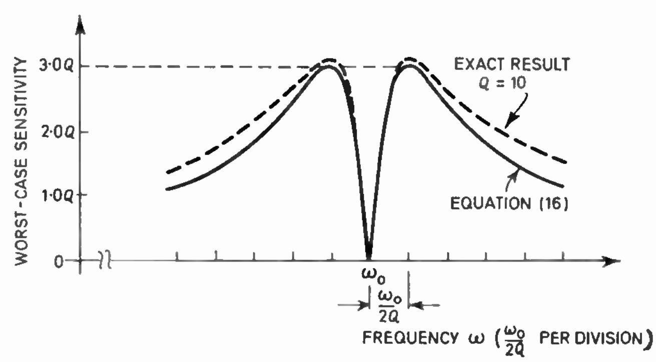


Fig. 1. Worst-case sensitivity of a single second-order section.

determined by $S_{x_i}^{\omega_0}$. This latter term is constant and low-valued for all the above mentioned low-sensitivity methods.

3 Cascaded Low-sensitivity Sections

Consider now the cascade realization of an Nth order narrowband transfer function of the form

$$T(s) = \prod_{i=1}^{N/2} \frac{\omega_{0i}s}{\left(s+\dfrac{\omega_{0i}}{Q_i}+j\omega_{0i}\right)\left(s+\dfrac{\omega_{0i}}{Q}-j\omega_{0i}\right)} \tag{18}$$

where all $Q_i \gg 1$. It follows from equation (17) that the worst-case sensitivity is

$$\sum |S_{x_i}^{|T(j\omega)|}| \simeq 3\left|\sum_{i=1}^{N/2} \frac{(2Q_i)^2\left[\pm\dfrac{(\omega-\omega_{0i})}{\omega_{0i}}\right]}{1+(2Q_i)^2\left[\dfrac{\omega-\omega_{0i}}{\omega_{0i}}\right]^2}\right| \tag{19}$$

where ω_{0i} are the centre frequencies of the ith sections. The $\pm$ sign option is used in this equation to illustrate the fact that the worst-case error $\pm\varepsilon$ of the elements x_i can cause the *individual* terms of equation (19) to be of either polarity. The implications of this sign option will be discussed by means of an example. This function may be obtained in graphical form directly from the normalized sensitivity curve in Fig. 1. For example, consider the derivation of the worst-case sensitivity of a sixth-order narrowband Butterworth transfer function by means of equation (19). Straightforward analysis of the pole locations of a narrowband Butterworth function gives

$$Q_1 \simeq Q_3 \simeq \frac{2\omega_{02}}{BW} \qquad Q_2 \simeq \frac{\omega_{02}}{BW}$$

$$\omega_{01} = \omega_{02} - \frac{BW}{2}\cos 30^\circ; \qquad \omega_{03} = \omega_{02} + \frac{BW}{2}\cos 30^\circ$$

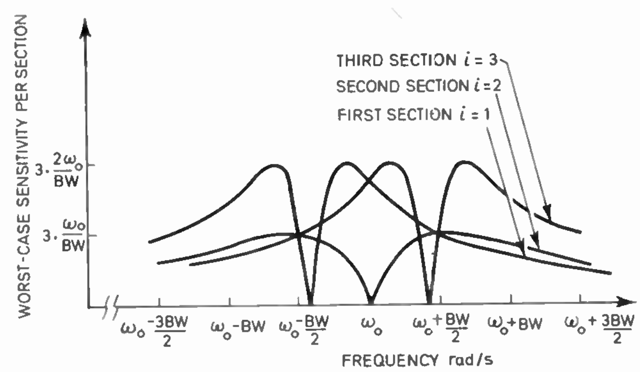
and ω_{02} is the centre frequency of the three-section realization. Using the above values for Q_i and ω_{0i}, the three terms of equation (22) are plotted in Fig. 2(a). The terms become zero at the corresponding ω_{0i} and exhibit maxima and minima of value $3Q_i$ at the corresponding frequencies $\omega_{0i} \pm \omega_{0i}/2Q_i$. The worst-case curve for $|T(j\omega)|$ depends on the sign options selected for each section. Thus, in Fig. 2(b) there are several possible interpretations of worst-case sensitivity.

(i) Worst-case sensitivity at any frequency
This is obtained by adding the moduli of the three curves, as in curve A in Fig. 2(b) and as given by equation (19).

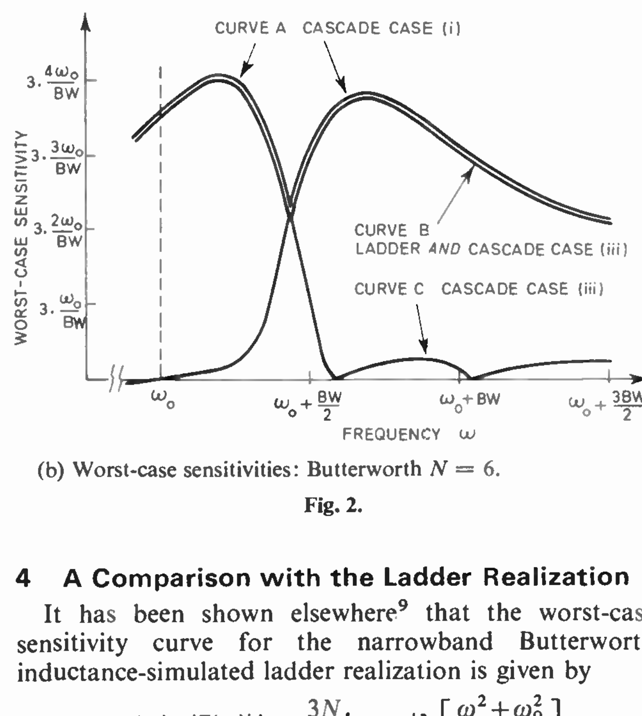
(ii) Worst-case sensitivity at band-edge $\omega_{02} \pm BW/2$
If the sign sequence in (19) for the terms $i = 1, 2, 3$ is $+, +, +$, or $-, -, -$, we obtain the worst-case band-edge sensitivity as curve B in Fig. 2(b). In this case, the curve peaks at $3{\cdot}806\omega_0/BW$.

(iii) Worst-case sensitivity at centre frequency ω_{02}
A sign sequence $+, -$, or $-, +$ for terms $i = 1, 2$ in (19) will give a worst-case result at the centre frequency ω_{02}. In this example, the worst-case centre frequency sensitivity is $3{\cdot}464\omega_0/BW$, as illustrated by curve C.

The absolute result of curve A is pessimistic because it will not be achieved at all frequencies simultaneously; it represents the absolute worst-case at any one frequency. The curves B and C are the particular sign options of $\pm\varepsilon$ that can contribute to the terms of equation (19) in such a way as to give the worst-case result at the centre frequency and the band-edge frequency, respectively.



(a) Single-section worst-case sensitivities: Butterworth $N = 6$.



(b) Worst-case sensitivities: Butterworth $N = 6$.

Fig. 2.

4 A Comparison with the Ladder Realization

It has been shown elsewhere[9] that the worst-case sensitivity curve for the narrowband Butterworth inductance-simulated ladder realization is given by

$$\sum_{\text{all } x_i} |S_{x_i}^{|T(j\omega)|}| \simeq \frac{3N}{3}|p(j\omega)|^2\left[\frac{\omega^2+\omega_0^2}{|\omega^2-\omega_0^2|}\right] \tag{20}$$

and that the function exhibits peaks at

$$\omega_{max} \simeq \pm\left\{(N-1)^{1/N}.\frac{BW}{2}\right\} + \omega_0 \tag{21}$$

of peak value

$$\sum |S_{x_i}^{T(j\omega max)}| \simeq 3(N-1)^{(N-1)/N}.\frac{\omega_0}{BW} \tag{22}$$

where N is the order, $|p(j\omega)|^2$ is the reflexion coefficient, BW is the bandwidth and ω_0 is the centre frequency. Consequently, it is possible to compare the cascade- and ladder-realizations in considerable detail by comparing

Table 1

Order N	2	4	6	8
LADDER Peak worst-case transition sensitivity (eqn. (22))	$3\ \omega_0/BW$	$6{\cdot}837\ \omega_0/BW$	$11{\cdot}47\ \omega_0/BW$	$16{\cdot}47\ \omega_0/BW$
CASCADE Peak worst-case transition sensitivity (eqn. (19))	$3\ \omega_0/BW$	$6{\cdot}786\ \omega_0/BW$	$11{\cdot}42\ \omega_0/BW$	$16{\cdot}43\ \omega_0/BW$
LADDER α_{max} (eqn. (21))	$\omega_0 + BW/2$	$\omega_0 + 0{\cdot}658\ BW$	$\omega_0 + 0{\cdot}654\ BW$	$\omega_0 + 0{\cdot}637\ BW$
CASCADE ω_{max}(eqn. (19))	$\omega_0 + BW/2$	$\omega_0 + 0{\cdot}707\ BW$	$\omega_0 + 0{\cdot}683\ BW$	$\omega_0 + 0{\cdot}653\ BW$

equations (19) and (20). For the sixth-order narrowband Butterworth ($N = 6$), equation (20) is given by curve B in Fig. 2(b). The passband response of the ladder exhibits a similar worst-case sensitivity to the corresponding case (ii) cascade realization and it will be observed in Fig. 2(b) that the peak *transition region* values are approximately $11{\cdot}4\omega_0/BW$ at approximately $\omega_0 \pm (0{\cdot}65)BW$ for both cascade- and ladder-methods.

Calculations of the peak Butterworth transition region worst-case sensitivities for $N = 2$, 4, 6 and 8 have been made directly from equations (21) and (22), for the ladder, and from equation (19) for the cascade-structure. The results are given in Table 1.

It will be observed that the peak worst-case sensitivities are similar for ladder- and cascade-methods and occur at similar frequencies in the transition region. It would therefore appear that, from the point of view of worst-case accuracy of the transfer function *in the transition region* there is little difference between the two methods. The cascade-method, however, *can* exhibit its worst-case peak sensitivity in the passband, in which case the transition band sensitivity is much improved (curve C) at the expense of a passband sensitivity that is much worse than the ladder.

5 Conclusion

The worst-case sensitivity of all-pole second-order narrowband bandpass functions may be evaluated in terms of the pole sensitivities and for biquad-, g.i.c.- and gyrator-type realizations *the sensitivity of the Q factor may be neglected.* An approximate expression may be obtained for the narrowband transfer function sensitivity that is only a function of Q, ω_0 and the sensitivity of ω_0 to the network parameters. This result allows an expression for the absolute worst-case sensitivity of cascaded sections to be obtained and it is found that the *absolute worst-case transition region sensitivity* is of similar magnitude and occurs at a similar frequency to that of the corresponding simulated-inductance ladder realization. The cascaded-section realization method is, however, inferior to the ladder method because it allows a far greater worst-case sensitivity in the passband.

Finally it should be noted that the sensitivity of the transfer function $KH(s)$ to the multiplier constant K has been neglected because S^k_{zl} is independent of frequency and therefore not generally of interest.

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A method of analysis and design of linear high-order systems

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SUMMARY

A stability criterion is defined for high-order linear systems using root-locus method. The general character of the system can be predicted by the trajectories plotted from this criterion. The sub-characteristic equations defined in the criterion afford a simple means for the design of systems with multiple adjustable parameters.

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1 Introduction

Stability is one of the primary concerns in the analysis and design of control systems. A number of methods for investigations of absolute and relative stability of linear systems has been developed. Nevertheless, a simple criterion for the high-order systems is yet desirable to be explored. This paper reports a procedure using root-locus technique for testing the stability of high-order systems and for analysing systems with multiple adjustable coefficients in their characteristic equations.

2 Stability Criterion

Consider the characteristic equation of a system with positive and real coefficients $A_0 \dots A_n$ given in the general form:

$$F(S) = A_n S^n + A_{n-1} S^{n-1} + \dots A_1 S + A_0 = 0 \qquad (1)$$

It can be written as

$$F(S) = F_o(S) + KF_e(S) = 0 \qquad (2)$$

where $F_e(S)$ and $F_o(S)$ are the even and odd parts respectively of the polynomial of eqn. (1) and K is a scaler. Let the roots of $F_e(S)$ be Z_j and the roots of $F_o(S)$ be P_j, where $j = 1, 2, \dots, n$. We may deduce a stability criterion for an nth-order system as follows:

For a linear nth-order system to be stable, necessary and sufficient conditions are that all the roots of the even part Z_j and the odd parts P_j of its characteristic equation lie on the imaginary axis of the S-plane.

We will show this criterion to be necessary and sufficient for stable physical systems by first verifying that:

(i) if all the real roots and all the real parts of the complex roots of the system characteristic equation are negative, i.e. all the zeros of the polynomial $F(S)$ lie in the left-half S-plane,

(ii) then if the polynomial $F(S)$ is written in the form

$$F(S) = F_o(S) + KF_e(S)$$

the zeros and poles of the function

$$H(S) = \frac{F_e(S)}{F_o(S)}$$

lie along the imaginary axis and must also interlace.

Starting with the factored form of the polynomial of equation (1) which reads

$$F(S) = A_n(S - r_1)(S - r_2) \dots (S - r_n) \qquad (3)$$

We have by multiplying out and collecting terms with like powers of S:

$$\frac{A_{n-1}}{A_n} = -(r_1 + r_2 + r_3 + \dots)$$

$$\frac{A_{n-2}}{A_n} = r_1 r_2 + r_1 r_2 + \dots r_{n-1} r_n$$

$$\frac{A_0}{A_n} = (-1)^n \prod_{j=1}^{j=n} r_j \qquad (4)$$

It is evident that if all the roots are real and negative, all the coefficients are positive and real. If some or all of the roots are in the form of conjugate complex pairs, it may likewise be shown by the use of equation (4) that all the coefficients are positive if the roots have negative real parts. Thus, it has been established that all the zeros of the polynomial $F(S)$ lie in the left half S-plane and the stability of the physical system is assured. A polynomial possessing this property is called a Hurwitz polynomial.

We will investigate the properties of the Hurwitz polynomial. Introduce the function

$$H(S) = \frac{F_e(S)}{F_o(S)} \tag{5}$$

The even and odd parts of $F(S)$ are defined by the equations

$$\begin{aligned} F_e(S) &= \tfrac{1}{2}[q(S)+q(-S)] \\ F_o(S) &= \tfrac{1}{2}[q(S)-q(-S)] \end{aligned} \tag{6}$$

Thus

$$H(S) = \left[\frac{q(S)+q(-S)}{q(S)-q(-S)}\right]$$

or

$$H(S) = \frac{\dfrac{q(S)}{q(-S)}+1}{\dfrac{q(S)}{q(-S)}-1} \tag{7}$$

We examine the nature of the quotient $q(S)/q(-S)$. In Fig. 1(a) is shown a point S_1 in the right half of the S-plane. This point, and the points $\bar{S}_1$, $-S_1$ and $-\bar{S}_1$ form a quad of points with the same magnitude for real and imaginary parts. In the same Figure are shown three zeros of $q(S)$ with left half-plane locations as required of Hurwitz polynomials. We observe that since d_1, d_2 and d_3 as identified in the Figure are the same phasor magnitudes:

$$\begin{aligned} |q(-\bar{S}_1)| &= |q(-S_1)| \\ &= d_1 d_2 d_3 \end{aligned} \tag{8}$$

In Fig. 1(b) we compare $|q(-\bar{S}_1)|$ with $|q(S_1)|$. Observe that d_1 and d_{10}, d_2 and d_{20}, d_3 and d_{30} have the same imaginary component, but the real component of d_{10} is greater than the real component of d_1, d_{20} greater than d_2 and d_{30} greater than d_3. Then

$$d_{10} > d_1, \qquad d_{20} > d_2, \qquad d_{30} > d_3$$

and since

$$|q(S_1)| = d_{10} d_{20} d_{30} > d_1 d_2 d_3$$

thus

$$|q(S_1)| > q(-S_1) \tag{9}$$

or

$$\left|\frac{q(S)}{q(-S)}\right| > 1 \quad \text{for Re } S > 0 \tag{10A}$$

By similar reasoning, we show that

$$\left|\frac{q(S)}{q(-S)}\right| = 1 \quad \text{for Re } S = 0 \tag{10B}$$

and

$$\left|\frac{q(S)}{q(-S)}\right| < 1 \quad \text{for Re } S < 0 \tag{10C}$$

Let

$$\frac{q(S)}{q(-S)} = u + \mathrm{j}v \tag{11}$$

so that for Re $S \geqslant 0$

$$\left|\frac{q(S)}{q(-S)}\right| = \sqrt{u^2+v^2} \geqslant 1 \tag{12}$$

Now

$$\begin{aligned} \text{Re } H(S) &= \text{Re}\left[\frac{u+\mathrm{j}v+1}{u+\mathrm{j}v-1}\right] \\ &= \left[\frac{u^2+v^2-1}{(u-1)^2+v^2}\right] \end{aligned} \tag{13}$$

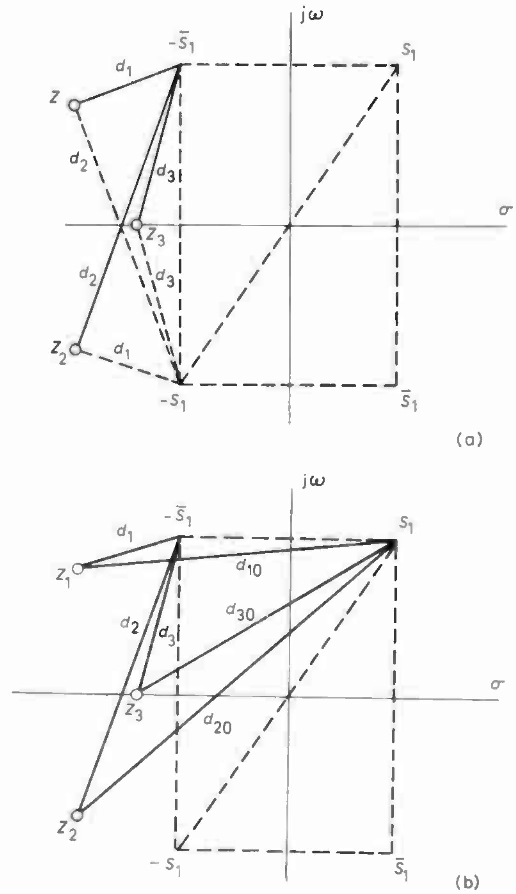

Fig. 1. Geometry shows (a) $|q(-\bar{S}_1)| = |q(-S_1)|$ and (b) $|q(S_1)| > |q(-S_1)|$ if S_1 has a right half plane location.

The denominator of this equation is always non-negative, and equation (12) compared with the numerator of this equation indicates that

$$\text{Re } H(S) \geqslant 0 \quad \text{for Re } S \geqslant 0 \tag{14}$$

This conclusion together with the observation from equation (7) that $H(S)$ is real when S is real tells us that $H(S)$ is a positive real function and its reciprocal $1/H(S)$ is also positive real.

Suppose the rational function $H(S)$ has a pole P_y in the right-half plane. Let us make a Laurent series expansion about P_y for $H(S)$

$$H(S) = \frac{b_{-m}}{(S-P_y)^m} + \frac{b_{-m+1}}{(S-P_y)^{m-1}} + \ldots + b_0 + b_1(S-P_y) + \ldots\ldots \quad (15)$$

In the neighbourhood of the pole P_y, $H(S)$ can be approximated by

$$H(S) = \frac{b_{-m}}{(S-P_y)^m} \quad (16)$$

Letting $(S-P_y)^m = \rho^m \mathrm{e}^{\mathrm{j}m\phi}$, and $b_{-m} = B\,\mathrm{e}^{\mathrm{j}\phi}$ one has

$$H(S) = \frac{B}{\rho^m}\,\mathrm{e}^{\mathrm{j}(\phi - m\theta)}$$

whence

$$\mathrm{Re}\,H(S) = \frac{B}{\rho^m}\cos(\phi - m\theta) \quad (17)$$

When θ varies from 0 to 2π, the sign of Re $H(S)$ will change $2m$ times. Since Re $H(S) \geqslant 0$ when Re $S \geqslant 0$, it is seen that any change of sign of Re $H(S)$ in the right half-plane will show that the function is not positive real. Therefore, the function $H(S)$ cannot have poles in the right half-plane. Since $H(S)$ is positive real, its reciprocal $1/H(S)$ is also positive real. Hence there cannot be any zeros in the right half-plane either.

With reference to the conditions given in equation (10) one is forced to conclude that the function $H(S)$ cannot have poles in either the right or the left half-plane. This fact restricts the poles of $H(S)$ to lie along the imaginary axis of the S-plane. Equation (17) imposes a further restriction upon these poles that the poles may exist on the jω axis if $m = 1$ and $\phi = 0$. The condition $m = 1$ implies that the pole is simple, and the condition $\phi = 0$ implies that the residue is positive and real. It is readily seen that zeros on the jω axis must also be simple. Thus, it follows that the highest powers of $F_e(S)$ and $F_o(S)$ as well as their lowest powers may not differ by more than unity and obviously they must differ at least by unity because $F_e(S)$ and $F_o(S)$ are respectively even and odd.

The independent variable S is a complex value which is written $S = \sigma + \mathrm{j}\omega$. If we write

$$H(S) = U(\sigma,\omega) + \mathrm{j}V(\sigma,\omega) \quad (18)$$

in which U and V are the real and imaginary parts. The Cauchy–Riemann condition equation shows that

$$\frac{\partial U}{\partial \sigma} = \frac{\partial V}{\partial \omega}, \qquad \frac{\partial U}{\partial \omega} = -\frac{\partial V}{\partial \sigma} \quad (19)$$

Since $U(\sigma,\omega)$ is identically zero for $\sigma = 0$ and positive for $\sigma > 0$, it follows that $\partial U/\partial \sigma$ is positive and hence $\partial V/\partial \omega$ is positive for $\sigma = 0$. In other words, when all the poles of $H(S)$ lie on the imaginary axis and its real part there is identically zero, one has

$$\left(\frac{\mathrm{d}H}{\mathrm{d}S}\right)_{S=\mathrm{j}\omega} = \frac{\mathrm{d}H(\mathrm{j}\omega)}{\mathrm{j}\mathrm{d}\omega} > 0 \quad (20)$$

A similar relation then also holds for the reciprocal function. This states that H is a continuously increasing function between poles. Thus, the consequence is that the zeros and poles of $H(S)$ which lie along the imaginary axis must interlace.

If the polynomials $F_e(S)$ and $F_o(S)$ are written in their factored form, then

$$H(S) = \frac{K(S^2 - Z_1^2)(S^2 - Z_2^2)\ldots(S^2 - Z_j^2)}{P_0(S^2 - P_1^2)(S^2\ldots P_2^2)\ldots(S^2 - P_j^2)} \quad (21)$$

The alternation of zeros and poles along the imaginary axis is expressed by the condition

$$P_0 < Z_1 < P_1 < Z_2 < \ldots$$

where P_0 is the root at the origin of the S-plane.

If $F(S)$ has poles on both sides of the imaginary axis, referring to Fig. 1, as S proceeds along the imaginary axis from $-\mathrm{j}\infty$ to $+\mathrm{j}\infty$ we still have at all times

$$\left|\frac{q(S)}{q(-S)}\right| = 1 \qquad \text{and} \qquad \arg\left(\frac{q(S)}{q(-S)}\right) = 2\arg q(S)$$

since $q(-S)$ and $q(S)$ are complex conjugates, but the modulus can now be unity elsewhere.

Each vector d in Fig. 1 drawn to a point on the imaginary axis $S_1 = \mathrm{j}\omega_1$ from a zero of $q(S)$ on the left increases its argument by π from $-\frac{1}{2}\pi$ to $+\frac{1}{2}\pi$. Due to each zero on the right this argument reduces by π. With n_l zeros on the left and n_r zeros on the right, arg $q(S)/q(-S)$ increases $2\pi(n_l - n_r)$ as S proceeds from $-\mathrm{j}\infty$ to $+\mathrm{j}\infty$. Hence $q(S)/q(-S) = 1$ only $(n_l - n_r)$ times and this is equal to the number of poles of $H(s)$ on the imaginary axis.

Hence if the condition (3) is satisfied, zeros of $q(S)$ cannot exist on both sides of the imaginary axis. Since in addition the coefficients $A_N \ldots, A_0$ of (1) are all positive, the sum of all roots of $F(S)$ must be negative. This demonstrates the sufficiency of the condition.

Now, manipulation of equation (2) yields

$$\frac{KF_e(S)}{F_o(S)} = -1 \quad (22)$$

which is a proper form for plotting root loci. Since the difference between the number of poles and zeros can only be 1, there is only one root locus coming from or going to infinity. The following examples serve to illustrate the proposed method for system stability analysis.

Example 1. Find the roots and hence character of a general 5th-order system

$$F(S) = S^5 + A_4 S^4 + A_3 S^3 + A_2 S^2 + A_1 S + A_0 = 0 \quad (23)$$

From equation (15)

$$\frac{KF_e(S)}{F_o(S)} = \frac{A_4[S^4 + (A_2/A_4)S^2 + A_0]}{S(S^4 + A_3 S^2 + A_1)} = -1 \quad (24)$$

The roots of equation (24) can be found using the standard method.

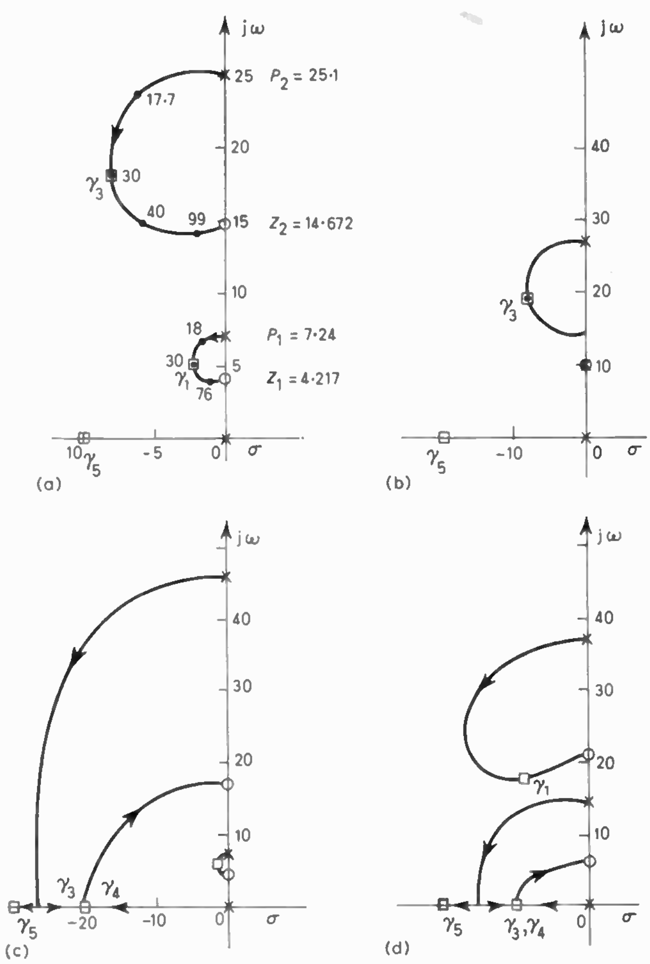

Fig. 2. Typical root loci for a general 5th order system.

Suppose that

$A_4 = 30, \quad A_3 = 683, \quad A_2 = 6992,$

$A_1 = 33072, \quad A_0 = 114830$

$$\frac{KF_e(S)}{F_o(S)} = \frac{30[S^4+233{\cdot}06S^2+3827{\cdot}67]}{S[S^4+683{\cdot}32S^2+33072]}$$
$$= \frac{30[S^2+(4.22)^2][S^2+(14{\cdot}67)^2]}{S[S^2+(7{\cdot}24)^2][S^2+(25{\cdot}1)^2]}$$
$$= -1 \qquad (25)$$

The root loci for this example is plotted in Fig. 2(a). From the plot it is predicted that this system is stable and has characteristic roots with reasonal damping character. Several root loci for different values of coefficients assigned to equation (24) are also plotted in Fig. 2, and from which the following observations can be made:

(i) A system is at its stability limit if a pole equals a zero.

(ii) A system has its characteristic roots with small damping ratio if a pole is very close to a zero.

(iii) A small (or large) value of the coefficient A_{n-1} can lead to poor damping character, as indicated by the gain marks in Fig. 2(a).

(iv) The frequencies of any pair of complex roots are limited by two pairs of poles and zeros which terminate the loci that pass through that pair of complex roots.

Example 2. Analyse the 9th-order system with the following equation

$$F(S) = S^9+80S^8+68S^7+1920S^6+1035S^5 +14400S^4+5543S^3+37920S^2 +8952S+28000 = 0 \qquad (26)$$

As before

$$\frac{KF_e(S)}{F_o(S)} = \frac{80[S^8+24S^6+180S^4+474S^2+350}{S[S^8+68S^6+1035S^4+5543S^2+8952]} \qquad (27)$$

After the roots of equation (27) have been found, the root loci are plotted in Fig. 3. The character of this system is predicted as follows:

(i) This system is stable.

(ii) There are two pairs of complex roots with frequencies respectively near $\omega = 1{\cdot}8$ and $\omega = 2{\cdot}7$ almost on the imaginary axis due to two zeros equal to two poles.

(iii) Another two pairs of roots approximately at $\omega = 1{\cdot}1$ and $\omega = 3{\cdot}5$ also have poor damping character because of the large value of coefficient A_{n-1}.

(iv) Since all the complex roots have small real parts, the real root is approximately equal to $-A_{n-1}$, i.e. -80.

From the results of these two examples it is seen that the stability and the general character of a system can be investigated by evaluating the real roots of the even and odd parts of the characteristic equation which we will define as sub-characteristic equations.

Since the order of the sub-characteristic equations is equal to half of the order of the characteristic equation, the effort for analysing a high order system is greatly

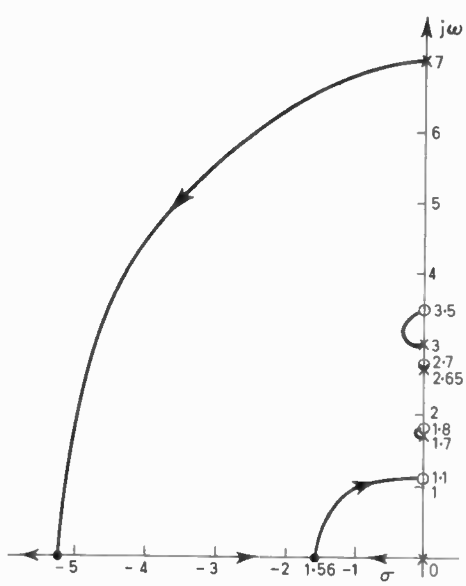

Fig. 3. Root-loci for Example 2.

reduced. This approach is also of interest in system design with multiple adjustable coefficients.

3 Adjustable Parameter Analysis

In finding the roots of a polynomial, graphical method sometimes is found convenient. Consider the polynomial having the general form

$$F(X) = C_m X^m + C_{m-1} X^{m-1} + \dots + C_1 X + C_0 = 0 \quad (28)$$

It can be written as

$$F(X) = F_1(X) + F_2(X) = 0 \quad (29)$$

where $F_1(X)$ and $F_2(X)$ are two separate parts of $F(X)$. A value of X which can make $F_1(X) = -F_2(X)$ is a root of equation (28). Graphically $F_1(X)$ and $F_2(X)$ can be represented by two curves plotted in the same plane. The intersections of these two curves are the roots of the equation. This concept leads us to develop a technique of analysis and design of systems with adjustable coefficients. Let $F(X)$ be the sub-characteristic equation. Let $F_1(X)$ be a curve and $F_2(X)$ be a straight line representing the function of adjustable parameters. $F_1(X)$ and $F_2(X)$ are aligned to give the desirable real roots for the sub-charactersitic equation. The root loci are then plotted and from which we can find the characteristic roots of the system.

Assume that the sub-characteristic equation of the system has one adjustable coefficient, C_k. The expression of equation (28) can be written as

$$F(X) = \sum_{i=k+1}^{m} C_i X^i + C_k X^k + \sum_{j=0}^{k=1} C_j X^j = 0 \quad (30)$$

Dividing throughout by X^k we can let

$$F_1(X) = \sum_{i=k+1}^{m} C_i X^{i-k} + \sum_{j=0}^{k-1} C_j X^{j-k} \quad (31)$$

and

$$F_2(X) = C_k \quad (32)$$

Similarly, if the sub-characteristic equation of the system has two adjustable coefficients, C_k and C_{k+1}, we have

$$F_1(X) = \sum_{i=k+2}^{m} C\, X^{i-k} + \sum_{j=0}^{k-1} C_j X^{j-k} \quad (33)$$

$$F_2(X) = C_{k+1} X + C_k \quad (34)$$

Equations (32) and (34) are straight lines, and their intersections with equations (31) and (33) respectively can be found easily. The best selections of adjustable coefficients represent the case where no poles or zeros are close to each other, thus a reasonable damping can be expected for any pair of complex characteristic roots. The design method is illustrated by an example.

Example 3. Design a 9th-order system with the following characteristic equation

$$F(S) = S^9 + 60S^8 + 1900S^7 + A_6 S^6 + A_5 S^5 + \\ + 5400000S^4 + 48000000S^3 + \\ + A_2 S^2 + A_1 S + A_0 = 0 \quad (35)$$

where A_0, A_1, A_2, A_5 and A_6 are adjustable coefficients. A pair of dominant roots at $\xi = 0{\cdot}4$ and $\omega = 4$ is desired. From equation (23):

$$\frac{KF_c(S)}{F_o(S)} = \frac{60\left[S^8 + \dfrac{A_6}{60} S^6 + 90000S^4 + \dfrac{A_2}{60} S^2 + A_0\right]}{S[S^8 + 1900S^6 + A_5 S^4 + 48000000S^2 + A_1]} \quad (36)$$

The sub-characteristic equations are

$$F(X_1) = X_1^4 + C_6 X_1^3 + 90000X_1^2 + C_2 X_1 + C_0 \quad (37)$$

$$F(X_2) = X_2^4 + 1900X_2^3 + C_5 X_2^2 + 48000000X_2 + C_1 \quad (38)$$

where

$$C_6 = A_6/60, \quad C_5 = A_5$$
$$C_2 = A_2/60, \quad C_1 = A_1$$
$$C_0 = A_0/60, \quad X_1 = X_2 = S^2$$

Thus,

$$F_1(X_1) = X_1^2(X_1^2 + C_6 X + 90000) \quad (39)$$

$$F_2(X_1) = C_2 X_1 + C_0 \quad (40)$$

and

$$F_1(X_2) = X_2^2(X_2^2 + 1900X_2 + C_5) \quad (41)$$

$$F_2(X_2) = 48000000X_2 + C_1 \quad (42)$$

There must be $m-2$ inflexion points on the curve defined by equation (33) since equation (34) represents a straight sub-characteristic equation. Equating the second derivative of equation (33) to zero, the maxima and minima of the curve can be found. Hence

$$\frac{dF_1(X_1)}{dX_1} = X_1(X_1^2 + -C_6 X + 45000) = 0 \quad (43)$$

$$\frac{d^2F_1(X_1)}{dX_1^2} = X_1^2 + \tfrac{1}{2}C_6 X_1 + 1500 = 0 \quad (44)$$

$$\frac{dF_1(X_2)}{dX_2} = X_2(X_2^2 + 1425X_2 + \tfrac{1}{2}C_5) = 0 \quad (45)$$

$$\frac{d^2F_1(X_2)}{dX_2^2} = X_2^2 + 950X_2 + \tfrac{1}{6}C_5 = 0 \quad (46)$$

Curves for equations (44) to (47) shown in Figs. 4 and 5 indicate the limiting values for C_6 and C_5 which yield only one inflexion point. These imply that for the system to be stable, the value of C_6 must be chosen larger than 500 and C_5 be less than 135×10^4. Equation (39) is plotted in Fig. 6 for several values of C_6. Choosing $C_6 = 580$ and adjusting the straight line defined by equation (40), four intersections with the curve are found. They are not too close to each other, and the system can be expected to have reasonable damping. The magnitude of the first root is smaller than ω^2. Similarly, equation (41) is plotted in Fig. 7 for various values of C_5. The selections of real roots are limited in this case because straight lines with constant slope as defined by equation (42) can only be drawn. The curve $C_5 = 600000$ is chosen, and the intersections of this curve and the straight line give four real roots apart.

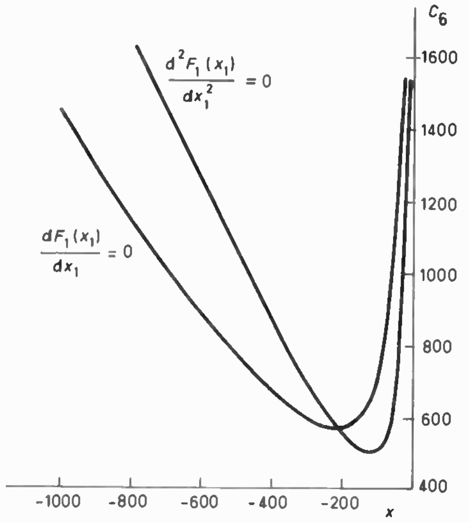

Fig. 4. Limiting value for C_6.

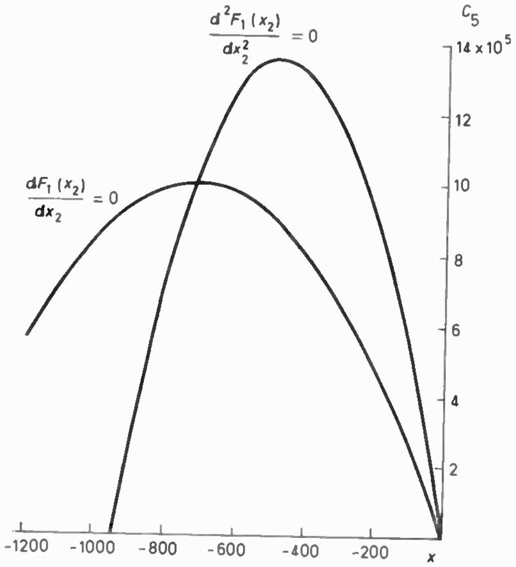

Fig. 5. Limiting value for C_5.

The magnitude of the first root is larger than ω^2. The root loci are shown in Fig. 8. Since the value of the coefficient A_{n-1} is not too small, the system is predicted to be able to meet the specifications approximately. Figure 3 gives

$$\frac{KF_e(S)}{F_o(S)} = \frac{\dfrac{60[S^2+(18\cdot6)^2][S^2+(13\cdot34)^2]}{[S^2+(6\cdot9)^2][S^2+(2\cdot8)^2]}}{\dfrac{S[S^2+(39\cdot08)^2][S^2+(16\cdot14)^2]}{[S^2+(9\cdot5)^2][S^2+(4\cdot65)^2]}} \tag{47}$$

which satisfies the stability criterion.

4 Conclusion

This paper presents a simple means of assessing the character of high-order linear systems with multiple adjustable coefficients. Compared with the commonly used techniques such as Mitrovic's chart or D-partition

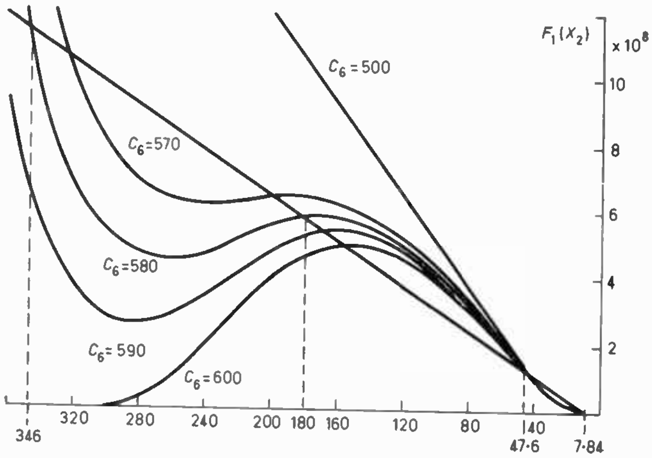

Fig. 6. Graphical solution of equation (38).

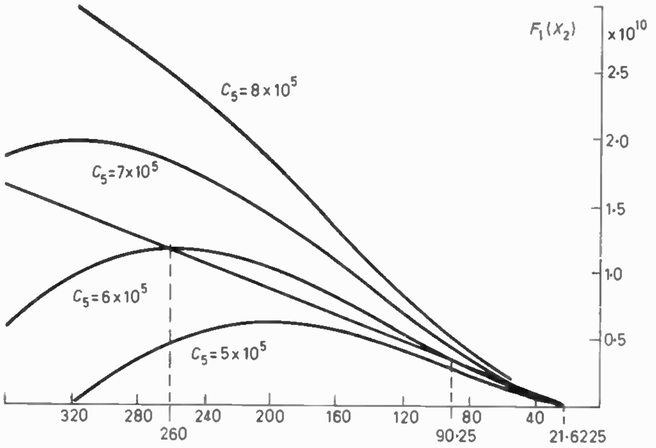

Fig. 7. Graphical solution of equation (39).

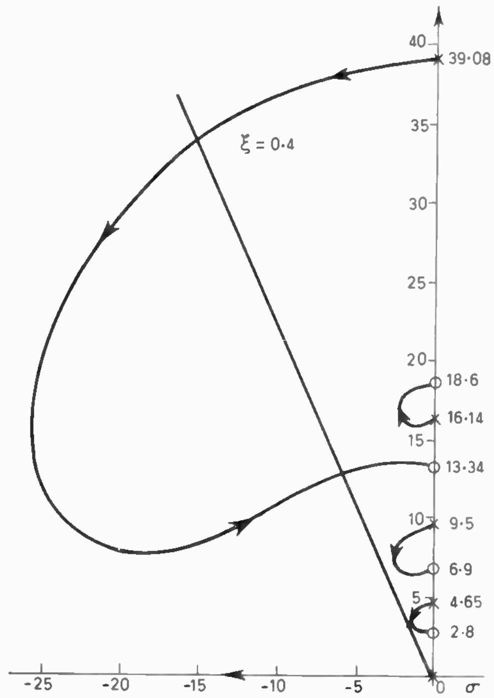

Fig. 8. Root loci for Example 3.

which require multi-dimensional space, the proposed method describes the sub-characteristic equations in two separated planes and considerably reduces the effort of calculations.

5 Acknowledgments

The author is grateful to the Referees for their helpful constructive criticism of this paper.

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Joint United Kingdom-Canadian Study of Quasars

Canadian and British radio astronomers have begun a joint experiment to shed new light on the nature of the baffling cosmic objects known as quasars. Scientists from the National Research Council of Canada, the University of Toronto and Queen's University on the one hand and the Radio and Space Research Station of the UK Science Research Council on the other, have combined the signals from quasars that were received simultaneously on radio telescopes in Canada and England, 3270 miles apart. The arrangement forms a radio interferometer having the ability to distinguish detail which improves the separation between the telescopes and with decreasing radio wavelength.

Quasars (the term is derived from 'quasi-stellar radio sources' because of their very small star-like appearance) are believed to be the most distant objects in the universe and are extraordinarily powerful emitters of radio waves. The radio emission is thought to be produced by high-energy electrons travelling in weak magnetic fields and to extend over a region which appears, as do distant stars, to be very small. So compact are the quasars that interferometers capable of very fine discrimination of detail are needed for any meaningful measurements at all.

The new interferometer uses an unusually short wavelength of 2·8 centimetres at both telescopes, a 150-foot (46 m) reflector at the Algonquin Radio Observatory, Ontario, Canada and an 85-foot (26 m) reflector at Chilbolton, England. The combined instrument can measure detail as small as 0·0004 seconds of arc—the equivalent of being able to stand in England and distinguish a marble held in Canada. Detail of this order has been found in several quasars.

The improvement in the interferometer comes about largely from the use of hydrogen maser atomic clocks, which allow the observations to be made at short wavelengths. Not only does this improve the discrimination of detail, but the detail itself is different at short wavelengths. Intriguing evidence from previous short wavelength observations of some quasars suggests that their size is rapidly changing. These observations can be interpreted in at least three ways: the quasars are expanding faster than the speed of light if they are at the distances inferred from the shift in wavelength of their spectral line; they are expanding at reasonable speeds and are much closer; or we are the victims of a form of optical illusion which depends upon special conditions in the quasar. Because of this puzzling variability in quasars the observations will be made at intervals of a few weeks for approximately a year.

For the experiment to work, observations from the two telescopes must be synchronized to within one ten millionth of a second and the tuning of the two receivers must be maintained identical to within one part in 10^{12}, a degree of accuracy made possible by the hydrogen maser clocks. Quasar signals are recorded at each telescope using video tape recorders and are finally combined after transporting both sets of tapes to the Astrophysics Branch of NRC's Radio and Electrical Engineering Division in Ottawa.

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Transient analysis on a small computer

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SUMMARY

A transient analysis program is described which is now operational on a small computer with 16k words of store. This program, entitled SOTRAP, is able to perform low-cost time response simulation of networks which contain bipolar transistors and diodes. Details are given of the program's convenient input-data format, analysis technique and built-in device models.

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1 Introduction

During the last decade, computer programs for circuit analysis have become an indispensable tool for design engineers. Many of the programs which are currently in use are suitable only for large computer installations and some of these[1,2] use slow integration algorithms which often make the time needed for transient analysis prohibitively long. A third generation of programs is now becoming available[3,4] where 'implicit' integration techniques are used in place of former 'explicit' methods. Such techniques have greatly reduced computing time and made time-response simulation of large non-linear networks a sound proposition.

However, there is still a need for transient analysis programs which operate in small computer systems, and such a program, entitled SOTRAP, has been implemented on a Honeywell DDP–516 machine of 16k word store. This program uses an implicit integration technique and has a convenient conversational input-data system which allows the engineer to work 'on-line' with the computer. Use has been made of a Burrough's fixed-head disk file, where numerical results are stored for subsequent use by a visual display program. However, the program does not use overlay or backing store, and is able to give the results of an analysis directly, without the aid of a disk file.

The program is written mostly in basic *Fortran IV* and is used under the BOS operating system. A small number of additional sub-routines were written in assembly language in order to save core space and to allow a flexible input-data structure. All numbers are read in free-format, this being made possible by the use of sub-routines which have been added to the standard *Fortran* library by members of the Department of Electronics at Southampton University.

2 Input Data

The program, in its present form, is able to analyse networks with up to about 25 nodes, up to ten transistors and a fair number of resistors, capacitors, inductors, diodes and independent sources. The components are represented in the input-data by a single or double letter command, e.g. R for resistor, T for transistor and IT for time-dependent current source. The command is usually followed by the numerical value of the component and the node numbers which designate its connexion points in the network. For diodes and transistors, the numerical value is replaced by a type number which refers to a list of device parameters stored as a separate item of data. Time-dependent current sources provide single pulses of the form shown in Fig. 1.

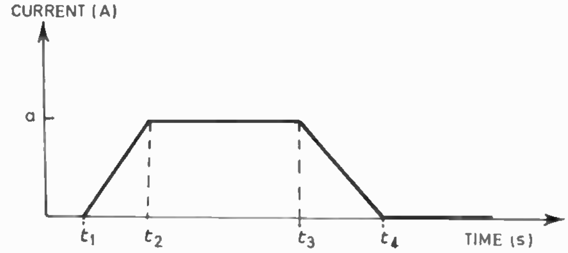

Fig. 1. Time-dependent current source.

A current source of the type shown in Fig. 1, connected between nodes n_t and n_2, with current flowing into node n_2, is entered by the following line of data:

$$\text{IT} \quad a \quad n_1 \quad n_2 \quad t_1 \quad t_2 \quad t_3 \quad t_4$$

Combinations of time-dependent current sources, together with fixed voltage and current sources may be used to model more complex excitations.

Components may be entered in any order, either from an on-line teletype or a prepared punched paper tape. A combination of both of these methods may be used. Standard units are assumed for all items of data, these being volts, amperes, seconds, hertz, ohms and farads. Additional single letter commands allow the network to be altered, or printed out for the purpose of checking. Errors in the input-data can quickly be rectified, on-line, by deleting and then replacing incorrect circuit components or other input-data statements.

The input data list reproduced in Table 1 defines the standard 't.t.l.-gate' circuit shown in Fig. 2. The statement beginning 'NN' defines the number of nodes, initial and final time of analysis, accuracy (volts) required and printing step to be used for the output of results. Diodes (D) are defined by their anode and cathode node-numbers followed by a type number; transistors (T) by the collector, base, emitter node number, type number and a final integer which is +1 for n-p-n and −1 for p-n-p. Fixed voltage sources (V) require a small series resistance which is specified after the voltage value and node-numbers. The data list must be supplemented by the transistor and diode parameters for each type number used. This information is used to define the device models and is usually stored on a standard reel of paper tape.

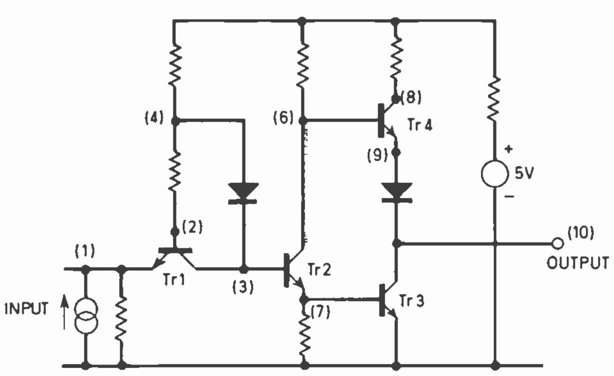

Fig. 2. T.T.L.–NAND gate driven by a single input pulse.

Table 1

Data for t.t.l.-gate shown in Fig. 2

'T.T.L.-GATE CIRCUIT'											
NN	10	0		200E–9				1E–9	E–3		
R	50			1		0					
R	300			4		2					
R	4000			5		4					
R	1600			5		6					
R	130			5		8					
R	1000			7		0					
R	1			5		0					
D	4			3		1					
D	9			10		2					
T	3	2	1		3		1				
T	6	3	7		2		1				
T	8	6	9		1		1				
T	10	7	0		1		1				
V	−5			5		0		1			
IT	·08			0		1		10E–9	70E–9	100E–9	120E–9

The analysis starts with d.c. node-voltages which are supplied as part of the input-data. When these voltages are not known, approximate values may be given if the action of time-dependent current is delayed to allow the network to 'settle down' to its initial state. This time delay need not be long as fairly accurate d.c. conditions can be realized in a few iterations.

Since the computer used has no line-printer, all printed numerical results must be obtained from a teletype. The quantity of numerical information which is available from the transient analysis program is considerable and thus a full print-out of results would take a prohibitive amount of time. It is therefore important to keep the amount of printing as small as possible, and the program allows the user to do this by specifying the node-voltages which are of interest to him. The voltages at only these chosen nodes are printed at each step, and even this output may subsequently be inhibited during the course of an analysis by the use of a sense-key on the computer console. Any results may be obtained after the completion of an analysis from the 'results' disk file.

3 Models of Diodes and Transistors

As a mathematical representation of the behaviour of a complex electrical device, a model should accurately predict its performance and be characterized by parameters which can be derived from practical measurements. Any model must be restricted to some extent in the range of usefulness in order to prevent it being too cumbersome. For example, it is difficult to include the conductivity modulation in the base and in the collector in a model of moderate complexity.

In the program SOTRAP the Ebers-Moll[6] models are used for non-linear transistor and diode representation. The diode is represented by the equivalent circuit shown in Fig. 3, where C is the sum of the depletion and diffusion capacitors and the voltage dependent current

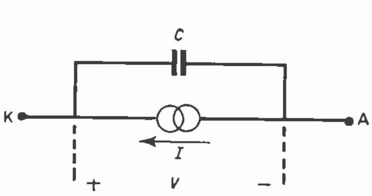

Fig. 3. Diode model.

source I is given by:

$$I = I_s(\exp(\lambda V)-1)$$

where I_s = diode saturation current (A)

λ = constant of the diode = $q/MkT\ (\mathrm{V}^{-1})$.

Both C and I are non-linear in nature and are calculated at each time-step of the analysis from six diode parameters which are entered as input-data.

Transistors are represented by equivalent circuits as shown in Fig. 4. Fifteen parameters are needed to characterize such a model, these being necessary to define the emitter and collector junctions with their depletion and diffusion capacitance as for the diode, the forward current gain (α_N), the inverse current gain (α_I) and the base resistance R_b.

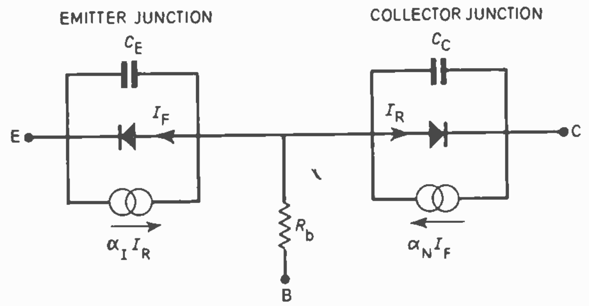

Fig. 4. Transistor model.

4 The Analysis Method

The program uses the nodal analysis method which was described in a recent letter.[7] This makes the setting up of network equations a very simple process, and in general uses less core storage than other standard techniques. Kirchhoff's current law may be expressed as follows:

$$GV(t)+C\frac{\mathrm{d}}{\mathrm{d}t}V(t)+\Gamma\int_0^t V(u)\,\mathrm{d}u = -I(t) \qquad (1)$$

where G = nodal admittance matrix of conductors

C = nodal capacitance matrix

Γ = nodal matrix of reciprocal inductance

$V(t)$ = vector of instantaneous node-voltage at time t

$I(t)$ = vector of instantaneous current-sources at time t.

Assuming that $V(t)$ has been calculated the node voltage vector at a small incremental time step Δt later, $V(t+\Delta t)$, may be calculated from an 'implicit' finite difference approximation to equation (1):

$$GV(t+\Delta t)+C\frac{1}{\Delta t}[V(t+\Delta t)-V(t)]+$$

$$+\Gamma\left[\int_0^t V(u)\,\mathrm{d}u+\Delta t V(t+\Delta t)\right] = -I(t+\Delta t) \qquad (2)$$

which gives:

$$\left[G+\frac{1}{\Delta t}C+\Gamma\Delta t\right]V(t+\Delta t)$$

$$=\frac{1}{\Delta t}CV(t)-I_L(t)-I(t+\Delta t) \qquad (3)$$

where $I_L(t) = \Gamma\int_0^t V(u)\,\mathrm{d}u$ is the vector of inductor currents at time t.

Writing equation (3) in the form

$$\mathrm{Y}\,.\,V(t+\Delta t) = J \qquad (4)$$

it can be seen that the node-voltage vector $V(t+\Delta t)$ is the solution of a set of simultaneous equations. When the network does not contain diodes or transistors, the matrix Y is independent of V and J, and equation (4) represents a set of linear equations which may be solved very easily, using a Gaussian elimination technique. It can be shown that Y must be non-singular for a well-defined circuit since each node will have at least one component connected to it. It can also be shown that the numerical solution of equation (4) is stable for any time-increment Δt.

When non-linear elements are present, J and Y become voltage dependent and a more complex, iterative technique must be employed to solve equation (4).

The program uses a modified Newton–Raphson technique[5] where at each iteration, non-linear devices are approximated by linear elements using the values of node-voltage which were calculated by the previous iteration. This gives a linear equation in the form of expression (4) which may be solved by Gaussian elimination. A new node-voltage vector $V(t+\Delta t)$ is hence obtained, which is a better approximation to the true situation than the previous voltage vector. A better approximation to the non-linear devices may now be calculated, and this gives an iterative process which quickly converges to the true node-voltage vector $V(t+\Delta t)$.

5 Examples

The circuit shown in Fig. 2 is a standard t.t.l.-NAND gate with ten independent-nodes, four transistors and two diodes. In order to make an analysis of this circuit, the a.c. parameters necessary to define an Ebers-Moll model for the diodes and transistors, were estimated from the geometry of the devices. D.c. parameters are easily calculated from d.c. characteristics given, for example, in the 'Texas Data Book'. More accurate results can, of course, be obtained if the a.c. parameters of the devices are evaluated from practical measurements. Figure 5 shows a plot of the results which were obtained from the analysis. The graph shows the input and output voltages plotted against time. A similar graph of any

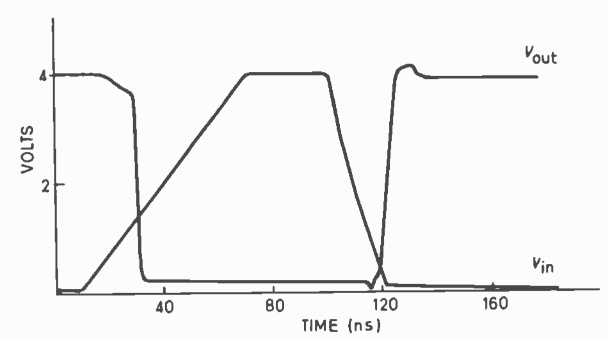

Fig. 5. Results of analysis of t.t.l.–NAND gate.

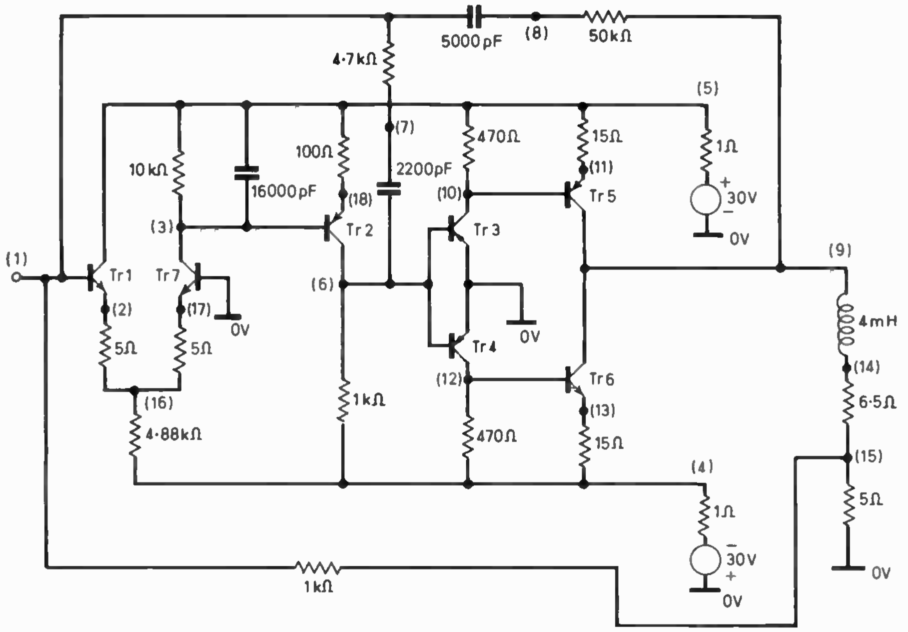

Fig. 6. Power amplifier with feedback components.

node in the network may be displayed directly on an 'ETOM' visual display unit. This graph is produced from the 'results' disk file created by SOTRAP, by means of a supplementary program which operates independently of the analysis program.

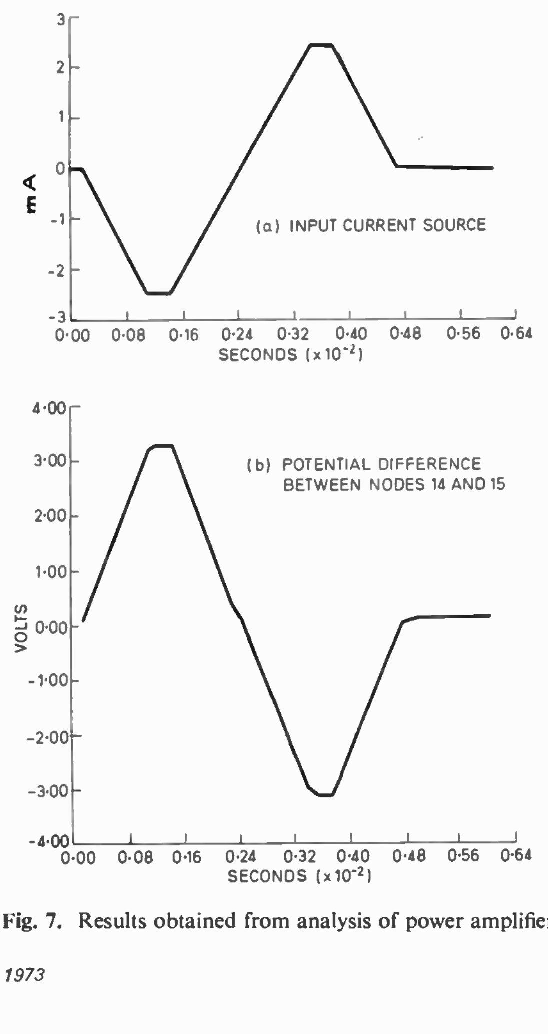

Fig. 7. Results obtained from analysis of power amplifier.

The time taken to perform the analysis was 8 minutes. This is quite long because of the fact that the computer is not equipped with fast, hardware floating-point arithmetic facilities. All floating-point arithmetic is performed by software sub-routines, which require about $\frac{1}{4}$ ms for each addition and multiplication and 2 ms for exponentiation. It is hoped that, in the near future, the program will be made very much faster by various programming techniques and possibly some hardware.

Another network which has been analysed successfully is shown in Fig. 6. The network is a power deflexion amplifier designed for use in a scanning electron-beam equipment. The circuit shown consists of seven transistors and twenty-six nodes (including the reference node and an internal node for each of the seven transistor models). Darlington pairs of transistors are represented by the single transistors Tr5 and Tr6 which have been given circuit parameters appropriate to the pairs they replace. The analysis was carried out in order to minimize the cross-over distortion for a given gain specification. To achieve this, the feedback parameters were to be adjusted. The input pulses were provided by two time-dependent current sources connected in parallel, the second with a negative value delayed until the first had returned to zero. The current in the output coil was deduced from the potential difference between nodes 14 and 15. The graph shown in Fig. 7 is typical of those obtained from the analysis.

The computation for these results was completed in approximately ten minutes.

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STANDARD FREQUENCY TRANSMISSIONS—March 1973

(Communication from the National Physical Laboratory)

March 1973	Deviation from nominal frequency in parts in 10^{10} (24-hour mean centred on 0300 UT)			Relative phase readings in microseconds N.P.L.—Station (Readings at 1500 UT)		March 1973	Deviation from nominal frequency in parts in 10^{10} (24-hour mean centred on 0300 UT)			Relative phase readings in microseconds N.P.L.—Station (Readings at 1500 UT)	
	GBR 16 kHz	MSF 60 kHz	Droitwich 200 kHz	*GBR 16 kHz	†MSF 60 kHz		GBR 16 kHz	MSF 60 kHz	Droitwich 200 kHz	*GBR 16 kHz	†MSF 60 kHz
1	+0·3	+0·1	0	719	643·3	**17**	−0·1	0	+0·1	711	615·3
2	0	+0·1	0	719	642·6	**18**	0	0	0	711	615·1
3	+0·2	+0·1	0	717	641·4	**19**	0	0	0	711	615·3
4	−0·1	+0·2	0	718	639·9	**20**	0	0	0	711	614·1
5	+0·1	+0·2	0	717	638·4	**21**	−0·1	0	+0·1	712	614·3
6	0	+0·1	0	717	620·1	**22**	0	−0·1	+0·1	712	615·0
7	+0·1	0	0	716	620·1	**23**	−0·1	0	0	713	615·3
8	+0·3	+0·1	+0·1	713	618·9	**24**	0	0	−0·1	713	615·6
9	0	+0·2	0	713	617·4	**25**	−0·1	0	−0·1	714	615·9
10	+0·1	+0·1	0	712	616·7	**26**	0	+0·1	0	714	615·4
11	+0·1	+0·1	0	711	615·8	**27**	0	0	0	714	615·2
12	0	+0·1	0	711	616·3	**28**	0	0	0	714	615·2
13	0	+0·1	0	711	615·8	**29**	0	0	0	714	615·6
14	0	+0.1	0	711	615·3	**30**	0	0	0	714	615·8
15	+0·1	+0.1	0	710	614·8	**31**	0	−0·1	0	714	616·8
16	0	0	0	710	615·0						

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

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

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

UDC 621.391.823 : 621.396.62 : 621.396.669.8

Impulsive noise reduction in radio receivers

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Based on a paper presented at the Conference on Radio Receivers and Associated Systems held at Swansea from 4th to 6th July 1972

SUMMARY

The spectrum of impulsive noise is quite different from that of an a.m. or s.s.b. radio signal, but receivers designed to discriminate against the former have only been moderately successful.

A mathematical analysis of a noise blanking receiver demonstrates the critical nature of system parameters. Signals in adjacent channels can give rise to excess noise in the wanted channel. Trapezoidal blanking functions are shown to have significant advantages, and the necessity for some means of delaying the signal before blanking is demonstrated.

A noise blanking receiver incorporating an ultrasonic glass delay line is described.

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1 The Problem of Impulsive Noise

It was early realized[1,2] that much of the noise which reduces the effectiveness of radio communication is impulsive in character. Since the noise impulse consists of a short burst of energy very broadly distributed across the radio frequency spectrum, it is, in principle, easily distinguished from a typical narrowband radio transmission. Carson[3] showed the limitations of linear networks for this purpose. Non-linear circuits were more promising and the limiter proposed by Dickert[4] saw widespread application, in a variety of modified forms. Weighton[5] provided a definitive mathematical analysis of limiters, as applied to a.m. systems, drawing attention to the need for relatively wideband circuits before the limiter to maximize the amplitude difference between the wanted and unwanted signals whilst keeping the noise impulse relatively short. Narrowband filters necessary to give the receiver its required adjacent channel selectivity needs should follow, not preceed, the limiter.

Parallel to the work on limiters (or clippers) for impulse noise reduction, there has also been considerable interest in blankers. Whereas a clipper limits the maximum signal excursion to a fixed value, a blanker consists of a circuit which reduces the receiver gain to a very low value for the duration of the noise pulse. Neither type of circuit shows a very marked advantage over the other, and the contemporary state of the art in respect of both is well reviewed by Pappenfus *et al.*[6]

Many parts of the electromagnetic spectrum are already very congested, and this situation is likely to get worse. Thus the present almost universal use of s.s.b. for professional communications in the h.f. band is likely to pressage a similar trend at other frequencies as the technical problems of using s.s.b. are overcome. For minimum bandwidth data transmission the situation is less clear, with both d.s.b. and p.s.k. as possible future lines of development. In any event, narrowband systems such as these are likely to be troubled much more by impulsive interference than the wider band systems used hitherto, unless special precautions are taken. In a recent review of trends in v.h.f. mobile radio techniques, Beusing[7] goes so far as to rule out completely the use of s.s.b. because of the problem of ignition interference, which he apparently considers to be insoluble. Theoretically, there is no reason why this should be so, since the noise pulse as received is quite different in character from the wanted signal. It ought therefore to be perfectly possible to design an effective blanker.

The present paper will review the problem of impulse noise reduction in a.m. and s.s.b. receivers, demonstrating the reasons why attempts to achieve in practice what is theoretically possible have met with limited success. The critical parameters of a blanking system will be identified, and an approach to the design of blankers described which partly circumvents these difficulties.

2 The Noise Blanking Receiver

A typical general block diagram for a noise blanking receiver appears in Fig. 1. The signal from the mixer passes to two channels: a noise channel, in which the occurrence of a noise pulse is detected and a suitable blanking pulse is generated, and a signal channel, which is fairly conventional except that it contains a blanker, an element which normally passes the signal with little loss, but which can introduce a very large attenuation when a control voltage is applied to it. The signal may undergo a time delay before entering the blanker, the purpose of which is to ensure that the blanking voltage from the noise channel has had time to bring the blanker into its high attenuation state before the noise pulse in the signal channel arrives at it.

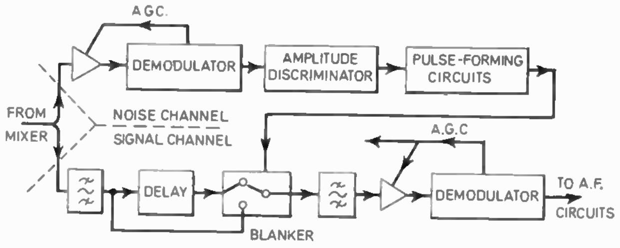

Fig. 1. A noise blanking receiver.

The noise channel must distinguish between impulsive noise and other signals in the pass band over which it operates. This is achieved by amplitude discrimination, thus the bandwidth of the noise channel is kept wide to give a large peak/mean ratio for noise impulses. Demodulation is by a simple diode, keeping circuit time-constants short, and a.g.c. is applied to the noise amplifier. The amplitude discriminator compares the instantaneous output from the noise channel with its mean, giving an output if the former exceeds the latter by more than a preset ratio. This output passes to pulse-forming circuits, in which a voltage of the right magnitude, waveform and duration is generated to give the required blanking action.

Although receivers of this general class have been described in the literature,[6] they have been only moderately successful, and there appears to have been no published analysis of the properties of noise blanking which fully considers all the problems of design optimization.

3 An Analytical Approach to Noise Blanking

In order to develop a mathematical model of the noise blanking receiver, it is first necessary to consider the nature of the noise against which the receiver is to be made resistant. For present purposes, the noise pulses will be treated as equivalent to true delta functions, having negligible duration but non-zero amplitude time integral. Although such pulses can occur randomly and sporadically, the most severe case from the point of view of signal degradation is that in which the pulse is repeated at regular intervals. This is the situation with interference from spark ignition vehicles and much electrical machinery. The range of pulse repetition rates typically extends from the bottom of the audio frequency range down to a few hertz.

The receiver multiplies the signal by a blanking function $y(t)$, which has the value zero whenever a noise pulse is presented to the blanker and a non-zero value at other times. Usually the blanking function will be trapezoidal, since the transitions between zero and maximum transmission take finite time, but initially the argument will be developed in terms of a rectangular blanking function, for which $y(t) = 1$ for a time T_0, and $y(t) = 0$ for $(T-T_0)$, where T is the repetition period.

Three types of modulation are of significance in connexion with noise blanking: a.m., d.s.b. with diminished or suppressed carrier, and s.s.b. Both a.m. and d.s.b. give rise to waveforms adequately represented for a simple sinusoidal modulation by $x(t)$, where, if the carrier frequency is ω_c and that of the modulation ω_m,

$$x(t) = [a+b\cos(\omega_m t)]\cos(\omega_c t) \tag{1}$$

where a, b are constants; s.s.b. can be represented as the sum of two d.s.b.s.c. signals, and need not, therefore, be considered independently.

After blanking, the waveform has an amplitude function $z(t)$, such that

$$z(t).\cos(\omega_c t) = y(t).x(t) \tag{2}$$

This waveform is then demodulated (synchronously in the case of diminished or suppressed carrier modulation) and the result is a voltage proportional to the amplitude function $z(t)$. $z(t)$ is not simply identical to the amplitude of $x(t)$, but contains a spurious (noise) component, due to the action of blanking on the signal. Thus

$$z(t) = y(t).[a+b\cos(\omega_m t)] \tag{3}$$

The function $y(t)$ may itself be expanded into its Fourier components:

$$y(t) = \sum_{n=0}^{\infty} \frac{1}{n\pi}\sin\left(n\pi\frac{T_0}{T}\right).\cos\left(2n\pi\frac{t}{T}\right)$$

$$= \frac{T_0}{T} + \sum_{n=1}^{\infty} \frac{2}{n\pi}\sin\left(n\pi\frac{T_0}{T}\right).\cos\left(2n\pi\frac{t}{T}\right)$$

assuming

$$0 \leqslant y(t) \leqslant \tfrac{1}{2}$$

Hence

$$z(t) = a.\frac{T_0}{T} + a.\sum\frac{2}{n\pi}\sin\left(n\frac{T_0}{T}\right).\cos\left(2n\pi\frac{t}{T}\right)+$$

$$+b.\frac{T_0}{T}\cos(\omega_m t)+b.\cos(\omega_m t)\times$$

$$\times\sum\frac{2}{n\pi}\sin\left(n\pi\frac{T_0}{T}\right).\cos\left(2n\pi\frac{t}{T}\right) \tag{4}$$

On the right-hand side of this equation, the first term is a d.c. component, which does not contribute signal or noise to the output, but may be used for a.g.c. It will be omitted in what follows. The second and fourth terms represent noise due to blanking, whilst the third is the wanted signal.

Let the powers due to the second, third (signal), and fourth terms be p_2, p_3 and p_4 respectively.

Since only power ratios will ultimately be used, as a matter of convenience powers will be calculated by treating the terms of equation (4) as voltages and assuming a load of unity nominal resistance. Subject to these conditions, the powers averaged over a period T are given by

$$\begin{aligned} p_2 &= a^2.\frac{T_0}{T}.\left(1-\frac{T_0}{T}\right) \\ p_3 &= \frac{b^2}{2}.\left(\frac{T_0}{T}\right)^2 \\ p_4+p_3 &= \frac{b^2}{2}\frac{T_0}{T} \end{aligned} \tag{5}$$

Hence, approximately,

$$p_4 = \frac{b^2}{2}\frac{T_0}{T}\left(1-\frac{T_0}{T}\right) \tag{6}$$

The total signal power $p_s = p_3$, whereas the total noise power p_n is

$$\begin{aligned} p_n &= p_2+p_4 \\ &= \left(a^2+\frac{b^2}{2}\right).\frac{T_0}{T}.\left(1-\frac{T_0}{T}\right) \end{aligned} \tag{7}$$

Thus the signal/noise power ratio R is

$$R = \frac{p_s}{p_n} = \frac{b^2}{2a^2+b^2}.\frac{T_0}{T-T_0} \tag{8}$$

In the case of suppressed carrier transmissions, the constant $a = 0$, hence in this case the signal/noise ratio is R_s, where

$$R_s = \frac{T_0}{T-T_0} \tag{9}$$

This ratio is plotted in Fig. 2: it is large for small blanking periods but falls to 0 dB when the blanking time is half the repetition period.

Where the carrier is not suppressed a normal a.m. signal results, and it is convenient to express the signal to noise ratio, R_{AM}, in terms of a modulation index m, where

$$m = \frac{b}{a} \qquad 0 < m < 1$$

Thus

$$R_{AM} = \frac{m^2}{2+m^2}.\frac{T_0}{T-T_0} \tag{10}$$

It will be seen that suppressed carrier transmissions are always at an advantage relative to a.m.; when the latter is 100% modulated this advantage is 5 dB, but at lower, and more normal, modulation indices the advantage is considerably greater.

The results will apply *pari passu* to a more complex modulating waveform such as speech, but two additional points must be made. The nature of speech is such that if peaks are to be accommodated the average modulation index for a.m. will be considerably less than 100%. Secondly, blanking continues to produce noise on a.m. even between words, where no modulation is transmitted. In suppressed carrier systems the receiver has no noise output at these times (ignoring adjacent channel effects to be considered in a subsequent section).

The actual value of blanking time $(T-T_0)$ chosen must be greater than the duration of the noise pulse appearing at the blanker. This will depend principally on the bandwidth and response shape of the band-pass filter preceeding the blanker. Such a filter will both delay and lengthen the pulse entering. Filters of the Butterworth or (worse) Chebyshev type commonly used in i.f. amplifiers are ill suited to this application, having outstandingly poor transient response, with a pulse extension which may amount to very many times the reciprocal of the half bandwidth.

A highly selective filter will require a long blanking period, with relatively poor signal/noise ratio due to blanking signal in the wanted channel. However, such a filter will virtually eliminate signals entering the blanker due to adjacent channels. As will be demonstrated subsequently, such adjacent channel signals can also contribute to noise in the receiver output. To pursue this point further it will be necessary to consider the effect of such signals.

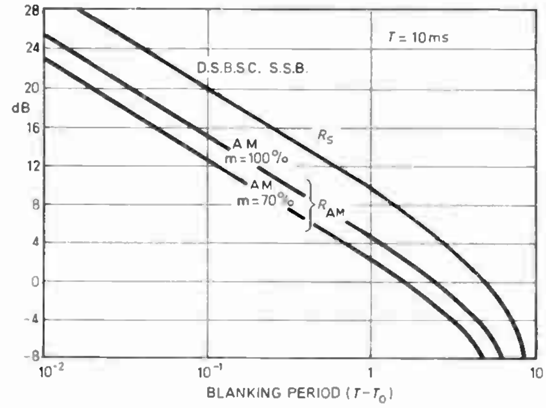

Fig. 2. Computed signal/noise ratios in the absence of an adjacent channel signal.

4 Adjacent Channel Effects

Signals in channels adjacent to that in which the wanted signal occurs also give rise to noise signals in the receiver output due to the effects of blanking. The consideration of this effect will, for simplicity, be developed for the case of an unmodulated carrier on the adjacent channel. For such a carrier, after blanking the waveform resulting is $u(t)$ where

$$u(t) = a' \sum_{-\infty}^{\infty} \frac{1}{n\pi} \sin\left[n\pi\left(\frac{T_0}{T}\right)\right] \times$$

$$\times \cos\left(2\pi n\frac{t}{T}\right).\cos\left[(\omega_c+\omega_s)t\right]$$

subject to the above conditions for $0 \leqslant u(t) < \frac{1}{2}$

where ω_s is the frequency difference between adjacent channels

and a' is the amplitude of the carrier.

Each term corresponds to a pair of sidebands spaced above and below the carrier at $(\omega_c+\omega_s)$, but only the lower sidebands can fall within the pass band of the wanted channel. Thus the spectrum of the lower set of sidebands is equivalent to $u''(t)$, where

$$u''(t) = \frac{a'}{2}\sum_{1}^{\infty}\frac{2}{n\pi}\sin\left[n\pi\left(\frac{T_0}{T}\right)\right]\times$$
$$\times\cos\left[\left(\omega_c+\omega_s-\frac{2n\pi}{T}\right)t\right] \qquad (11)$$

It is of particular interest to consider the component $u(\omega_c)$ centred at the frequency of the wanted carrier, for which

$$\omega_s - \frac{2n\pi}{T} = 0$$

or

$$n = \frac{\omega_s T}{2\pi}$$

Hence

$$u(\omega_c) = \frac{a'}{\omega_s T}\sin\left(\frac{\omega_s}{2}T_0\right)\cos(\omega_c t)$$

The sine function can have any modulus up to 1: in what follows this maximum value will be assumed, as giving the worst case. If the component at ω_c due to the adjacent carrier is to equal the wanted carrier component after blanking, the ratio of carrier amplitudes in the wanted and adjacent channels is

$$\frac{a'}{a} = \frac{\omega_s}{2}(TT_0)^{\frac{1}{2}} \qquad (12)$$

This ratio should preferably be large, because satisfactory operation is usually essential with considerably stronger signals in the adjacent than in the wanted channel.

5 Choice of Blanking Period

Since both the signal/noise ratio due to effects of blanking on signals in the wanted channel (eqn. (8)) and also the ratio of permissible adjacent to wanted channel signal amplitude increase with T_0 (the un-blanked period), the shortest possible blanking time should be adopted. What, then, is the lower limit to this parameter?

The blanking period must be sufficiently long to ensure that all disturbance due to the noise pulse is over at the input terminal of the blanker before the blanking period is complete. This period must therefore equal the sum of the delay time (T_D) between the detection of a noise pulse in the noise channel and its appearance at the blanker input, and the duration (T_P) of the noise pulse at that point. Both T_D and T_P are dependent on the characteristics of the delay device and the filter which preceed the blanker in the signal channel. A highly selective filter will lengthen both T_D and T_P. From the point of view of adjacent channel rejection (desirable both to reduce adjacent channel blanking effects and also the risks of cross modulation), the filter should be as narrow as possible, however, use of a narrow filter is ruled out by conflict with the need to keep T_0 large.

For example, in a d.s.b.s.c. system, if no filter is used the blanking period can be reduced, although for other reasons (to be discussed subsequently) probably not below about 50 μs. This would yield a signal/noise ratio of 16 dB at a noise repetition frequency of 500 Hz (T = 2 ms). The adjacent channel rejection (12·5 kHz channelling) would be 38 dB.

6 Trapezoidal Blanking

So far it has been assumed that the blanking function $y(t)$ is rectangular. Advantages may, however, be obtained by making it trapezoidal, i.e. causing the insertion loss of the blanker to increase over a non-zero time on receipt of the blanking control voltage. The Fourier components of a trapezoidal function decline more rapidly than those of a rectangular function, giving advantages in respect of adjacent channel effects particularly.

The trapezoidal blanking function will be assumed to have a value of unity for a time T_0, as in the rectangular case, but the transitions between T and 0 will be assumed to take place linearly over a time T_1. Thus the period for which the function is zero will be $(T-T_0-2T_1)$. Hence the blanking function may be expressed as $y(t)$, where

$$y(t) = t/T_1; \qquad 0 < t < T_1$$
$$y(t) = 1; \qquad T_1 < t < T_1+T_0$$
$$y(t) = 2-\frac{t-T_0}{T_1}; \qquad T_1+T_0 < t < 2T+T_0$$
$$y(t) = 0; \qquad 2T_1+T_0 < t < T$$

The expression of $y(t)$ as a Fourier series is quite straightforward:

$$y(t) = \sum_{-\infty}^{\infty}\frac{1}{n\pi}\sin\left(n\pi\frac{T_0+T_1}{T}\right)\times$$
$$\times\frac{\sin\left[n\pi\left(\frac{T_1}{T}\right)\right]}{n\pi\left(\frac{T_1}{T}\right)}.\cos\left(2n\pi\frac{t}{T}\right)$$
$$= \frac{T_0+T_1}{T} + \sum_{1}^{\infty}\frac{2}{n\pi}\sin\left(n\pi\frac{T_0+T_1}{T}\right)\times$$
$$\times\frac{\sin\left[n\pi\left(\frac{T_1}{T}\right)\right]}{n\pi\left(\frac{T_1}{T}\right)}.\cos\left(2n\pi\frac{t}{T}\right)$$

If $T_1 < T_0$, then so far as the effect of blanking on the wanted signal is concerned,

$$\frac{\sin\left[n\pi\left(\frac{T_1}{T}\right)\right]}{n\pi\left(\frac{T_1}{T}\right)} \to 1; \qquad n,\ T_1 \text{ small}$$

Thus the effect of blanking the wanted signal with a trapezoidal waveform is not significantly different from that of rectangular blanking except that T_0 is replaced by (T_0+T_1).

For larger n, however, as in the case of sidebands generated by blanking signals in the adjacent channel,

$$\frac{\sin\left[n\pi\left(\frac{T_1}{T}\right)\right]}{n\pi\left(\frac{T_1}{T}\right)} < \frac{1}{n\pi\left(\frac{T_1}{T}\right)}$$

$$< \frac{2}{\omega_s T_1}$$

Thus, provided that

$$T_1 > \frac{2}{\omega_s} \qquad (13)$$

an advantage in adjacent channel rejection will be obtained using trapezoidal blanking. The rejection in this case will be given by the approximate expression (comparable with equation (12) above):

$$\frac{a'}{a} = \frac{\omega_s^2 (T T_0)^{\frac{1}{2}}}{2} \cdot \frac{\left(\frac{T_1}{2}\right)}{\sin\left[\omega_s\left(\frac{T_1}{2}\right)\right]} \qquad (14)$$

The expression for noise due to in-band blanking will now be (from equation (8), substituting (T_0+T_1) for T_0):

$$R = \frac{b^2}{2a^2+b^2} \cdot \frac{T_0+T_1}{T-(T_0+T_1)} \qquad (15)$$

It is of interest to consider the optimum value of T_1 relative to T_0 and T, which will maximize a'/a. Since T_1 and T_0 are interdependent, it is helpful to rewrite equation (12) in terms of new variables. If the blanking time (i.e. time for zero signal transmission) is T_B, then

$$T_0 = T - 2T_1 - T_B \qquad (16)$$

Hence

$$\frac{a'}{a} = \frac{\omega_s}{2}[T_c T - 2T_1 - T_B]^{\frac{1}{2}} \cdot \frac{\left[\omega_s\left(\frac{T_1}{2}\right)\right]}{\sin\left[\omega_s\left(\frac{T_1}{2}\right)\right]} \qquad (17)$$

Taking the worst case, in which the sine function has the value unity and differentiating

$$\frac{d}{dT_1}\left\{\frac{a'}{a}\right\} = \frac{\omega_s^2 T^{\frac{1}{2}}}{4}\left[(T-2T_1-T_B)^{\frac{1}{2}} - T_1(T-2T_1-T_B)^{-\frac{1}{2}}\right]$$

$$= 0 \text{ if } T_1 = \tfrac{1}{3}(T-T_B)$$

This can easily be shown to be the condition for a'/a to be a maximum, and, from equation (16) leads to a value of $T_0 = T_1 = \frac{1}{3}(T-T_B)$. For this value of T_1 the ratio a'/a is very large:

$$\left[\frac{a'}{a}\right]_{\max} = \frac{\sqrt{3}}{4}\omega_s^2 T^{\frac{1}{2}} T_0^{\frac{3}{2}}$$

Hence if, as is usual, $T_B \ll T$, to good approximation

$$\left[\frac{a'}{a}\right]_{\max} \simeq \frac{\omega_s^2 T^2}{12} \qquad (18)$$

For example, in the case considered above for which the channel spacing is 12·5 kHz and T is 2 ms, the unwanted carrier rejection is as high as 67 dB. In fact even this calculation is pessimistic, since the sine function of equation (14) has been assumed equal to unity, and will in fact be less, giving a few decibels further improvement.

Practically, the use of the optimum value for T_1 has two serious drawbacks. The first of these is that the signal/noise ratio resulting from the blanking of the wanted (in-band) signal will be relatively poor. This makes the use of shorter times T_1 obligatory, in order to maintain acceptable S/N ratios due to wanted signal blanking. Secondly, it is obvious that the delay in the signal channel before the input to the blanker (Fig. 1) must exceed T_1, since otherwise blanking is not complete by the time a noise pulse arrives. For T having typical values of a few milliseconds, a delay of the order of a millisecond will be required if T_1 is to have its optimum value. Delay lines of this magnitude operating at i.f. are not available, and hence T_1 of this magnitude cannot be employed.

Furthermore, any delay line, unless so perfectly matched at input and output that there is literally no energy reflexion at the ports, will in fact give additional outputs at odd multiples of the nominal delay period due to reflected energy making 3, 5 etc. traverses of the line. These spurious responses can be made small, but to avoid at least the first it would be desirable for the blanking period to be greater than three times the delay period, so that the first spurious response will be blanked as well as the normally delayed noise pulse. Thus a long delay would result in an even longer blanking period, with correspondingly very poor signal/noise ratios.

Thus, for these reasons, T_1 will normally be made very much smaller than the optimum value, the choice being largely determined by the availability of delay components. In an experimental receiver described in a later section the delay component used is a 64-μs glass ultrasonic delay line, designed originally for colour television receivers.

The use of trapezoidal blanking has another advantage beyond those already considered. Practical blankers consist either of voltage controlled attenuators or variable gain amplifiers, and in either event usually depend on a change of d.c. level by up to several volts at some point in the circuit in order to introduce the required signal attenuation. This change of d.c. level corresponds to a waveform which may have Fourier components within the pass band of the receiver. This d.c. transient effect can much more easily be eliminated in the trapezoidal case.

Calculated S/N ratios for receivers using trapezoidal blanking are shown in Fig. 3. For suppressed carrier transmissions the S/N ratio does not fall below 10 dB in the absence of adjacent channel signals even with p.r.f.s as high as 500 Hz. However, at this p.r.f. adjacent channel discrimination is only 35 dB.

7 Experimental Results and Discussion

A receiver has been constructed using an ultrasonic glass delay line giving approximately 64 μs delay, and in accordance with Fig. 1, the bandwidth of the pre-blanker i.f. filter being 100 kHz, whilst the post-blanker filter is a

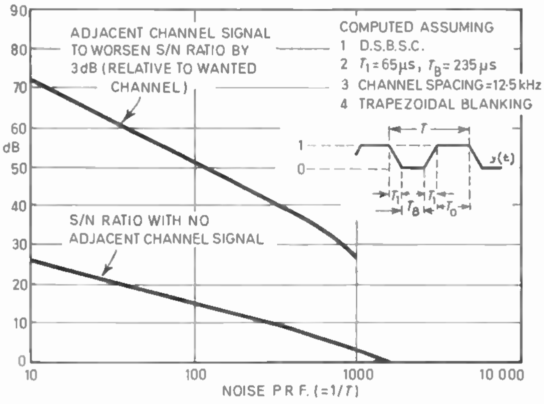

Fig. 3. Effect of adjacent channel signal.

conventional s.s.b. crystal i.f. filter (pass-band 2·5 kHz wide). The intermediate frequency is 5·2 MHz.

Preliminary results only are available at the time of writing, and although these are in general agreement with the theoretical treatment given, detailed confirmation has been complicated by the fact that the blanking pulse, although already relatively slow compared with a noise pulse, is extended by the subsequent narrow i.f. filter. This may be seen in Fig. 4. The upper trace is a simulated noise pulse injected into the noise channel, and the second trace shows the consequent blanking pulse applied to the gate of the f.e.t. noise blanker. This pulse is rectangular and of 200 μs duration. The third trace is the blanked i.f. before entering the narrow i.f. filter, and the bottom trace is the demodulated a.f. signal. A 'smoothed' version of the blanking pulse is observed, but accompanied by a damped 'ring' at about 600 Hz of about 1½ cycles duration, and a higher frequency 'ring' at about 3 kHz of many cycles duration.

Trapezoidal blanking (Fig. 5) almost entirely eliminates the latter effect, but the lower frequency 'ring' remains, and will typically worsen observed S/N ratios by 3–4 dB, compared with predicted values. The delay line receiver is a significant step in the direction of a truly noise-resistant receiver, but does not provide a complete answer.

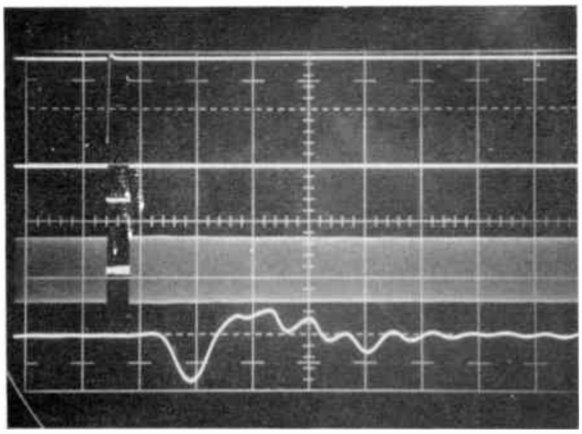

Fig. 4. Effect of rectangular blanking (1 large horizontal division = ½ ms).

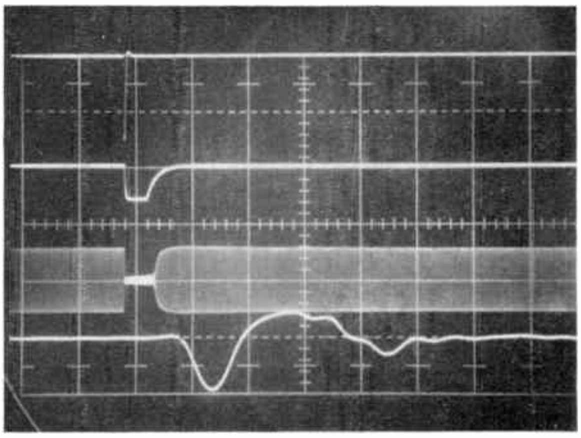

Fig. 5. Trapezoidal blanking: note virtual elimination of high frequency ring on the bottom trace.
(1 large horizontal division = ½ ms).

Fundamentally, the difficulty arises because at some stage the received signal must pass through a narrow band i.f. filter, and such a filter cannot be designed both to give the required adjacent channel selectivity and also to have a good response to transients, whether noise pulses or blanked breaks in the received signal. The blanker reduces the amplitude of the transient to equal the instantaneous amplitude of the received signal, and this is its advantage over a receiver not equipped with noise reduction circuits, but a serious transient still remains.

A possible approach to this problem is to arrange a system which reduces the magnitude of the blanking transient, for example by holding the signal amplitude constant during the blanking period rather than reducing it to zero. Since blanking operations must be performed at i.f., and at a signal level of only tens of microvolts, circuits which will perform this function adequately present serious difficulties.

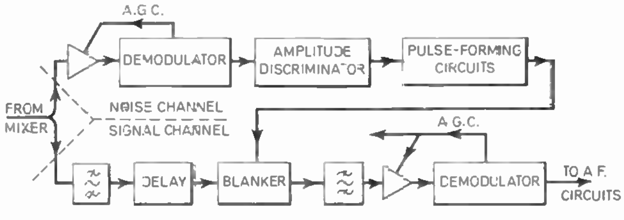

Fig. 6. An alternative noise-reducing receiver design.

One solution currently under investigation, is indicated in Fig. 6. The received signal normally passes through a broad-band delay line, but when a noise pulse is detected the delay line is switched out of circuit and the signal passes directly to the subsequent receiver circuits. Provided that the delay is short compared with the shortest modulation period, there will be little change in the signal amplitude, hence the transient superimposed on the incoming signal will be much reduced.

8 Conclusions

Although impulsive noise is quite different in character from d.s.b. or s.s.b. radio signals, the design of a receiver which will discriminate against the former, and in favour of the latter, is not easy.

It is critically important that the noise blanking operation should be carried out on the received signal before it passes through the narrow i.f. filter, and hence blanking must operate at very low level. The use of a signal delay line is essential, both to permit the blanking circuit to come into operation before any of the noise pulse energy reaches the following filter, and also to allow the operation of the blanker to be relatively slow, thus reducing the component of i.f. energy passed to the filter which is due to the blanking operation itself.

Results obtained experimentally suggest that further progress will be made only if some means of noise cancellation less crude than simple blanking can be employed. The aim of any new approach must be to reduce the disturbance to the wanted signal caused by the operation of noise cancelling circuits.

9 Acknowledgments

This work formed a part of a research programme supported by the Home Office, and the author wishes to thank the Director of Telecommunications, Home Office, for permitting its publication.

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British Components Industry Reorganizes

The Electronic Components Board has been reorganized and, with effect from 1st January, 1973, the separate constituent Associations responsible for active electronic components have ceased to exist as separate entities. The two Groups of VASCA covering professional valves and tubes, and semiconductor devices have become product groups of the new organization, and a third product group is responsible for domestic valves and television tubes, taking over the function of the BVA.

There will be two categories of members:

(a) Direct Members, being Companies engaged in the UK in the manufacture and sale of electronic components, and who are approved by the Council of the Board, who will pay an inclusive subscription direct to the ECB.

(b) Corporate Members, who will be members of the RECMF, which will pay a block subscription to the ECB on their behalf.

Thus the RECMF will continue as a separate autonomous but affiliated organization under their Director and Secretary, A. C. Bentley (Companion).

The objects of the Board are:

(i) To promote, foster and protect the interests of the British Electronic Components Industry.

(ii) To provide a united organization to speak for the Electronic Components Industry with one voice to the UK Government, to other sectors of the Industry and to other trade associations, both at home and overseas.

(iii) To provide a forum in which all matters appertaining to the Electronic Components Industry can be fully discussed, and where necessary coordinated views ascertained for appropriate action.

(iv) Generally to maintain and improve the image and prestige of the British Electronic Components Industry at home and abroad.

It is emphasized that the demise of the separate active components Associations at the end of the year in no way affected the activities previously carried out on behalf of their member Companies which are now transferred in their entirety to the Electronic Components Board Product Groups.

The Chairman of the Board will be Mr. A. Deutsch, Technical Director of the Thorn Group. The Director of the Electronic Components Board is Sir Ronald Melville, K.C.B., who was appointed on 1st May, 1972, and Mr. Michael Mason, whose connection with VASCA and BVA goes back to 1931 remains as Secretary. Mr. P. A. Fleming is Engineering and International Secretary.

Engineering Consultant to the Board, a new post, will be Mr. K. G. Smith. Mr. Smith who is a Past Chairman and Vice President (1960-1964) of the RECMF was, until his retirement at the end of 1971, Joint Managing Director of NSF Limited and also a Director of Simms Motor and Electronic Corporation Ltd., and other companies in the Lucas Group.

The offices of the Electronic Components Board and of the Radio & Electronic Component Manufacturers Federation are now at Liberty House, 222 Regent Street, London W1R 5EE. The new telephone number is: 01-437 4127.

IERE News and Commentary

Advance Notice of Institution Dates

The Annual General Meeting of the Institution will be held at the London School of Hygiene and Tropical Medicine, Keppel Street, Gower Street, London WC1, on Thursday, 25th October 1973, at 6 p.m. Formal notice and Agenda for the meeting will be published shortly.

Members of the Council and of Standing, Specialized Group and Local Section Committees and representatives of the Institution on outside bodies may like to note that the Dinner of Council and its Committees will take place at the Savoy Hotel, London, on Wednesday, 21st November 1973. Further details for this popular social evening, at which members may be accompanied by ladies, will be announced in due course.

1973 Survey of Professional Engineers

Following the recent CEI Board decision that a further Survey of Professional Engineers should be conducted in 1973, a Steering Committee has been set up under the chairmanship of Mr. B. Hildrew (representing the Institutions of Gas, Mining and Production Engineers, the Institute of Marine Engineers, the Institution of Mining and Metallurgy, the Royal Aeronautical Society, and the Royal Institution of Naval Architects). He is assisted by Mr. H. J. Foxcroft (representing the Institution of Chemical Engineers, and the Institute of Fuel), Mr. T. M. D. James (representing the Institutions of Civil, Municipal, and Structural Engineers), Mr. E. R. L. Lewis (representing the Institutions of Electrical, and Electronic and Radio Engineers) and Mr. Llewellyn Young (representing the Institution of Mechanical Engineers).

The main task of the Steering Committee has been to design the questionnaire leaflet which will be distributed to members of the IERE in Great Britain with their copies of the June issue of *The Radio and Electronic Engineer*.

Overseas Recognition of the IERE

Advice has been received from the Registrar of the Professional Engineers Board, Republic of Singapore, that Corporate Members of the Institution who have been admitted by a recognized degree or diploma or by passing the CEI Examinations are eligible for consideration, under the terms of the Professional Engineers Act of 1970, for registration as Professional Engineers in Singapore.

Microwave 73

The technical conference programme for Microwave 73 to be held in Brighton from 19th to 21st June next has now been published. The Conference takes place in the Hotel Metropole as does the associated Exhibition.

In the Systems sessions such topics as satellites and specialized terrestrial communications systems, air traffic control in the 80s, computer-controlled microwave receiving systems, c.a.d., and automated test equipment, as well as industrial and commercial applications, are being covered.

Main subjects of the Modules sessions—which are being held in parallel with the Systems sessions—are antennas, broadband microwave modules, advances in solid-state devices, microwave tubes, and surface acoustic waves.

Over 70 papers from nine countries are scheduled to be presented at the conference, and 150-word abstracts of each are included in the booklet.

Professor J. H. Collins of the University of Edinburgh is chairman of the conference. Session chairmen include Professor D. E. N. Davies, (University College London), Dr. W. Veith (Siemens AG, Munich), Dr. D. Dorsi (GTE, Milan), Professor P. Weissglas (Microwave Institute Foundation, Stockholm), and Mr. J. Gremilles (Thomson-CSF, Orsay).

The conference, which has been organized in association with the Institution of Electronic and Radio Engineers, is to be officially inaugurated by H.R.H. the Duke of Kent on the Tuesday afternoon. IERE members benefit from a reduced conference fee of £25, which includes a full set of three days' preprinted proceedings; the equivalent fee for other delegates is £30.

The programme and registration forms are obtained from Microwave Exhibitions & Publishers Ltd., 21 Victoria Road, Surbiton, Surrey. (Telephone 01–390 0202).

An Industry Golden Jubilee

Belling & Lee Ltd., the electronic components manufacturing Company formed by C. R. Belling and E. M. Lee celebrated its Golden Jubilee in December last.

In the early 'twenties, the Company started manufacturing radio receivers incorporating such contemporary advanced techniques as variable selectivity, tone control and remote control and steadily progressed to the design and production of electrical components. In 1933, an electrical gramophone pick-up was produced and the Company entered the radio frequency interference suppression market with a capacitor mains filter, followed later by heavy duty inductors.

The introduction of a regular television service in the mid-1930s widened the Company's activities in the electrical components business. Later the main contributions to the war effort by Belling-Lee were components such as connectors using the well-known vibration-proof O-Z construction, fuseholders and valve holders, including the famous B9A for the EF50 valve, and aircraft aerials.

With the re-opening of the BBC television service after the War the Company expanded enormously and opened up a special environmental test laboratory to ascertain that its products met the requirements of industrial standards. The success in the U.K. caused the Company to seek further avenues for expansion and an Australian company, Belling & Lee (Australia) Pty. Ltd., was registered in December 1955.

In 1966, following the death of Mr. Belling, a controlling interest in the Company was acquired by Ada (Halifax) Ltd., a holding company of the Philips Group.

The Company now has many divisions and the TV Equipment Division, in close co-operation with the Philips Group in Eindhoven, designs, manufactures and markets a comprehensive range of cable TV equipment.

Mr. Lee, one of the two founder directors of the Company, who died just six months short of the Company's 50th anniversary, was a Fellow of the Institution for nearly 30 years and for many years represented the Institution on BSI Committees. (See *Journal* for July 1972, p. S.98).

LONDON MEETINGS

Wednesday, 23rd May

COMMUNICATIONS GROUP MEETING

'Optimum Linear and Non-linear Transversal Equalizers'

By Dr. A. P. Clark (*Loughborough University of Technology*)
IERE Lecture Room, 6 p.m. (Tea 5.30 p.m.)

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

Thursday, 14th June

JOINT IERE/IEE MEDICAL AND BIOLOGICAL ELECTRONICS GROUP IN COLLABORATION WITH THE BRITISH INSTITUTE OF RADIOLOGY, THE HOSPITAL PHYSICISTS' ASSOCIATION, AND THE BIOLOGICAL ENGINEERING SOCIETY.

Colloquium on
X-RAY IMAGE INTENSIFIER SYSTEMS

University College London, 11 a.m. to 5.15 p.m.
Advance Registration necessary. £2.20 to members of sponsoring bodies; £4.40 to non-members (prices include VAT). Applications to Meetings Secretary, IERE.

Part of the meeting will be devoted to recent developments in X-ray intensifier tubes with caesium iodide phosphors, and associated systems, and leading experts in Europe have agreed to contribute. The remainder of the meeting will be concerned with more practical, but nevertheless important, aspects of the use of X-ray image intensifier systems in hospitals, for example in testing their performance. The meeting will conclude with a panel discussion which will provide an opportunity for radiologists, physicists and engineers to consider the advantages and limitations of X-ray image intensifier systems, and the ways in which further improvements in performance might be achieved and monitored.

Introductory Papers

'Medical Requirements in X-ray Image Intensification'
By Dr. G. R. Airth (*Southmead Hospital*)

'The Mechanisms and Limitations of X-ray Image Intensifier Systems'
By G. A. Hay (*University of Leeds*)

Experience with X-ray Intensifier Systems in Hospitals

'Checking System Efficiency with special reference to X-ray Beam Quality'
By Dr. G. M. Ardran (*Nuffield Institute for Medical Research*) and H. E. Crooks (*A.E.R.E.*)

'Practical Difficulties of Routine Testing'
By Dr. M. Davison (*Western Regional Hospital Board, Glasgow*)

'Test-object Design for Routine System Evaluation'
By G. A. Hay (*University of Leeds*)

'Simple Apparatus for Checks on X-ray Image Intensifiers'
By P. Jilbert (*A.W.R.E.*)

Design Aspects of X-ray Intensifier Systems

'Technical Problems associated with High Speed Pulsed Cinefluorographic Systems'
By O. F. Clarke (*University of Leeds*)

'Recent Advances in X-ray Image Intensifiers'
By Professor F. Gudden (*Siemens AG*)

'Design Aspects of New Image Intensifier Tubes'
By Dr. W. Kuhl (*NV Philips Gloeilampenfabrieken*)

'Evaluation of X-ray Imaging in Terms of Transfer Function'
By F. Timmer (*NV Philips Gloeilampenfabrieken*)

Obituary

The Council has learned with regret of the deaths of the following members, all of whom died last autumn although news was not received by the Institution until last month.

Mr. Frank Arthur Leake (Associate 1950, Student 1947) died on 9th September 1972, aged 51 years. A Londoner, by birth, Mr. Leake's first appointment was with Standard Telephones & Cables, New Southgate, in 1937, where he was a Senior Laboratory Assistant prior to joining the R.A.F. in 1943. He returned to STC as a Test Gear Engineer on demobilization in 1947. He then attended the Northern Polytechnic from 1947 to 1949, obtaining a Diploma in Radio Engineering. An appointment as an Engineer on test gear development with EMI at Hayes followed and in 1950 he joined the Air Technical Publications Directorate of the then Ministry of Supply at Chessington as a Technical Author. In 1966 Mr. Leake was appointed as Safety Adviser on Electronic and Electrical Equipment in the Ministry of Technology's Safety Service Organization, where he was concerned with all aspects of electronic equipment ranging through guided weapons to noise problems. He represented the Organization on a B.S.I. Committee and, although dogged by ill-health, was known to his colleagues as an expert and as a tireless worker who was always prepared to help others.

Mr. Leake leaves a widow.

Major Samuel Reginald Rickman, T.D., R.Sigs. (Retd) (Member 1949) died on 7th October 1972, at the age of 65.

Major Rickman was at Sherborne School from 1920 to 1924, and in the following two years he was at Birmingham University. During the twelve years prior to the War, he was employed in the electrical engineering and quarrying departments of the Clay Cross Co. in Derbyshire. He was commissioned in the Royal Corps of Signals in 1939, serving as an Instructor from 1941 to 1943, then as Deputy Assistant Director Research and Development from 1943 to 1944 with the Ministry of Supply. Major Rickman then returned to instructional duties at Catterick, being appointed Chief Instructor, Signals Wing in 1946 and subsequently as Technical Staff Officer at Catterick. He presented papers on modern communications techniques at North Eastern Section Meetings and served on Council from 1952 to 1954. This was followed by a tour of duty in Malta from 1955 to 1957, and an appointment at the War Office. He retired to live in Sussex and will be remembered particularly for his active participation in educational matters affecting his Corps.

He leaves a widow.

Mr. Walter Bryan Savage (Fellow 1938, Member 1936) died on 7th October 1972, aged 76 years.

Mr. Savage was awarded a Mathematical Scholarship at Eastbourne College from 1910 to 1914, and was commissioned in 1915, serving from then until 1919 in the Queen's Royal West Surrey Regiment, the Machine Gun Corps and as an Inspector of Munitions. A keen radio amateur, he formed his own radio and sound engineering business in London in 1929, using the trade name 'Savage Sound'. This eventually became W. Bryan Savage Ltd. The Company moved to Wiltshire in the early part of the Second World War and Mr. Savage formed a new company, Savage Transformers Ltd., at Devizes in 1945. He retired from active participation in the business in the late 1960s.

He leaves a widow who had always been associated with him as a co-Director in his Companies.

Members' Appointments

CORPORATE MEMBERS

Mr. W. V. Richings (Fellow 1968, Member 1967) has recently been appointed General Manager, Dawe Instruments Ltd. Mr. Richings joined Dawe Instruments as a Development Engineer in 1945 and became Chief Development Engineer and then Chief

Engineer before being promoted to Technical Director in 1966. He has been active on British and International Committees concerned with ultrasonics and with noise and vibration measurement, and he has been a member of IERE Group and Conference Organizing Committees.

Mr. F. E. Parr, B.Sc.(Eng.) (Fellow 1964, Member 1951) has recently been appointed Naval Weapons Production Overseer (Southern Area) in the Ministry of Defence (Navy Department). He was formerly Head of Technical Services with the Metrication Board and for a number of years served on the Southern Section Committee as Programme Secretary.

Mr. J. Hockley (Member 1972) has been promoted Senior Scientific Officer with Government Communications Headquarters, Cheltenham. Mr. Hockley joined GCHQ in 1964 as an Experimental Officer.

Mr. F. Haslam (Member 1966) will shortly be leaving his present post as a Senior Lecturer in the Electrical Engineering Department of the City of Birmingham Polytechnic to take up the appointment of Head of the Department of Electrical Engineering, Matthew Boulton Technical College, Birmingham.

Mr. P. Crossland (Member 1972, Graduate 1965) has taken up a one year assignment at IBM's manufacturing plant in Boulder, Colorado, U.S.A. Mr. Crossland joined IBM (UK) Ltd. in 1970 as Associate Engineer in their Manufacturing Research Laboratories in Hampshire.

Squadron Leader R. B. Game, RAF (Member 1971) has been appointed Officer in Charge RTPS, RAF North Luffenham, Oakham, Rutland. Sqn. Ldr. Game was formerly stationed at RAF Medmenham, Bucks.

Mr. B. Scott (Member 1968), previously with the Television Equipment Division of E.M.I. Sound and Vision Ltd., has been appointed Project Manager on the Regional Broadcasting Training Centre project for the Malaysian Government in Kuala Lumpur.

Wing Commander D. W. Hills, RAF (Member 1962), formerly with the Ministry of Defence S(OPS)1, Central Staff, is now Chief, Advanced Traffic Control and Landing Systems Division, HQ Air Force Communications Service, USAF, Missouri, USA.

Mr. R. Cooke (Member 1972, Graduate 1969) has joined Mimic Electronics Ltd., Dartford, as Chief Engineer, Digital Monitoring Equipment. Mr. Cooke was previously a consultant with Y-ARD (Consultants) Ltd., Glasgow.

Mr. A. S. Omosule (Member 1972, Graduate 1971) who was formerly Engineer-in-Charge (Installation), with the Nigerian Broadcasting Corporation in Lagos, has been appointed Lecturer in Electrical Engineering at The Polytechnic, Ibadan.

Mr. A. J. Kenward, B.Sc. (Member 1959), Secretary of the Society of Electronic & Radio Technicians, has been appointed a member of the Technician Education Council by the Secretary of State for Education and Science, Mrs. Margaret Thatcher. Mr. Kenward was for 16 years on the staff of the IERE as Education Officer

responsible for all membership, education and examination activities. He has been Secretary of SERT since its formation in 1965 and of the Radio, Television and Electronics Examination Board since the same date. He is currently chairman of the Technician Engineer Section of the Engineers' Registration Board.

Mr. H. B. Maughan (Member 1965, Graduate 1961), who joined Sevcon Engineering Ltd., Gateshead, in 1966 as a Senior Development Engineer, to become Works Manager shortly afterwards, has now been appointed Director of Manufacturing of the Company. Mr. Maughan has been a member of the North Eastern Section Committee for several years.

NON-CORPORATE MEMBERS

Mr. D. M. Dawson, B.Sc. (Graduate 1972) who was previously with Plessey Memories (Towcester) as a Graduate Trainee/Memories Design Engineer, is now a Power Engineer with Smiths Industries Ltd., Cheltenham.

Mr. G. K. Atigh (Graduate 1972) has been appointed Deputy Technical Director in the Iranian Meteorological Department Tehran.

Squadron Leader L. A. Bull, M.B.E., RAF (Associate 1967) who was posted to the Ministry of Defence (Air) as a Special Project Officer last year, has been appointed Officer Commanding Electronic Engineering Squadron, Ground Radio Servicing Centre (GRSC), RAF North Luffenham, Rutland.

Mr. D. R. Edwards (Graduate 1967, Student 1963) recently completed the M.Sc. course in microwave and solid state physics at Portsmouth Polytechnic and has now taken up a teaching post at Sydney Technical College. Previously Mr. Edwards spent three years in Africa training Zambia Police Signals technicians.

Mr. D. J. Cousins (Graduate 1962) who joined Sundstrand Aviation, Rockford, Illinois, as Senior Project Engineer and Leadman for the Electrical Controls Group, has been appointed Group Engineer, Test Equipment Section.

Mr. D. Hudson (Graduate 1970) formerly a Senior Electronics Technician at Bradford University and a Yorkshire Section Committee member, has been appointed Laboratory/Workshop Superintendent in the Department of Psychology at Brunel University.

Mr. S. M. A. Phillips (Graduate 1970) has been promoted and transferred to the Headquarters of the Home Office Directorate of Telecommunications as a Design Engineer. Mr. Phillips was formerly Resident Engineer for the police and fire services in Gloucestershire, based in Cheltenham, and was latterly promoted to a position at H.O.M.U., Bishop's Cleeve.

Mr. A. E. Sloan (Associate 1972, Graduate 1967) has been appointed Director of Engineering at Sevcon Engineering Ltd., Gateshead. Mr. Sloan, who joined the Company at its formation in 1963 as Senior Development Engineer, was promoted to Chief Development Engineer and later to Technical Manager.

Mr. N. Weisbloom (Graduate 1959) previously Technical Manager of the Government Electronics Division, Mullard Limited, now joins the Computer Electronics Division as Group Product Manager for Electronic Assemblies.

New Books Received

All the books which are described below are available in the Library and may be borrowed by members in the United Kingdom. A postal loan service is available for those who are unable to call personally at the Library

Electronic Engineering Processes

C. E. Jowett. Business Books Ltd., London 1972. 23·4 cm × 15·4 cm. 215 pp. £6·30.*

Contents: Section 1: Process Standard for Integrated Circuits. Coating and Encapsulation Semiconductor Manufacture. Gold-Plating Process for Semiconductor Manufacture. Process for Silicon Integrated Circuits. Thin-Film Circuit Production Process. Chemical Processes in Semiconductor Manufacturing. Section 2: Process for Assembly of Hybrid Circuits into Flatpacks. Processes in Modular Packaging Systems. Section 3: Electronic Packaging Processes with Plastics. Surface Preparation when using Adhesives. Coating of Electronic Assemblies using Polyurethane. Application of Nylon Coating by Fluidized Powder. Packaging Electronic Components. Section 4: Printed Wiring Board and Circuits. Multilayer Printed Wiring Board Plating Process. Plated Through-Hole Process. Weldable Printed Circuit Boards with Plated Through-Holes. Joining Processes. Electrical Wiring Process for Panels, Junction Boxes and Electronic Equipment. Cable Lacing Process. Electronic Component Assembly Wiring and Soldering Process. Shrinkable Sleeves Assembly Process. Etching, Cleaning, Preparation Processes. Roller and Immersion Tinning Process. Stripping Solutions for Electroplated and Surface Conversion Films. Section 5: Production Process of Solid Tantalum Capacitors. Monolithic Capacitor Process. Processing of Polyester Film Capacitors. Superconducting Tape Process. Flexible Printed Wiring Process. Flexible Magnet Process. Plated Wire Memories Process. Ferrite Magnet Process. Processing Wirewound Resistors.

It is extremely difficult to review this book and yet be fair to the author and the subject matter. The attitude to the book adopted by the reader will depend entirely on his standpoint. For example, an equipment manufacturer may feel that there is far too much emphasis on components and the component manufacturer will undoubtedly feel that his subject has not been dealt with fully. The treatment of the subject in various chapters is quite mixed; varying from much detail in some cases to outline treatment in others.

The electronics industry has for some time been in a changing situation with the introduction of integrated circuits. As the integrated circuit covers a larger and larger proportion of an equipment design, so the line drawn between the equipment manufacturer and the component manufacturer begins to get quite confused. To use the title 'Electronic Engineering Processes' in my view would have required more information on processes used in equipment manufacture well beyond the simple packaging and connecting up of solid-state devices and one or two other chosen components.

I applaud the constant reference to the need for well-controlled processes. Here I agree with the author that this is the only way to achieve a good quality product, but I feel that it would have been better to have laid down the principles on which control should be based rather than quote specific temperatures and other parameters which will obviously vary as the state of the art progresses.

The book would be a useful addition to the library of the newly qualified student about to enter the Electronics Industry to enable him to obtain a quick grasp of some of the processes. The index and bibliography are well organized and informative.

S. J. H. STEVENS

(Mr. Jowett is currently a consultant to several electronics companies and British editor of an international journal on microelectronics.)

Basic Electronic Instrument Handbook

Ed. Clyde F. Coombs, Jr. McGraw-Hill Inc., USA 1972. 22·7 cm × 14·5 cm. 836 pp. £12·85.*

Contents: Introduction to Instrumentation. Fundamentals of Electronic-measurement Instruments. Fundamentals of Signal-generation instruments. Using Electronic Instruments. Instrumentation Systems. Current-and-Voltage-measurement Devices. Circuit-Element Measuring Instruments. Signal-generation Instruments. Frequency- and Time-measurement Instruments. Recording Instruments. Special-function Instruments. Microwave Passive Devices.

The intended readership of this comprehensive handbook includes the considerable numbers of engineers, technicians and students who are not necessarily qualified in electronic and electrical engineering and who need to make measurements using electronic instruments in a wide range of science and technologies. It aims to give guidance on the choice of instruments, their limitations, proper modes of operation and the interpretation of results. The range of electronic measurements is so extensive that almost any electronic engineer will find the handbook a useful reference on measurements that are less familiar to him.

(Mr. Coombs is Manufacturing Manager of Hewlett-Packard Singapore Ltd; the 23 contributors are drawn from universities, government laboratories and electronics companies in the United States.)

Le Thyristor: Definitions, Protections, Commandes

G. Maggetto. Presses Universitaires de Bruxelles, Eyrolles Editeur, Paris. 1972. 24 cm × 16 cm. 264 pp. 450 F.B. (In French.)

Contents: Principles of operation. Definition of the characteristic parameters of thyristors. Protection of thyristors. Control of thyristors. Special thyristors, triacs, diodes. Standards, presentation of characteristics.

This is the first volume of a series entitled 'The Thyristor in Electrotechnology'. It is a detailed introduction to the principles and circuit performance of the device, the applications of which will follow in later volumes.

(M. Maggetto is in the Electrical Machinery and Industrial Electronics Laboratories of the Free University of Brussels.)

Video Recording. Record and Replay Systems.

Gordon White. Newnes-Butterworths, London, 1972. 21·5 cm × 13·5 cm. 208 pp. £3·25.*

Contents: Introduction to Video Systems. Television Waveforms and Colour Systems. Factors Affecting the Design of Video Tape Recorders. Theory of Magnetic Recording and Reproduction. Frequency Modulation in a Recording System. Broadcast Video Tape Recorder. Servo Systems of a Colour Recorder. Helical Scan Recorders. Editing, Professional Cartridge Systems and Duplication of Video Tape. Disc Recording. Wideband Analogue Recording. Manufacture and Care of Video Tape. Electronic Video Recording (EVR). R.C.A. Selecta Vision. Video System Comparisons.

The role of video recording in broadcasting is well established: in education, in industry and in the home it represents areas of enormous potential growth. The appearance of this book is therefore timely and, while the designer or intending designer of video recorders will probably require more detail, the approach adopted of providing the user with meaningful information on the various techniques and equipments makes this a useful guide and reference.

(Mr. White (Member 1968, Associate 1966) was for several years Video Engineering Manager with Amplex Ltd. and is now with Sesco Security.)

MOS Integrated Circuits

Prepared by the Engineering Staff of American Micro-systems, Inc. Edited by William M. Penney and Lillian Lau. Van Nostrand Reinhold Co., New York, 1972. 474 pp. 22·5 cm × 14·5 cm. £9·00.

Contents: MOS LSI technology. Basic theory and characteristics of the MOS device. MOS processing. MOS circuit design theory. Logic design with MOS. System design with MOS arrays. MOS memory products. Topology—array layout. Reliability of MOS devices and arrays.

The objective of this book is to present the techniques that are necessary to implement systems using metal oxide semi-conductor large-scale integration. It is primarily written for practising systems and logic designers in the digital equipment business, although circuit development engineers should find the circuit design data and topological considerations useful. The level of presentation assumes a general understanding of basic semiconductor theory and a familiarity with Boolean algebra as applied to digital systems design. Much of the material in this book has evolved from an engineering course in m.o.s. technology.

Book Supply Service

As a service to members, the Institution can supply copies of most of the books reviewed in the *Journal* at list price, plus a uniform charge of 25p to cover postage and packing.

Orders for these books, which are denoted by an asterisk (*) after the price, should be sent to the Publications Department at Bedford Square and must be accompanied by the appropriate remittance.

INSTITUTION OF ELECTRONIC AND RADIO ENGINEERS

Applicants for Election and Transfer

THE MEMBERSHIP COMMITTEE at its meetings on 20th March and 12th April 1973 recommended to the Council the election and transfer of 47 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: 20th March 1973 (Membership Approval List No. 156)

GREAT BRITAIN AND IRELAND

CORPORATE MEMBERS

Transfer from Graduate to Member

COLEMAN, Peter Leonard *Tunbridge Wells, Kent.*
COLES, David Charles *West Bergholt, Essex.*
DRUMMOND, Francis Gavin *Enfield, Middlesex.*
GLASBY, Peter Dawson *Crawley, Sussex.*
HOLDSTOCK, Eric John Paul *Rainham, Kent.*
KNIGHT, Marcus George *Chelmsford, Essex*
MACLEOD, Donald Finlay, Squadron Leader, RAF *Malvern, Worcestershire.*
MATTINSON, David *Newport, Isle of Wight*
POSTANCE, Richard, Flight Lieutenant, RA *Thame, Oxfordshire.*
TAYLER, Michael Ernest *Sandy, Bedfordshire.*

Direct Election to Member

BIRD, John Bolton, B.Sc., Squadron Leader, RAF *Elston, Newark, Nottinghamshire.*
BUTLER, Dennis Baden, B.Sc., Inst. Lt. Cdr., RN *Lee-on-Solent, Hampshire.*
CLAYDEN, Ronald Thomas *Twickenham, Middlesex.*
CROCK, John Richard *Chelmsford, Essex.*
JOHN, George Alan, Squadron Leader, RAF *Marlow, Buckinghamshire.*
McDONALD, Allan Kenneth *Glasgow, S.6.*
OAKES, Robert Allan, B.Sc. *Folkestone, Kent.*
THOMSON-JACOB, Ian *London, WC1.*

NON-CORPORATE MEMBERS

Transfer from Student to Graduate

HERRING, Malcolm, B.Sc. *Abingdon, Berkshire.*

Direct Election to Graduate

CHRISTIE, John Alexander, B.Sc. *Irvine, Ayrshire.*
FIELD, Richard John *Aylesbury, Buckinghamshire.*
FIELD-RICHARDS, Hugh Sherwood, B.Sc. *Ringwood, Hampshire*
HAWES, Anthony David McIntosh *Cambridge.*
WOOD, Andrew Richard, B.A. *Camberley, Surrey.*

Direct Election to Associate

REYES, James Louis *Merstham, Surrey.*
RICHARDSON, David Frank *London, W.13.*
SMITH, Duncan *Bearsden, Glasgow.*

STUDENTS REGISTERED

ALDOUS, Jack Maurice *Bracknell, Berkshire.*
BEJIDE, Olubunmi Emmanuel *London, N.5.*
BERRIMAN, Peter *Southampton.*
KNIGHT, Anthony John *Nottingham.*
SHIELD, Peter *Bath, Somerset.*
SIVANESAN, Appukuddy *Brighton, Sussex.*
VIRDI, Sukhdev Singh *Southall, Middlesex.*

OVERSEAS

CORPORATE MEMBERS

Transfer from Graduate to Member

DUTTA, Nirmal Kumar *Baroda Dist., India.*

Direct Election to Member

CAMPBELL, Hugh *Kuwait.*

NON-CORPORATE MEMBERS

Direct Election to Graduate

CHATTOPADHAYA, Arabinda *Thana, Bombay.*

Direct Election to Associate

ALAWIYE, Abdulfatai Ademola *Ibadan, Nigeria.*

STUDENTS REGISTERED

WU, Hon Fai *Kowloon, Hong Kong.*
YASSIN, Elkhider *Khartoum, Sudan.*

Meeting: 12th April 1973 (Membership Approval List No. 157)

GREAT BRITAIN AND IRELAND

CORPORATE MEMBERS

Transfer from Member to Fellow

GRIMM, Frank. *Cambridge.*

Transfer from Graduate to Member

ATTWOOD, Anthony John. *Hythe, Hampshire.*
BAMFORD, James. *Thurso, Caithness.*
BARHAM, John Joseph. *West Molesey, Surrey.*
COGSWELL, John Frederick, Flt. Lt., RAF. *Saltford, Bristol.*
EVANS, Arthur Gwyn. *Neath, Glamorgan.*
FREEMAN, Terence John, Flt. Lt., RAF. *Benson, Oxon.*
HOBBS, Robert Clive. *Swansea, Glamorgan.*
HURRICKS, Maurice Edward. *Gillingham, Kent.*
JONES, John Graham. *Prestwich, Manchester.*
MIDDLETON, Michael Robert. *Bishops Stortford, Hertfordshire.*
REID, John Michael. *Laughton, Sheffield.*
SCOTT, Thomas. *Barton-on-Sea, Hampshire.*
SMITH, John Derek. *Stivichall, Coventry.*
THOMSON, David. *Twyford, Berkshire.*
TRANTER, Edward Albert. *Dorchester, Dorset.*
WEEKS, Selwyn Howard. *Cwmbran, Monmouthshire.*
WILSON, Roy. *Cleethorpes, Lincolnshire.*
WOOD, Roger Jeremy. *Balsham, Cambridge.*
WOOD, Walter Alfred Spencer. *Abingdon, Berkshire.*
YOUSEFZADEH, Bijan, M.Tech. *Enfield, Middlesex.*

Direct Election to Member

LAWRENCE, Ronald Edward. *Aylesbury, Buckinghamshire.*
McKENZIE SMITH, Ian, B.Sc. *Milngavie, Glasgow.*
WEIGHTMAN, Bruce Arthur. *Leicester.*
WINDER, John Graham, B.Sc.(Eng). *Glenfield, Leicester.*

NON-CORPORATE MEMBERS

Transfer from Student to Graduate

HARDING, Denis Arthur. *Leigh on Sea, Essex.*
FARROW, Nicholas Charles. *Canterbury, Kent.*

Direct Election to Graduate

BAMPTON, Alvin David, B.Sc. *Eversley, Hampshire.*
JONES, Robert Arthur. *Hoveton St. John, Norfolk.*

STUDENTS REGISTERED

BARKER, David Vernon. *Chesterfield, Derbyshire.*
BAGSHAW, Philip Michael. *Leicester.*
BELL, Michael Geoffrey. *Chesterfield, Derbyshire.*
BURROWS, John Anthony. *Old Tupton, Chesterfield, Derbyshire.*
COATES, Keith Martin. *Bradford 3, Yorkshire.*
GOODWIN, Peter Geoffrey. *Stonebroom, Derbyshire.*
JHALERA, Dinkarrai Shantilal. *Wembley, Middlesex.*
LONGSTAFF, Ian. *Spennymoor, County Durham.*
LOWE, Kenneth Roy. *London E2.*
TODD, Michael Charles. *Ipswich, Suffolk.*

OVERSEAS

CORPORATE MEMBERS

Transfer from Graduate to Member

CHARLTON, Brian John, Flt. Lt., RAF. *B.F.P.O. 53.*

Direct Election to Member

MUKHERJI, Biswa Nath. *Ranchi 4, Bihar, India.*

STUDENTS REGISTERED

FONG, Shek Kwong. *Tsuen Wan, Hong Kong.*
GOH, Beng Koon. *Singapore 15.*
IBRAHIM, O. E. *Khartoum, Sudan.*
TAN, Gim Hong. *Singapore 2.*

Notice is hereby given that the elections and transfers shown on Lists 153 and 154 have now been confirmed by the Council.