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Various technical aspects of satellite communications and the vulnerability of satellites from the military point of view has led to a reassessment of HF and a renewal of interest in this portion of the radio spectrum. Both civilian and military requirements will be explored.

The state-of-the-art in microprocessors, synthesizers, modems and other equipment has led to the concept that HF communications can be adaptive using presently developed components. New aspects of system design and coding to meet the propagation problems will be explored.

This Lecture Series, sponsored by the Electromagnetic Wave Propagation Panel of AGARD, has been implemented by the Consultant and Exchange Programme of AGARD.

The material in this publication was assembled to support a Lecture Series under the sponsorship of the Electromagnetic Wave Propagation Panel and the Consultant and Exchange Programme of AGARD presented on 21—22 April 1986 in Brussels, Belgium, 24—25 April 1986 in Issy-les-Moulineaux, France and 28—29 April 1986 in Copenhagen, Denmark.

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# AGARD

ADVISORY GROUP FOR AEROSPACE RESEARCH & DEVELOPMENT, Paris.

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AGARD LECTURE SERIES No.145

# Propagation Impact on Modern HF Communications System Design.

NORTH ATLANTIC TREATY ORGANIZATION

DISTRIBUTION AND AVAILABILITY
ON BACK COVER

AGARD-LS-145

NORTH ATLANTIC TREATY ORGANIZATION

ADVISORY GROUP FOR AEROSPACE RESEARCH AND DEVELOPMENT

(ORGANISATION DU TRAITE DE L'ATLANTIQUE NORD)

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## THEME

Lecture Series 145 is concerned with high frequency communications and is sponsored by the Electromagnetic Wave Propagation Panel of AGARD and implemented by the Consultant and Exchange Programme.

The aim of these lectures will be to survey problems and progress in the field of HF Communications. The lectures will cover needs of both the civil and military communities for high frequency communications. It will discuss concepts of real time channel evaluation, system design, as well as advances in equipment, in propagation, and in coding and modulation techniques. The lectures are aimed to bring non-specialists in this field up to date so that HF communications can be considered as a viable technique at this time. The problems, difficulties and limitations of HF will also be outlined.

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## CONTENTS

# HIGH FREQUENCY COMMUNICATION: AN INTRODUCTION

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SUMMARY

The aim of this lecture series is to survey problems and progress in the field of HF communication. The region of the spectrum used, 3-30 MHz, allows observers of the RF signal energy to detect both ground wave and sky wave. It is a much used part of the spectrum but it is vital to be able to fully utilize its unique capabilities.

## 1.0 INTRODUCTION

It is easy to see why the disillusion with HF communications set in after World War II - at least with the planners and those interested in technological advances.

The propagation problems for example are numerous. (1) In the prediction area it is possible to develop long range models for forecasting monthly medians of frequencies to be used for a specific path in a particular month and phase of the sunspot cycle. From the viewpoint of the user the enormous range of values used to formulate that monthly median curve means that, for example, over a few hours communications are not assured. The techniques for developing long term models and those for forecasting short term variations are very different; they can be said to be different arts. (2) Several propagation problems are and will continue to be destructive. At high latitudes polar cap absorption, which at its worst can last for days, will wipe out any system of HF communications that relies on transmitting thru the polar cap ionosphere. Auroral fading and absorption are probably the most difficult problems of HF. Fast fading and selective path absorption wreaks havoc on HF signals. At equatorial latitudes fading produced by F layer irregularities is active in regions between plus and minus $20^{\circ}$ from the magnetic equator; it can be a devastating effect.

The 3-30 MHz region is only a small portion of the radio spectrum. At times and under certain propagation and operational conditions even 4 kHz seems to be a wide band. The ionospheric characteristics narrow the band used. The extensive use of HF further bounds the medium with signals piled on top of signals.

Finally there is the possibility of jamming with high power and highly directive signals blotting out the communications -- and there is the interception problem.

## 2.0 EVALUATING AND DEALING WITH CHANNEL PROBLEMS

Any new system will encounter many of these problems. The ideal new system tries to address each problem in a creative way. It deals with the auroral problems by knowing the geographical and signal characteristics of fading and absorption during periods of severe effects. The ideal new system provides for path diversity, rerouting messages to minimize the effect of auroral and polar cap absorption. The ideal system uses coding for corrections and changes transmission characteristics as a function of its analysis of real time channel problems.

The words holding together modern HF radio are Real Time Channel Evaluation; the system envisaged for the future for any use is adaptive. The adaptive aspects contemplated include some or all of the following:

a. Frequency band selection - what general frequency range should be tried for a particular path at a particular time.
b. Channel selector - precise frequency to be used after determining levels of interference, noise, occupancy.
c. Path selector - in the case of routing possibilities the real time channel evaluation would also determine the relay paths utilized.
d. Propagation mode selectivity is also of importance with ground and ionosphere paths available. Within the ionosphere there are many modes, i.e. single hop, multiple hop, sporadic E etc.
e. An area deemed of great importance for networks is to control the power level with excess power utilized only for reducing interference.
f. Antenna nulling is a more recent possibility for selective point to point communications.
g. Channel equalization.
h. Modulation selection, coding, error correcting and network protocal are methods where adaptive systems have and will change HF communications.

## 3.0 THE USERS

The largest consumers of HF systems at the present are governmental units ranging from National Police to postal and communications services within small nations. There are other groups using HF including data services and telephone systems. With regard to the latter remote areas in Canada utilize HF within their telephone system for communications. In many of these governmental uses there is difficulty in siting that preclude the use of VHF and UHF. The need in almost all these cases is a voice channel with the users frequently requesting secure communications.

In the area of civil aviation there are extensive plans for HF, the need pushed by the investment that would be required for an effective satellite system. There are firm requirements for high performance from an HF digital data link. Channels must be available under many conditions; message and symbol error states must be minimized for full utilization of the data capability. Civil aviation incidentally allows for a tuning of an adaptive system to conditions on individual routes.

The aim of these lectures is to tie together some of the points raised in this introduction. We hope to outline the problems of HF communications, the state of the art in coding, equipment, propagation forecasting. Finally we hope to discuss the research and development efforts necessary to fully exploit HF communications.

# MILITARY SYSTEM REQUIREMENTS

by

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For the military in many nations, if not in all nations, HF is a primary means of communications rather than a backup. In some countries no other communication techniques are contemplated for such services as ship to shore, aircraft to ground over distances beyond the line of site etc. The vulnerability of satellites makes nations consider HF communications as a primary military communications medium even when satellites are part of their military operations.

In some countries the moratorium on HF developments that lasted for more than a decade has been lifted with research and development focused on developing an HF radio system that is adaptive to the changing ionospheric medium, encompassing countermeasures to the electronic threat and performing the necessary signal processing for swift and reliable communications.

The requirements of a military system range from receptive broadcasts of simple commands on teletype to secure digital communications. Global operations demand the HF system adapt to propagation conditions in regions varying from equatorial to polar. Each area has its separate propagation problems and there may be a need to tailor systems to propagation requirements of the area.

In the architecture of military systems there are needs as follows:

(a) Connectiveness: the requirements may range from many users to many users to a requirement confining the operation to a limited group.

(b) Master station: In almost all cases a master station is contemplated with command headquarters organizing the distribution of the system. In this case multiple networks are needed to allow communications within limited groups.

(c) Relay: Relay may consist of rebroadcasting or rerouting. Path diversity, i.e. moving a signal to a path with minimum propagation outages is an area on which military research and development must concentrate.

(d) Identification: A requirement in military systems.

The technical requirements from users include many items. While the list may seem to be a wish list, firm needs can be shown in each of the following areas:

(a) Voice
(b) Data
(c) Adequate anti-jam margin
(d) Security
(e) Interoperability
(f) Coverage: ground and skywave
(g) Channel optimization
(h) Message error control

# INTRODUCTION: CIVILIAN AND DIPLOMATIC REQUIREMENTS

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## ABSTRACT

HF radio is used extensively to meet civilian and diplomatic communication requirements. Briefly described are the non-military requirements for HF communication by exploration companies, aeronautical and marine operations, the diplomatic corp, radio amateurs, as well as by users in remote and isolated areas where other means of communications are not available.

## TYPES OF SERVICES

HF radio provides the following types of services:

### Fixed Service

A service of HF radio communications between specified fixed points or locations.

External Affairs or the diplomatic corp are major users of this service. An HF service is provided between embassies and missions in various countries as well as directly to the home country. The radio facility is usually located in the embassy or mission building with the antennas mounted on the roof. The physical environment restricts types of antennas that can be used on the building and the high RF noise levels in urban areas can seriously degrade the HF service. Antennas should be of a low profile to avoid destruction in case of internal conflict in the country.

Telephone and private companies are also major users of the fixed service. The HF service is provided to isolated or remote areas where more reliable services are not presently available. Generally the traffic carried is low capacity and limited resources are provided by the user to improve the system. In many cases small companies use HF to avoid cost of long distance telephone charges of the carrier networks. In many mining and exploration operations, HF radio is used in the initial stages of development of their operation because of the length of time it takes to establish a regular telephone service using microwave relays or land lines.

### Mobile Services

A service of HF communications between mobile and land stations or between mobile stations such as aircraft or ship.

The bands available for these services are very congested and the performance of the systems vary depending on the priority the user places on the need for reliable communications. Aeronautical mobile services are required by airlines providing regional, national and international flights. For small airline companies the radio equipment is inexpensive and performance rather marginal. For airlines with national and international services the aeronautical land based stations and equipment is of high quality and the performance is satisfactory to the user. However, the HF service is not generally used when other more reliable radio services are available in the area. Many areas of the world have limited radio services outside the use of the HF band. The mobile service also provides communication between coast stations and ships or between ships. The ground wave mode of propagation enables communication over large distances along coastal regions as compared to the use of the VHF frequencies. The range for skywave communications varies from a few miles to distances half way around the world. Depending on the size of the ship and its operation, there is often a requirement for highly reliable communications back to home base.

### Amateur Service

A service of self training, intercommunication and technical investigations carried on by amateurs, that is, by duly authorized persons interested in radio techniques solely with a personal aim and without pecuniary interest.

The portions of the HF band reserved for amateur usage are always congested. Because of the large market for equipment by this group, new techniques are often tested on and by this group and if successful are later incorporated into commercial equipment.

# THE DESIGN OF STATIC & MOBILE HF COMMUNICATION SYSTEMS

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SUMMARY

The lecture deals with more specific aspects of HF communication system design arising from a previous lecture entitled "HF system design principles" presented as part of AGARD Lecture Series 127 in 1983.

The major topics considered are

(i) the design of point-to-point systems;
(ii) the design of systems involving mobile terminals;
(iii) system control and frequency management;
(iv) resistance to interception and disruption.

The lecture is intended to povide a technical framework for the lectures covering the more detailed aspects of HF communication system design.

## 1. INTRODUCTION

In this lecture, the design of static and mobile HF communication systems will be considered from the viewpoints of present practice and future trends. The lecture is intended

(i) to introduce the problems associated with the design of static and mobile HF communication systems and their control and operation;

(ii) to identify the techniques offering greatest promise of improved performance in future designs;

(iii) to provide a general framework for the more detailed aspects of HF communication to be discussed in later lectures.

Before the HF propagation medium can be used efficiently for communications purposes, its strengths and weaknesses must first be appreciated; the most important of these are listed below (Darnell, 1983a)

### 1.1 Strengths of the HF Medium

(a) The ionosphere is a robust propagation medium which recovers rapidly after major perturbations, eg polar cap events (PCE's), sudden ionospheric disturbances (SID's) and high altitude nuclear bursts.

(b) Long-term (monthly mean) propagation parameters are predictable with reasonable accuracy.

(c) Each communication link exhibits unique characteristics, eg fade rates and depths, multipath structure, noise levels, etc, which potentially can be used to isolate that path from the effects of other transmissions in the HF band.

(d) Only simple equipment and operating procedures are necessary to achieve access to the medium; this allows it to be exploited by simple mobile terminals.

(e) Equipment costs are low in comparison with those of other types of long-range communications systems.

### 1.2 Weaknesses of the HF Medium

(a) The ionosphere is subject to sudden unpredictable disturbances such as the PCE's and SID's mentioned above.

(b) Although the long-term parameters of the propagation path are relatively predictable, significant departures from such predictions can be expected in the short-term (day-to-day).

(c) Levels of manmade interference are high, particularly at night, and the nature of such interference is inadequately characterised.

(d) A high level of system availability and reliability requires considerable user expertise in manually-controlled systems.

(e) The available capacity of a nominal 3 kHz HF channel is limited to a maximum of a few kbits/s; data rates of, at most, a few hundreds of bits/s are more realistic if high levels of availability and reliability are necessary.

As will be shown in Sections 4 and 5, care must be taken to employ HF systems operationally for the types of traffic and service to which they are well matched, and not to impose "unnatural" requirements for which the medium is fundamentally unsuitable.

### 1.3 Mobile Problems

In addition to the inherent natural weaknesses noted above, systems incorporating mobile terminals suffer from other fundamental limitations. When contrasted with static (land-based) terminals, mobile installations have the following disadvantages (Darnell, 1985):

(a) Comparatively low transmitter powers, limited by the generation capacity of the mobile platform.

(b) Simple, narrowband, and relatively inefficient transmitting antennas, with minimal directivity.

(c) Electromagnetic compatibility (EMC) problems due to the HF communications equipment being co-sited with multiple EM systems such as radars and communications transmitters using other frequency bands.

(d) Dependance upon a propagation path which, in addition to natural temporal variability, is also affected by variations in mobile position.

Land based stations, on the other hand, can use high powers and efficient and directive antennas, coupled with electrically quiet receiving locations. It is also relatively straightforward to provide interconnections between sites to give a capability for networking.

In military applications, mobile terminals may represent a particularly vital element in the operational strategy: as a consequence there may be additional requirements placed on the mobile to function in such a way that the probability of its transmissions being decoded, intercepted, located or jammed is minimised.

Mobile terminals can thus be considered as a class of "disadvantaged" users of the HF spectrum, who may nevertheless be required to communicate reliability and securely over a wide geographical region, possibly via an ionosphere exhibiting severe irregularities.

### 1.4 Design Philosophy

One of the main problems associated with the operation of current HF systems is that they are employed primarily to pass traffic at a constant rate - even up to 2.4 kbits/s and above - in the same way as with line circuits or other less dispersive media. However, as will be shown in Section 2, the capacity of an HF skywave link is constantly varying over a wide range which ideally requires adaptation of the signal parameters and/or signal processing procedures in accordance with the available capacity at any time. The design concepts developed in this lecture attempt in some circumstances to mitigate, and in others to exploit, the natural variability of HF paths in various ways.

For both static and mobile systems, the basic requirement is for more effective and responsive control and adaptation procedures to be incorporated into the system design, rather than the relatively inflexible frequency utilisation procedures and transmission formats employed by many HF systems at present. An essential pre-requisite for this improved control and adaptation is the availability of an appropriate channel model obtained by real-time channel evaluation (RTCE), which is the subject of a companion lecture (Darnell, 1986) and a lecture in the previous HF lecture series (Darnell, 1983b).

Two communications scenarios which are encountered in practice are illustrated in Figs. 1(a) and 1(b). Fig. 1(a) shows an "open-loop" situation in which there is no feedback path between receiver and transmitter, whilst Fig. 1(b) shows a "closed-loop" situation in which such a path does exist. Ideally, the aim of the control procedure should be to made the transmitted signal, x(t), and the received signal, y(t), identical. In both cases, before any form of optimal or sub-optimal control can be applied to achieve the desired purpose of the system, it is necessary to characterise the propagation path in order to produce an appropriate model for use in the control algorithm, ie the path must be "identified".

For the open-loop system this can be achieved in three ways:

(a) By using a priori knowledge of the nature of the path, eg from previous practical experience and via off-line propagation analysis programs.

(b) By multiplexing RTCE signals with the traffic signals so that the receiver can derive information to model the path and hence can adjust its parameters accordingly.

(c) By deriving RTCE information at the receiver from the "natural" operating signals of the system.

In the closed-loop system, the feedback link enables information extracted from the received signal(s) to be used to characterise the path and then to be passed back to the transmitter to allow its parameters to be varied adaptively.

The design of many HF systems incorporating mobile terminals takes only limited account of the specific problems mentioned in Section 1.3. In Section 5 of this paper, the manner in which these problems can be minimised in mobile systems is discussed. The design philosophy exploits the strengths of fixed terminals and minimises the effects of any problems associated with mobile terminals, at the same time placing as few constraints as possible upon mobile operational flexibility. In essence, the complexity of the design resides in the fixed terminals of the system, with the mobile equipment being kept simple.

## 2. OPERATIONAL ENVIRONMENT & REQUIREMENTS

Currently, there seem to be some misconceptions about the nature and potential of the HF band as a vehicle for communication. These misconceptions arise from the following factors:

(a) A tendency to consider the problem of reliable HF transmission as one of exploiting Rayleigh-fading ionospheric skywave and/or steady-signal surface wave paths, with a background of Gaussian white noise (GWN). In practice, the propagation mechanisms may be more complex and the noise background is certainly non-Gaussian in many instances.

(b) An assumption that constant transmission rates are appropriate for HF systems, whereas the available information transmission capacity of any given channel may vary over a wide range in a short time interval.

(c) An assumption that HF communication can replace, or back-up, other types of less dispersive communication media and still maintain a similar level of reliability. In fact, there are classes of traffic, which can be carried by say a 3 kHz bandwidth telephone channel, for which a similar bandwidth HF link is fundamentally unsuitable.

(d) An incomplete understanding of the medium in terms of the strengths and weaknesses listed in Sections 1.1 and 1.2.

In the following discussion, it will be assumed that conventional frequency assignment and bandwidth limitations will apply in any HF system design, ie a typical communication system will have a limited set of assigned channels within which to operate, with each channel having a nominal maximum bandwidth of 3 kHz. Additionally, it will be assumed that there are no geographical limitations upon the operational deployment of the systems.

### 2.1 Propagation, Noise & Interference Models

In the design of HF communication systems, it is convenient to use off-line propagation analysis programs, eg (Barghausen et al, 1969) and (Lloyd et al, 1981), to dimension the system in terms of required transmitter powers, antenna gains, etc. There are several basic problems associated with the use of these programs:

(a) They are derived from limited data bases.

(b) They can only treat effects such as sporadic E-layer propagation in a rather simplistic manner.

(c) The modelling of noise and co-channel interference effects tends to be rudimentary, and can lead to inappropriate system design criteria (Darnell, 1984).

(d) Their outputs are typically in the form of relatively long-term, eg monthly mean, predictions of path parameters at hourly intervals.

Even allowing for the shortcomings listed above, off-line analysis can yield reasonable estimates of required transmitter powers and antenna characteristics. However, it does not provide information on the short-term, ie minute-to-minute or second-to-second, variability of the propagation medium which would enable more detailed consideration to be given to the form of signal generation and processing procedures appropriate for a particular path.

To some extent, channel simulators overcome this deficiency by making use of a deterministic model of HF propagation [see for example (Watterson, 1975) which may, or may not, be representative of the conditions on a given path at a given time]. Simulators still suffer from the lack of an adequate model of co-channel interfering signals - an important factor in determining the reliability of many communication systems.

In order to assess the types of transmission formats and services for which the ionospheric medium is appropriate, it is first helpful to describe the propagation and interference environment in a way which is meaningful to a communication system designer. An information-theoretic model of propagation is postulated in (Darnell, 1982a); this describes ionospheric skywave propagation in terms of 3-dimensional "windows" with dimensions of frequency, time and distance, as shown in Fig. 2. The sides of the window cover the ranges

$$\left.\begin{array}{lcl} f_1 & + & \Delta f \\ t_1 & + & \Delta t \\ \text{and} \quad d_1 & + & \Delta d \end{array}\right\} \qquad [1]$$

Within a given window, a specified signal-to-noise power ratio (SNR) will be maintained for the ranges of f, t and d shown. In general, the lower the required SNR, the greater will be the dimensions of the window; thus in Fig. 2, window 2 corresponds to a higher SNR than does window 1. For a given communication system, the "volume" of an individual window will depend upon the nature of the signal processing procedures used at the receiver, eg modulation and coding, since these affect the required SNR at the receiver; antenna characteristics, diversity processing, diurnal and seasonal variability, etc, will also cause the dimensions to change.

A systematic study of the nature of co-channel interference in the HF band has been carried out by (Gott et al, 1983). In the same way as propagation windows can exist between communications transmitter and receiver, similar windows may also occur between interfering transmitters and the communications receiver; any overlap between communications and interference windows will tend to reduce the effective volume of the former. Fig. 3 illustrates this situation.

It is instructive to consider what a window model of ionospheric propagation might look like for sporadic E-layer and meteor-burst modes. When sporadic E is present, it can give rise to stable, single-mode propagation: however, a given region of ionisation tends to be usable only over a somewhat restricted geographical area. Thus, it might be expected that windows due to this mechanism would be relatively large in the f and t dimensions, but small in the d dimension. Because of the physical nature of meteor-burst communication paths, ie short-duration and wideband, which are virtually unique for given transmitter and receiver locations, but relatively large along the f dimension. Fig. 4 shows typical windows for various ionospheric modes: although, for simplicity, regularly-shaped windows are shown, in practice they would tend to be irregular volumes.

## 2.2 Transmission Characteristics arising from the Window Model

From the discussion in the previous section, it can be seen that the lower the SNR required at the receiver, the greater will be the volume of any transmission windows available. Therefore, if a transmission wndow, associated with a signal-to-noise ratio $SNR_1$, has maximum dimensions $\Delta f_1$, $\Delta t_1$ and $\Delta d_1$, any window capable of supporting a ratio $SNR_2$ where

$$SNR_2 > SNR_1 \qquad [2]$$

and lying in the same ranges of f, t and d will normally have maximum dimensions $\Delta f_2$, $\Delta t_2$ and $\Delta d_2$ given by:

$$\left.\begin{array}{lcl} \Delta f_2 & < & \Delta f_1 \\ \Delta t_2 & < & \Delta t_1 \\ \Delta d_2 & < & \Delta d_1 \end{array}\right\} \qquad [3]$$

This implies that the $SNR_2$ volume is less than the $SNR_1$ volume, as indicated in Fig. 2

Using the Shannon upper bound for error-free channel capacity C (bits/s), ie

$$C = \Delta f \log_2 [1 + SNR] \text{ bits/s} \qquad [4]$$

where $\Delta f$, the transmission bandwidth, and distance are assumed fixed, it is possible to view the variation of SNR with time as a variation in channel capacity, as shown in Fig. 5.

This variation in channel capacity is due to changes in the window structure with time and might involve a number of physical factors, eg one mode becoming dominant in a multipath situation or a fading co-channel interfering signal. These effects are illustrated in Fig. 6. In the region $\Delta t$, one of the wanted propagation modes is dominant with respect to the other mode and the interfering signal, thus giving rise to a high effective SNR over that interval. On a given ionospheric link, $\Delta t$ values may range from milliseconds to hours, dependent upon the required value of SNR at the receiver.

Many existing transmission schemes employ techniques which, to some extent, are capable of smoothing out SNR variations, eg diversity processing and adaptive equalisation. However, as is explained in (Darnell 1984), these techniques cannot overcome capacity changes due to the presence of co-channel interfering signals. Therefore, the basic problem that the communication system designer has to solve is one of making effective use of a propagation medium with a time-varying information transmission capacity.

From the preceding discussion, it is evident that systems operating at constant transmission rates will, in many cases, be fundamentally mismatched to the variable capacity medium. In Fig. 5, a system operating at a constant rate R would normally have

$$R < C_{min} \qquad [5]$$

for error-free reception over a given time interval - which represents an inefficient utilisation of available capacity for most of the transmission period. From this simplified consideration of ionospheric path characteristics, it would seem that a more natural mode of transmission would involve a variable rate, selected according to the prevailing path conditions. Using the model of propagation and interference shown in Fig. 6, even in the presence of extremely high levels of interference there will be occasional windows when the unwanted signal is in a deep fade and one of the wanted modes is dominant; this may be the case even if the average SNR is very low. If these windows can be identified, it will be possible to maintain some residual channel capacity, even in the face of extremely high-level co-channel signals. It should be stressed, however, that in severe interference these windows may only persist for fractions of a second at a time.

The similarity between the above model of normal ionospheric skywave propagation with high levels of interference and the characteristics of meteor-burst modes (Bartholomé & Vogt, 1968) is obvious; both can be considered as intermittent paths. The situation with normal skywave propagation and relatively low levels of co-channel interference is rather more complex; instead of the path being completely intermittent, it gives rise to transmission windows of variable capacity, possibly with some low-level of capacity being available on a continuous basis.

In addition to the normal propagation and interference environment discussed in the previous two sections, HF systems may be required to operate under extreme conditions of poor signal propagation or high manmade interference levels. These situations are examined in the following two sections.

### 2.3 High Latitude Effects

Propagation mechanisms can be very irregular in polar regions, with the major effects being (Thrane, 1983):

(a) polar cap absorption (PCA) events due to the influx of high energy protons and alpha particles penetrating to the lower ionosphere: such events may persist for several days, causing severe attenuation of skywave signals and an effective communication "blackout" over a wide geographical area;

(b) auroral absorption, caused by high energy electrons penetrating the ionospheric D-region: this source of signal attenuation typically persists for a few hours only and affects a more restricted geographical area than does a PCA;

(c) irregular behaviour of sporadic E-layer, spread F-layer and auroral oval propagation modes, giving rise to abnormal and variable multipath time spreads;

(d) abnormal doppler shifts and phase instability of received signals due to the variable nature of the high-latitude ionosphere, manifesting itself in the form of very rapid signal fading;

(e) large variability of maximum usable frequencies (MUF's) from very low to abnormally high values: the former may result in spectral congestion and hence high levels of co-channel interference, whilst the latter may allow HF type propagation to take place in an extended frequency range above 30 MHz.

Clearly, the propagation environment in polar regions is relatively unstable and unpredictable, thus limiting the effectiveness of off-line propagation analysis programs. Also, the degree of adaptation required of systems operating in these regions is likely to be greater than for those deployed in temperate latitudes.

### 2.4 Interception and Disruption

In Fig. 3, the interference window could arise from the activities of a jammer attempting to disrupt the operation of the communication system. Similarly, the paths from communications transmitter to interception receivers will also have a characteristic window structure.

The problem for the designer and operator of the cummunication system is thus to minimise the probability of coincidence between interfering/jamming/interception windows and those for the communication system. If partial or complete coincidence does occur, there is an opportunity for disruption or interception of the communications traffic. The design strategy in this environment should therefore be to

(i) Maximise the size of the communications windows, and

(ii) Minimise the coincidence probability between communications and interference/jamming/interception windows by improved system control.

In general, it can be assumed that a potential jammer will have a significant radiated power advantage relative to the communicator; also, it is probable that such a jammer will have available directive and steerable antennas, and be capable of operating in both narrowband and wideband modes. Thus, in Fig. 3, the window dimensions for the jammer will on average be rather larger than those for the communications system. It is instructive to consider the fundamental limitations on ECM effectiveness and then to identify those techniques which can exploit those limitations, together with the strengths of the medium listed in Section 1.1, to the benefit of the communicator. Attention should not be confined solely to the conventional HF band; the possibilities offered by the use of higher frequencies should also be considered.

## 2.5 Fundamental Limitations on ECM Effectiveness (Darnell, 1982b)

As illustrated in the simple models discussed in Section 2.1, each HF path is, to some extent, unique in its characteristics. The probability of complete overlap of communications and ECM windows can therefore, in principle, be made low.

### 2.5.1 Propagation of Analysis Time Limitations

Consider the arrangement illustrated in Fig. 7: the network shown comprises a communications transmitter and receiver, together with a jamming and interception site. The propagation times in the network are:

$$\left.\begin{array}{lll} \text{(i)} & \text{Communications Transmitter} \longrightarrow \text{Communications Receiver} & = \tau_{TR} \\ \text{(ii)} & \text{Communications Transmitter} \longrightarrow \text{Jamming and Interception Site} & = \tau_{TJ} \\ \text{(iii)} & \text{Jamming and Interception Site} \longrightarrow \text{Communications Receiver} & = \tau_{JR} \end{array}\right\} \quad [6]$$

Assuming that the jammer cannot react to changes in communications system parameters before its associated interception site has monitored these changes, the fundamental propagation time advantage $\tau_A$ for the communicator over the jammer is:

$$\tau_A = [\tau_{TJ} + \tau_{JR}] - \tau_{TR} \qquad [7]$$

In addition, however, the interceptor will require a further time, $\tau_{AN}$, in order to analyse changes in the nature of the communications traffic before he can command the jammer appropriately. Thus, the overall time advantage now becomes:

$$\tau_A = [\tau_{TJ} + \tau_{JR} + \tau_{AN}] - \tau_{TR} \qquad [8]$$

This represents a fundamental limitation on jammer effectiveness which could only be overcome to some extent, and at inadmissible cost, by continuous, wideband, "blanket" jamming.

Practically, it is difficult for co-located interception facilities to look through a high power jammer, and therefore the intercept location may well be separated physically from the jammer. In this case, an additional propagation time between interceptor and jammer must be included in expression [8].

Typically, with modern intercept and jamming systems $\tau_A$ could be expected to have a minimum value of a few tens of milliseconds - the majority of that time being taken up by signal analysis. Two approaches suggest themselves as being able to exploit this time limitation on ECM effectiveness:

(i) The use of a communications system employing rapid frequency agility under cryptographic control such that the time spent on each frequency is $\tau_A$. This would inevitably require high signal levels which would be relatively simple to monitor and track.

(ii) The use of a transmission format which has no unique characteristics and is thus difficult to distinguish from other HF traffic in which the jammer has no interest. In this case, the "sorting" problem faced by the interceptor is increased with a consequent lengthening of analysis time, $\tau_{AN}$.

### 2.5.2 Corrupted Reference Limitations

A further advantage held by the communicator over the jammer/interceptor is the fact that the communications receiver will always be in the position of knowing the exact range of possibilities for the transmitted signal prior to reception. Therefore, the actual received signal can be compared with a set of perfect reference signals in a maximum likelihood detection scheme. The interception site, on the other hand, should at best only have available a corrupted estimate of the set of possible transmitted signals. Potentially, this means that the communications system has at least a 3dB advantage over the interceptor (eg the difference between coherent and differentially-coherent phase-shift keying). This advantage can be further enhanced by means of techniques such as correlation reception, quenched filtering, accurately-controlled tone filtering, time adaptation, frequency agility, encryption, signal format variation, etc, which will be discussed further in the later sections.

2.5.3 Data Base Limitations

Related to the analysis time and noisy reference limitations identified above is a more general limitation of the data base available to the interceptor and jammer. An effective jamming control strategy requires detailed knowledge of communication network configuration and of which are the key links. Propagation and analysis time limitations may well mean that it will take an ECM controller a considerable period to build-up such a data base. A communicator can accentuate this problem by making his communication system time-variable to the maximum extent, eg by altering frequency plans, traffic flows, network topology, cryptographic procedures and signal formats.

Direction finding (DF) can, and will, be employed by the interceptor as a vital sorting parameter: in many cases, it may well be the only available means of signal propagating to the interception site and short transmission durations can be used to increase the errors associated with DF.

As a general principle, the amount of "real" traffic passed by a communication system should be minimised. This then affords the option for the system transmission rates to be contracted by the use of more powerful source and channel coding techniques in the face of an ECM threat, whilst still retaining sufficient capacity to pass essential information. Hence, the overall system should not be organised in such a way that the peacetime traffic levels must be maintained in order to ensure its effective operation in time of stress.

In the following three sections, the effect of the factors discussed in Sections 1 and 2 will be considered in the context of a generalised HF system design methodology.

## 3. SYSTEM DESIGN: CURRENT STATUS

In a generalised HF communication system incorporating both static and/or mobile terminals, there may be a requirement for any of the following:

(a) static terminal-to-static terminal communication;

(b) static terminal-to-mobile terminal communication;

(c) mobile terminal-to-static terminal communication;

(d) mobile terminal-to-mobile terminal communication.

The manner in which each of the links above is currently implemented will now be outlined briefly.

### 3.1 Static Terminal-to-Static Terminal

Because relatively high transmitter powers and directive antennas can be installed at fixed sites, the engineering of a static terminal-to-static terminal link can normally be accomplished to give a reasonable communications reliability - at least for simple forms of traffic, eg low-speed telegraphy data and analogue speech. However, digital data, or digital speech, at say 2.4 kbits/s will present a greater problem, and circuit reliabilities for this form of traffic will be relatively low.

If a static link operates with a defined frequency complement on a regular schedule, there will be a significant frequency "airing" effect which will tend to deter other users of the spectrum from encroaching upon those frequencies, thus minimising the levels of co-channel interference.

### 3.2 Static Terminal-to-Mobile Terminal

Static terminal-to-mobile terminal communication is frequently implemented via a broadcast-type system; this again allows high transmitter powers and directive antennas to be used at a fixed site in order to radiate signals simultaneously to multiple mobile terminals, possibly situated anywhere within a wide geographical region.

A basic broadcast system operates on the principle of long-term frequency selection diversity; Fig. 8 illustrates that principle. It is seen that at any time of day, the same traffic signal will be radiated on a number of distinct frequencies or components, with the assumption being that at least one of the components will propagate effectively to any given mobile. The frequencies must therefore be chosen to provide a compromise coverage over the complete mobile operational area for both day and night conditions.

Typically, each component will comprise multiple sub-channels arranged within a nominal channel bandwidth, as shown in Fig. 9. The sub-channels correspond to traffic signals originating from various message sources; a given mobile will only extract those sub-channels of interest to it from the multiple sub-channel raster.

If all sub-channels are not assigned, it is possible to transmit the same information on more than one sub-channel to provide in-band frequency diversity. In Fig. 9, the $i^{th}$ sub-channel is employed for engineering order wire (EOW), or system control, purposes.

The system thus has the potential for frequency diversity at two levels: a given mobile selects the best component in terms of received SNR and, within that component, may also be able to apply in-band frequency diversity combining.

It is also possible for the same broadcast traffic to be radiated simultaneously from different static sites, giving path diversity in addition to frequency diversity.

### 3.3 Mobile Terminal-to-Static Terminal

The mobile terminal-to-static terminal situation is, in many respects, the complement of that discussed in Section 3.2. Two basic options exist as indicated in Fig. 10: the first where the mobile must access a single static terminal; the second where a number of alternative static terminal accesses are available. For the situation shown in Fig. 10(a), there may also be occasions when a bi-directional, full duplex link must be engineered between the two terminals. This can be difficult to maintain because of changes in mobile antenna characteristics as it changes position and orientation, coupled with the need for co-ordinated frequency changes.

In Fig. 10(b), the mobile has a choice of static terminals with which to communicate. If the static terminals are widely separated, there will be a high probability that both propagation and co-channel interference effects will be largely decorrelated at the various locations. Thus, it can be expected that at least one of the static terminals will be able to receive the mobile traffic in a reasonably optimum manner over the complete ranges of time and position; this traffic can then be passed to the desired recipient by HF or non-HF communications links from the static terminals.

### 3.4 Mobile Terminal-to-Mobile Terminal

Fig. 11 shows the two basic methods by which a mobile terminal can communicate with another mobile terminal. In Fig. 11(a), the mobiles communicate directly via skywave or surface wave HF propagation. In the latter case, vertically-polarised antennas mounted on ships allow links to be established over distances up to a few hundreds of kilometres. Skywave links over comparable ranges necessitate high-angle propagation via horizontally-polarised antennas, which may be difficult to mount on ships. Both terminals suffer from the fundamental limitations listed in Section 1.3, and thus direct mobile links are difficult to operate reliably over a wide range of distances.

Fig. 11(b) illustrates an alternative arrangement in which traffic is routed from one mobile to another via a static "relay" terminal. In this case, the advantages of the static station can be exploited in its relay role and communication between mobiles can be maintained over a wider range of distances with greater reliability - at the expense of more sophisticated control procedures.

As stated previously, HF links of various types are normally controlled by using data obtained from off-line propagation analysis programs. More recently, various forms of real-time channel evaluation (RTCE) are being used to aid frequency selection and system control.

Sections 4 and 5 respectively discuss design trends in HF communication systems

(a) involving static terminals alone;

(b) in which mobile terminals are present, possibly in addition to static terminals.

Since (b) presents a more significant and important practical problem, attention will be concentrated in this area. Prior to this discussion, however, the topic of diversity processing will be reviewed briefly because potentially it can provide a substantial improvement in performance for all types of HF communication systems at reasonable cost.

### 3.5 The Value of Diversity Processing in HF Communications

It has long been recognised that diversity processing of various types can be applied with advantage to HF communications. The most widely-used types are:

(a) spaced receiving antenna diversity;

(b) diversity obtained from receiving antennas of different polarisations;

(c) frequency diversity, either within a given assigned channel or over different assigned channels;

(d) time diversity, in which the same signal is repeated at different time offsets;

(e) diversity arising from the availability of independent multipath components.

All forms of diversity combining require there to be available two or more independently-derived received versions of the wanted signal. Under these circumstances, fades or interference affecting one version may be decorrelated with similar mechanisms affecting the other version(s), thus allowing a better overall signal estimate to be derived by using information from all versions (Stein & Jones, 1967).

Fig. 12 shows the reduction in received SNR admissible, under Gaussian White Noise (GWN) and Rayleigh fading conditions, when 2nd and 4th-order selection diversity combining is used rather than non-diversity reception. It is seen that for dual-diversity combining, an SNR reduction of about 10dB can be expected for an error probability of 1 in a 1000 - equivalent to a ten-fold increase in transmitter power, or the difference between an omnidirectional and a highly directional receiving antenna.

### 3.6 Speech Transmission at HF

One of the major technical problems associated with HF communication, which has yet to be solved satisfactorily, is that of transmitting digitised, and hence secure, speech reliably over such a time-variable channel. Existing vocoders typically operate at data rates of between 1.2 and 2.4 kbits/s: at such rates, reliable HF transmission cannot be guaranteed and, indeed, in spectrally congested environments such as the central European region, it is improbable that a satisfactory grade of service can ever be achieved - particularly at night.

Two options present themselves: first, users must accept that secure/digitised speech cannot be transmitted reliably over many types of HF channel - a situation which is operationally unacceptable; secondly, more sophisticated speech source encoding and system management schemes must be developed to give an average transmission rate not exceeding a few hundreds of bits/s.

One such system, involving a combination of speech-recognition and synthesis, coupled with user interaction with visual or audio prompts, has been demonstrated practically (Chesmore and Darnell, 1985). The information rate for this syyystem is approximately 25 bits/s, making it suitable for use with simple, low-power, mobile terminals.

## 4. DESIGN TRENDS IN STATIC SYSTEMS

To date, as indicated in Section 3.1, static HF systems have tended to exploit high transmitter powers and directive antennas for both transmission and reception purposes. However, the majority of developments in land-based static networks are now directed towards the design of systems with powers ranging from a few hundreds of W to a few kW; also, since such systems are often required to be redeployable, the antennas used must be of types which can be dismantled and re-erected rapidly. As a consequence, more emphasis has to be placed upon system control and channel coding/decoding procedures in order to achieve the desired level of performance (Darnell, 1983a).

### 4.1 Real-Time Channel Evaluation

Referring to the window model of HF propagation introduced in Section 2.1: for a static system, the position of the terminals is determined by operational considerations; thus, the problem of system control reduces to one of selecting values of frequency and transmission time for the given path. This selection procedure can be implemented by employing some form of RTCE. RTCE procedures have been developed for both the open loop and closed loop scenarios introduced in Section 1.4. At present, however, RTCE techniques are only designed to monitor and select the best of a set of alternative transmission channels at a specified time; in general, they are not well matched to the task of monitoring the short-term time variability of those channels. Thus, in the context of the simplified HF channel model postulated in Section 2.1, current RTCE algorithms search for transmission windows which are likely to persist for the complete duration of a fixed constant rate transmission; at the same time, they attempt to minimise the long-term overlap between communication and noise/interference windows.

In future, since the utilisation of the HF medium will remain at a relatively high level, or even increase, more emphasis should be given to the development of RTCE systems which enable the available capacity of the medium to be used more effectively by monitoring the shorter-term time fluctuations of a set of assigned HF channnels. This RTCE data could then be used in the control of an adaptive transmission rate system in which the transmission rate at any time could be matched more accurately to the available capacity. To implement this form of channel monitoring economically, it would be necessary to time/frequency multiplex RTCE probing signals and measurement periods with the traffic signals, possibly with the input data stream being buffered during the probing/measurement intervals.

One of the main disadvantages of many RTCE systems as currently implemented is that they are separate from the communication systems which they are designed to support; they require dedicated and expensive units which are comparable in cost to the units of the communication system itself. The future trend should be for RTCE to be integrated (or embedded) into the communication system design and hence to employ the same basic RF and signal processing equipment (Darnell, 1986). Also RTCE in the form of ionospheric sounding, in which energy is radiated across a major portion of the HF spectrum, should be restricted to the purposes of ionospheric research and not employed to support communication systems. Rather, integrated communication and RTCE systems operating only in assigned channels should be developed. In this way, the spectral pollution associated with dedicated RTCE systems would be minimised.

## 4.2 System Control

It should again be emphasised that, in the system control context, RTCE represents the process of identification, or modelling, which must take place before optimal control can be applied. Clearly, the path parameters are time varying and thus any control algorithm needs to be adaptive in response to such changes. If manual control procedures are used, response times will be at best a few tens of seconds and there will be no chance of the system being able to utilise relatively short duration, high capacity transmission windows (analagous to the meteor burst situation). RTCE, however, provides the essential information to enable the system control procedures to be automated - hence potentially providing a greatly improved response time.

The majority of HF systems operate in a closed loop, or 2-way mode. In addition to RTCE, the control of this type of system requires the ability to be able to transfer information concerning reception conditions between receiver and transmitter sites, ie the provision of a high integrity engineering order wire (EOW) facility. It is a fundamental requirement of such an EOW that it should be capable of passing the essential control data reliably for a short interval after the quality of the received traffic has deteriorated to an unacceptably low level due to changes in channel conditions. The transmission rate required of an EOW normally will be considerably lower than that of the traffic channel, thus enabling robust channel encoding (modulation and coding) procedures to be adopted for protection of the system control data.

The system control procedures should also allow for:

(i) Continuous checking of traffic quality: an automatic repeat request (ARQ) technique would be suitable for this purpose.

(ii) Automatic re-establishment of the circuit should contact be lost: probably based upon the use of back-up accurate time and frequency transmission schedules.

In the case of an open loop configuration with no EOW available, eg a broadcast system, the potential for adaptive operation is relatively limited. All adaptation must be carried out at the receiver in response to RTCE data embedded in the transmitted signal. Receiver processing techniques such as equalisation, adaptive filtering, diversity combination of two or more independent versions of the transmitted signal, antenna null and beam steering, etc are relevant in this situation.

From the preceding discussion, it is evident that automatic HF system control requires the application of considerable information processing power. Therefore, future HF communication systems will inevitably be processor based, with manual intervention only in exceptional circumstances.

## 4.3 Channel Encoding/Decoding

The purpose of the channel encoding procedure in a communication system is to condition the information to be transmitted in such a way that it is resistant to the types of noise and other perturbations likely to be encountered during its passage over the channel. The key aspects of channel encoding are therefore modulation and error control coding.

The range of channel encoding/decoding techniques, which offer promise of economic and effective implementation with variable capacity HF systems, has expanded considerably with the advent of increased signal processing power.

As has been indicated previously, signal processing in existing HF systems tends to be relatively simple and based upon the premise of non-adaptive, constant rate operation. If variable rate operation is contemplated, a fresh consideration of potential signal processing techniques is necessary. Those procedures which appear to offer promise are now outlined briefly.

(a) Adaptive filtering at the receiver: in which the parameters of baseband or IF receiver filters are modified to counter changes in the nature of interfering signals in the communication channel.

(b) Quenched filtering: applied to a synchronous transmission system enables a receiver filter to be opened to receive a wanted signal only over the intervals when that signal is expected to be present. In this way, the filter is not initialised by noise or interference and the detection decision is made when the detector output SNR can be expected to be a maximum. This type of detection scheme is used in the PICCOLO system (Bayley & Ralphs, 1972). For a given form of modulation, if synchronous or coherent detection can be applied, a significant improvement in system performance can be achieved.

(c) Soft-decision decoding implies that a "hard" detection decision, say 1 or 0, is augmented by information concerning the confidence level associated with that decision, eg based upon signal amplitude, phase margin, etc. When coupled with error protection coding, soft-decision decoding can enhance the performance of a basic detection system substantially (Chase, 1973).

(d) Possibly one of the most powerful techniques for use in a high interference environment is that of correlation reception. For this, sets of relatively long sequences with approximately impulsive autocorrelation functions are required so that matched filtering can be applied at the receiver. Sequences with suitable autocorrelation properties also allow RTCE data, say in the form of a channel impulse response, to be extracted conveniently.

(e) Bearing in mind the variable nature of HF channel capacity discussed previously, it would appear that some form of "embedded" data encoding would be appropriate. Here, several versions of the source data, each keyed at a different rate, would be combined into a single transmitted signal format. At the receiver, data would be decoded at the highest rate which the channel capacity at that time would allow. System control and re-alignment would be accomplished at intervals by an ARQ arrangement operating via the EOW (Darnell, 1983a).

(f) With the advent of cheap and powerful computing capacity, it is now possible to consider the introduction of post reception processing at the receiver. The unprocessed received signal could be stored, either at a low IF or baseband, and then processed rapidly off-line using different signal processing parameters and algorithms until a best estimate of the transmitted signal is obtained.

## 4.4 Radiated Power Levels

In the HF band, high radiated power levels are undesirable since they tend to cause spectral pollution and an increase in transmitter and antenna sizes - and hence costs. The use of RTCE in optimising the primary parameters of an HF communication system will tend to reduce the necessity for higher radiated powers by selecting frequencies and transmission times for which the received SNR is maximised and interference avoided. As a general principle, system availability and reliability should be improved by the use of RTCE and more effective signal processing, rather than by transmission at higher power levels; the latter should be held in reserve for particularly vital links.

The RF units of the system should be capable of rapid frequency agility so that full advantage can be taken of automatic and adaptive system control procedures. The ability to change frequency rapidly will also facilitate the RTCE process of finding a channel with an acceptable SNR. This will eliminate the "resistance" of system operators to changing frequency except where absolutely essential and thus lead to improved flexibility and lower radiated power levels.

## 4.5 Antenna Characteristics

Diversity using spaced receiving antennas is an effective method of improving HF system performance and should, where possible, be incorporated into a system design. Also, a simple 2-element receiving antenna system can be used to place a null in the direction of an interfering signal. The depth of the null will depend upon the instantaneous propagation conditions, but an average depth of a few dB may well be achievable for skywave signals. Against more stable groundwave interfering signals, the nulling procedure will be considerably more efficient.

Again, the systematic use of RTCE and improved system control procedures will tend to reduce dependence upon antenna characteristics.

### 4.6 System Design with Embedded RTCE

It is clear that, particularly in a high-interference environment, a constant rate transmission system will provide a variable level of performance, and at rates of 2.4 kbits/s is likely to have a low overall reliability. Many regions of the world have high interfering signal densities such that the probability of finding a clear frequency slot in which to place a 2.4 kbits/s transmission is extremely low (Gott et al, 1983). Thus, it can be inferred that requirements for constant data rate transmissions of this form are in many cases incompatible with the fundamental nature of the propagation medium when its properties are considered on a world-wide basis. This has a particularly serious implication for digital speech links, which normally operate at rates in the range 1.2 - 2.4 kbits/s, since in regions where the HF spectrum is congested, eg in central Europe, they can rarely be expected to provide a satisfactory grade of servie over extended periods. The use of real-time channel evaluation (RTCE) to select the optimum assigned channel at any instant will only be of limited effectiveness in such an environment because a channel with the required transmission characteristics will frequently not be available.

There are certain techniques which can smooth out the variatios in channel capacity as seen by a non-adaptive communication system (Darnell, 1977). In the context of the window model described earlier, this corresponds to a reduction in required SNR with a consequent expansion of window dimensions. These techniques include:

(a) diversity processing;

(b) adaptive equalisation;

(c) error control coding;

(d) antenna beam-forming.

In practice, RTCE can be used to select a frequency channel having suitable propagation characteristics and clear of significant levels of co-channel interference. This then enables techniques such as those listed above to be applied more effectively to increase the transmission window volume.

The potential of the ionospheric medium to transfer certain forms of traffic reliably, even in the face of high levels of interference or disruption, is of great operational importance. However, in order to realise this potential fully, HF communication systems must be designed and operated in a rather different manner from that associated with the majority of systems currently in service. Fig. 13 shows the elements of a duplex ionospheric communication system incorporating many of the features discussed previously. It is assumed that the system will use any appropriate mode of propagation. The function of the RTCE sub-system is to select the channel and characterise it in terms of its window structure as a function of required SNR at the receiver; this then allows the parameters of the signal processing unit to be adjusted to match the nature of the available windows, eg in terms of block length, modulation type, etc. Data is buffered until the transmission system is ready to accept it, with simultaneous transmission in each direction being based upon an ARQ mode of operation where control data and traffic signals are interleaved. At the receiver, the incoming data blocks are re-assembled in another buffer prior to output. It should be noted that both data and RTCE signals are multiplexed together and thus make use of single transmitters and receivers at each site, with no dedicated RTCE RF equipment, ie an embedded RTCE system. Receivers would normally be of the diversity type.

## 5. DESIGN TRENDS IN SYSTEMS INCORPORATING MOBILE TERMINALS

As explained previously, mobile terminals represent an important class of disadvantaged users of the HF spectrum and it is important that the energy they do radiate is received in the most efficient manner possible. The technique offering most promise of achieving this may be termed "geographical diversity reception".

In Section 3.3, a network of widely-separated static terminals for the reception of mobile transmission was described. This concept will now be extended to that of a geographical diversity system in which data from a number of spaced receiving stations can be combined in a systematic manner to provide an improved estimate of a desired mobile transmission. The value of such path diversity in improving communication availability has been demonstrated experimentally (Rogers & Turner, 1985).

Fig. 14 is a schematic diagram of a geographical diversity network involving multiple static terminals interconnected by means to be described in Section 5.1. The basis of operation of the network is that transmissions emanating from a given mobile terminal will be received at the static terminals via propagation paths which will normally be skywave, but may also be surface wave; these paths can be considered as independent if the separation of the static termials is chosen appropriately. Similarly, interfering transmissions will also be received via independent paths. Thus, taking the network as a whole there will be a high probability that the signal and interference conditions on those links which are propagating between mobile and static terminals will be uncorrelated. This situation lends itself to some form of diversity combining in which all received versions of the mobile signal are brought together at a central control and processing site to allow an improved signal estimate to be made. There are several possible combination algorithms which could be considered, eg:

(a) selection of the path giving the greatest SNR at any time, ie selection diversity combining;

(b) alternative classical forms of diversity combining such as maximal ratio (Stein & Jones, 1967);

(c) operation on the demodulated and shaped versions of the propagating received signals by say majority voting;

(d) more elaborate combining techniques applied to demodulated and shaped signals, taking account of soft-decision data associated with the received data symbols.

The most important considerations associated with the establishment of a geographical diversity network will be reviewed in the following sections.

### 5.1 Interconnection Considerations

The static terminal interconnections shown in Fig. 14 are clearly vital to the operation of the system; they need therefore to be robust and reliable. The following media may be considered for interconnection purposes:

(a) HF skywave and surface-wave point-to-point links;

(b) meteor-burst links;

(c) ionospheric or tropospheric scatter links;

(d) satellite links;

(e) some form of common-user telephone/telecommunication system, eg such as those implemented by PPT's.

Links of types (a), (b) and (c) are essentially "stand-alone", point-to-point interconnections, whilst (d) and (e) require the availability of separate communications systems, possibly controlled and operated by independent organisations. Hence, from the viewpoint of survivability in the face of physical or electromagnetic attack, (a), (b) and (c) are to be preferred. The most promising candidates will now be discussed in greater depth.

#### 5.1.1 HF Interconnection

In a geographical diversity network of the form proposed, fixed receiving terminals may be separated by distances from say a few hundreds of kilometres to several thousands of kilometres. Interconnection via HF links will thus involve skywave, and possibly surface-wave, paths exhibiting a wide variety of propagation characteristics. There is a need for data transfer between terminals to be effected with high integrity and resistance to disruption. This will require links with sophisticated and possibly embedded RTCE (Dawson & Darnell, 1985), together with adaptive error control (Hellen, 1985). Inevitably, the data rates which can be sustained reliably will be relatively low.

#### 5.1.2 Meteor-Burst Interconnection

Meteor-burst links (Bartholomé & Vogt, 1968) make use of the short-duration ionised trails, caused by the passage of micrometeorites through the lower ionosphere, in order to reflect radio wave energy. Such links are naturally intermittent, although the statistics of meteor trail occurrence are relatively predictable. It has been demonstrated experimentally that average data rates of a few hundreds of bits/s can be sustained over a complete 24-hour period, which would normally be sufficient for the role envisaged here. The diurnal and seasonal variability of meteor-burst links is less than for the corresponding HF skywave links; they are also less susceptible to excess absorption effects following severe ionospheric disturbances.

The most effective frequency range for meteor-burst operation is from abot 25 to 100 MHz which allows compact, efficient and directive antennas to be used.

#### 5.1.3 Ionospheric or Tropospheric Scatter Interconnection

Ionospheric scatter propagation makes use of radio wave energy scattered by the lower ionosphere (Bartholomé & Vogt, 1965). Tropospheric scatter propagation employs scattered energy emanating from the troposphere (Yeh, 1960). Since the troposphere is at a much lower altitude than the lower ionosphere, the ranges achievable by tropospheric scatter are substantially less. In both cases, the scatter loss is high which necessitates the use of high transmitter powers and directive antennas to illuminate the "scattering volume" efficiently.

Propagation via these scattering mechanisms is relatively reliable, although significant fading may occur.

5.1.4 Satellite Interconnection

Inexpensive and reliable satellite links are becoming available, although there may be coverage limitations at high latitudes. In a military context, the HF system should operate as an independent entity under threat conditions which might involve the satellite being jammed. For this reason, such links are not considered suitable for interconnection purposes, although they may be useful in benign operating environments.

5.1.5 Common-User System Interconnection

Possibly the cheapest interconnection medium is a PTT-type of network in which all fixed HF terminals and control sites are subscribers. Again, this has the disadvantage that it can be disrupted by physical or electromagnetic attack. However, the diversity of propagation media and routes make it inherently more robust than a satellite system alone.

Practically, a combination of interconnection media would appear to offer the best prospect for reliable operation over the whole range of operational conditions. Transmission systems covering say the range from 2 to 60MHz, having relatively high radiated powers (a few kW) and directive antennas, could form the primary basis for interconnection; they could operate in HF skywave, HF surface-wave, meteor-burst or ionospheric scatter modes, as available and appropriate. Also, a parallel interconnection could be established via a PTT network.

Obviously, some delay will be introduced into the reception of mobile terminal signals by the need to transfer data around the network. This delay will tend to increase as the number of mobile terminals accessing the network simultaneously increases. The longer the delay which can be tolerated, the greater is the potential for accurate reception. In general, it can be expected that the fixed terminal interconnections will be able to sustain a higher average data rate than that which will be associated with the mobile-to-static terminal transmissions. The effect of delays can be minimised by the signal design and network control strategies to be discussed in the following sections.

5.2 Signal Design Considerations

From the discussion in Sections 1 and 2, it is seen that ionosphere can be subject to severe natural disturbances giving rise to multipath delay, signal fading, low signal levels and possibly abnormal doppler spreads or extreme phase instability in polar regions. Under these circumstances, it is necessary to use robust transmission formats which are resilient to such perturbations.

5.2.1 Mobile Signal Formats

Normally, it will be necessary for a geographical diversity network to handle signals originating from several distinct mobiles. A block transmission format would be convenient since it would allow simple time and frequency multiplexing to take place, with only one mobile using a particular frequency at a specific time. To allow this, accurate time and frequency standards would have to be available at all terminals in the system, but this can now be accomplished relatively simply and economically.

The major requirements for a robust mobile transmission can be summarised as:

(a) that its time of occurrence and frequency of propagation are accurately controlled;

(b) that the SNR required at the receiver for a specified error rate is minimised;

(c) that no elaborate synchronisation techniques are necessary;

(d) that long multipath delays can be tolerated;

(e) that high levels of phase instability can be tolerated;

(f) that moderate doppler frequency offsets can be tolerated;

(g) an ability to operate in any of a limited number of modes which can become progressively more robust as propagation conditions deteriorate;

(h) that signals should be digital to enable data to be protected by encryption.

A possible signal format on which the mobile transmission could be based is basic wide-shift FSK, in which detection is accomplished by a Law assessor arrangement (Alnatt et al, 1957) givng the effect of dual, in-band diversity under frequency-selective fading conditions. Simple block error control codes could be applied to the data blocks, with more powerful (more redundant) coding being used as path conditions worsened. In extreme cases, the mobile would simply repeat its message many times to allow signal averaging to be employed at the receivers. Under white noise conditions, signal averaging improves the effective SNR by a factor proportional to the square root of the averaging time (Hewlett-Packard J, 1968).

Under poor SNR conditions, a trade-off must be made between SNR and transmission time; ie as the SNR reduces, the user must allow more time for the transmission of a given message if it is to be received with a specified level of fidelity. In effect, variable redundancy error control is required (Hellen, 1985).

In circumstances where frequency offsets are negligible and short-term phase stability reasonable, systems of the PICCOLO type (Bayley & Ralphs, 1972), or reduced multiple FSK systems, would provide robust communication.

In general, transmissions should be designed so that soft-decision data, or confidence information, can be extracted readily at the receiver.

### 5.2.2 Interconnection Signal Formats

Most of the comments made above in the context of mobile signals also apply to signals transferred over the interconnection links between static terminals.

Labelled data blocks relating to the signals from specific mobiles would be a convenient vehicle for passing data between the static nodes and control/processing site.

As indicated previously, the design and operation of reliable point-to-point links is a simpler problem than is the design of a mobile-to-static link because more efficient antennas and higher transmitter powers are available. Again, an ARQ mode of operation would seem to be appropriate (Dawson & Darnell, 1985), since this would be well-matched to the characteristics of the interconnection media proposed, ie HF skywave, meteor-burst or ionospheric scatter.

## 5.3 System Control Considerations

From the previous section, one of the potential problems which can be identified is that the maximum capacity required for the interconnection links is considerably greater than for the mobile-to-static links. This arises because the interconnection links may be carrying simultaneously information relation to:

(a) a number of distinct mobile transmissions;

(b) system control data for the network.

These will now be discussed in greater depth, together with other considerations affecting the control of the network.

### 5.3.1 Data describing Mobile Transmissions

One mode of operation for a geographical diversity system is where the various versions of the mobile signal received at the different static terminals are demodulated and reconstituted at those terminals, with the reconstituted data being forwarded to the site where combination is to be carried out. A further refinement to this process could be the transmission of soft-decision data, relating to each of the detected data elements, to the combination site. Alternatively, the complete analogue baseband signal from each static receiving terminal could be forwarded to the combination site. In the latter two cases, the information carried by the interconnecting links would be considerably increased in comparison with that contained in the original signal transmitted by the mobile. However, since the mobile transmissions would normally be of short duration, this data expansion would be acceptable if some decoding delay is admissible.

### 5.3.2 System Control Data

It will also be necessary for system control data to be transferred on a regular basis between static terminals, simply to maintain the integrity of the network. This system control data would relate to frequency selection options, available data rates, changes in coding levels, etc. It is envisaged that, in a ARQ-based system, the capacity overhead associated with this data would be relatively small.

### 5.3.3 Frequency Management

Frequency management of a geographical diversity system would logically be carried out at two levels:

(a) on a link-by-link basis;

(b) on a network basis.

In the former case, each of the interconnection links would have its own embedded RTCE and protocols (Darnell, 1986) operating over the frequency range of from 2 to about 60 MHz. The propagating mechanism giving the greatest predicted SNR would typically be selected.

At the network level, there is a need for a more sophisticated RTCE procedure which attempts to assess the general propagation state of the region in which the static and mobile terminals are situated. This could be accomplished by either a vertical-incidence ionosonde placed at the eastern edge of the region, or by oblique-incidence sounding carried out between the static terminals of the network. Measurements of this type would indicate the onset of ionospheric disturbances and possibly their approximate geographical extent.

An additional function of the overall frequency management would be to ascertain the extent to which frequencies could be shared or re-used by different static terminal interconnection links.

### 5.3.4 Other Control Functions

In military operations, the control of the communications system must, in part, be directed towards maintaining the security of the mobile traffic. Thus, the mobile would, for example, be assigned frequencies which would have the minimum probability of being intercepted or disrupted in the prevailing propagation conditions. Also, the mobile would exercise radiated power control (RPC) in response to control data sent to it by the static network.

The system control procedures would also seek to make maximum use of relatively local and transient propagation phenomena as they became available. There would also be a need to reconfigure the static terminal interconnections in order to avoid localised ionospheric disturbances. In this context, Fig. 15 and 16 illustrate the benefit to be gained from the use of a geographical diversity network. Fig. 15 shows a situation in which an ionospheric disturbance affects the skywave path between the mobile (M) and a static terminal (F1); the skywave path to static terminal (F2), at a greater distance from M, avoids the disturbed region completely, as does the interconnecting path between F1 and F2. Figure 16 is a more complex situation where the region of disturbance affects both the mobile-to-static terminal (F1) path and interconnections from F1 to F3, and from F1 to F4. In this case, the interconnecting links must be re-routed as shown to maintain network connectivity.

Clearly, similar re-configuration procedures could be employed to offset attempts to disrupt mobile transmissions reaching specific static terminals.

### 5.4 System Design

An attempt will now be made to coalesce many of the concepts outlined in the previous sections of this paper into an overall system design to provide improved HF communications reliability and survivability, particularly in high-latitude regions.

Fig. 17 illustrates a possible format for such a system in which mobile-to-static transmissions are received by a geographical diversity network, whilst static-to-mobile transmission is carried by single broadcast system sited at a convenient location.

System control and signal generation and combination are carried out at a specified control and processing site. Interconnections are implemented by 2 to 60 MHz ionospheric links, or via a PTT network, as appropriate. Broadcast traffic for all mobiles in the operational area is transitted using a multiple sub-channel format from which each mobile can extract the data addressed to it. The EOW sub-channel of the broadcast is used to request repeats of mobile data blocks received in corrupted form, and also to disseminate frequency management data to all mobiles who may wish to access the system.

Data transfer between the HF system and message originators and recipients can be effected via a PTT network or possibly via other dedicated HF or non-HF links.

Fig. 18 shows an alternative system format where the centralised broadcast system of Fig. 17 is replaced by a distributed network of lower power transmitters, possibly co-sited with the static receiving terminals. Thus, in principle, for any specified mobile position, a broadcast transmitter site having maximum probability of successfully propagating to that mobile can be selected - as illustrated in Fig. 18.

## 6. CONCLUDING REMARKS

This lecture has attempted to provide an introductory technical framework prior to a more detailed discussion of specific aspects of HF communications in subsequent lectures.

From the preceding discussions, the following points should be made in conclusion:

(a) That the variable capacity nature of the HF path should be both recognised and its characteristics exploited in the design of HF communication systems.

(b) That improved performance should be achieved by the application of more advanced signal processing and system control procedures, rather than by the application of higher powers and large, fixed antenna arrangements.

(c) That the key to improved performance is the effective integration and use of RTCE in an HF system design - as will be discussed further in a companion lecture.

(d) That HF systems should have a capability of operating in a frequency range above the traditional upper limit of 30MHz. This would provide additional resistance to ECM activity.

(e) That the most effective method of enhancing the communications reliability of mobile terminals would seem to be via the use of a geographical diversity system. A distributed network of the type illustrated in Figs. 17 and 18 poses additional control problems. However, the increase in operational flexibility and robustness potentially available from such a system would appear to merit further research. In particular, the ability to reconfigure in the face of ionospheric disturbances, or physical or electromagnetic attack, would be most valuable.

(f) That the price paid by an interceptor or jammer for his activities should be maximised - both from the cost viewpoint and that of the complexity of analysis he must employ.

(g) That the HF medium should be used for types of communication which are compatible with its fundamental properties.

Following from the discussion above, it would appear that forms of traffic which are best matched to the characteristics of the ionospheric path include:

(i) transmissions involving a low constant data rate, R, eg low-speed telegraphy, where R conforms to expression [5];

(ii) transmissions involving a low average data rate, where the data can be block-formatted and transmitted in packet form - possibly at relatively high, but variable, rates;

(iii) transmissions for which it is important to transfer limited amounts of information reliably, perhaps in the face of deliberate attempts at disruption.

The transmission of digitised speech at rates between 1.2 and 2.4 kbits/s will always be difficult to maintain with an acceptable level of reliability: other techniques, such as the combination of speech recognition and synthesis mentioned in Section 3.6, allow a capability for speech message transmission at very low data rates.

From the technical viewpoint, the areas requiring further investigation include:

(a) The problems of networking at HF.

(b) The design of efficient embedded RTCE procedures.

(c) The design of adaptive signal generation and processing systems.

(d) The design of more realistic simulation facilities to aid HF system development, particularly in respect of co-channel interference models and the effects of RF equipment characteristics.

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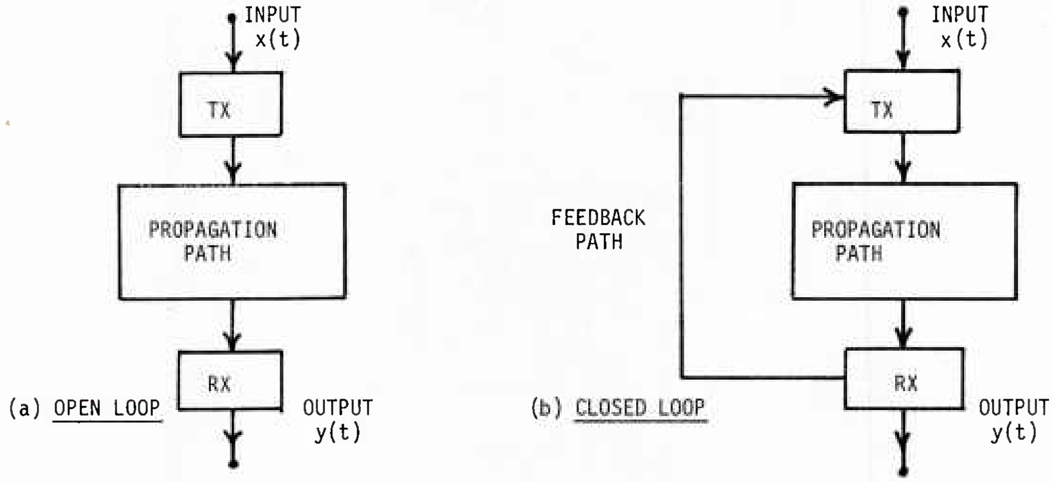


FIG. 1 BASIC COMMUNICATION SCENARIOS

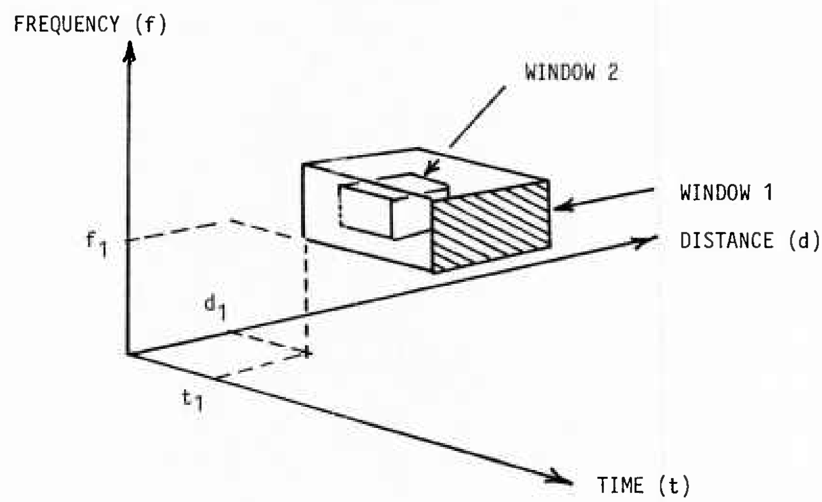


FIG. 2 WINDOW MODEL OF IONOSPHERIC PROPAGATION

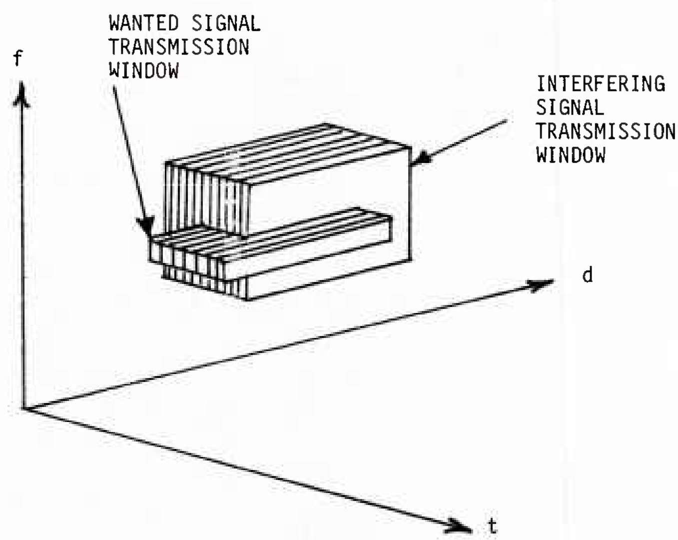


FIG. 3 EFFECT OF INTERFERENCE IN THE WINDOW MODEL

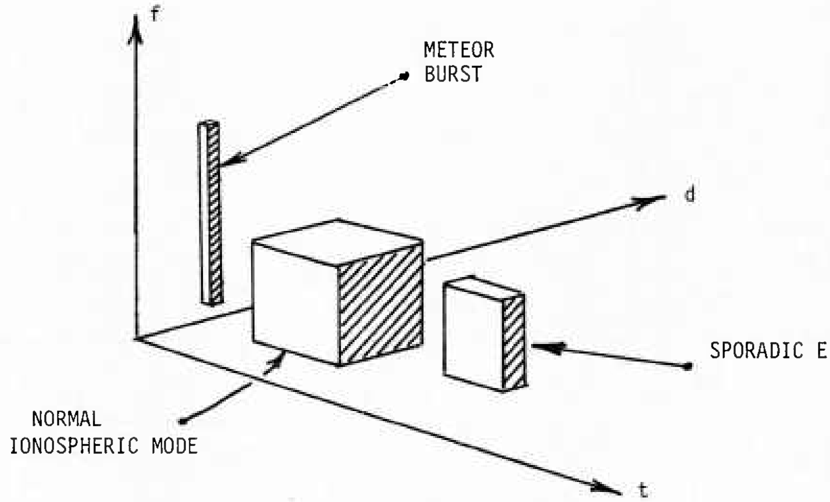


FIG. 4 WINDOWS FOR DIFFERENT PROPAGATION MECHANISMS

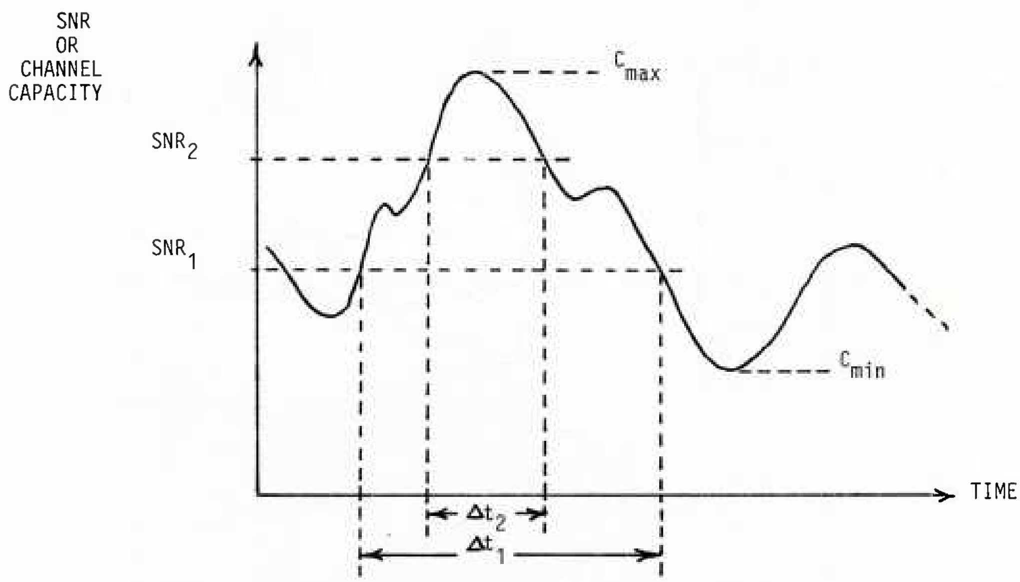


FIG. 5 VARIATION OF SNR/CHANNEL CAPACITY WITH TIME

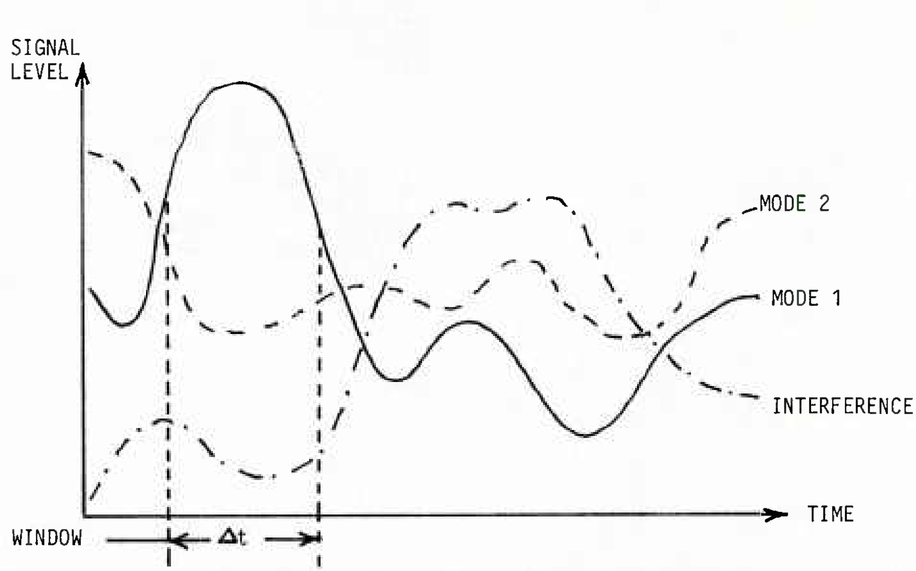


FIG. 6 EXAMPLE OF A PHYSICAL MECHANISM GIVING RISE TO IONOSPHERIC WINDOWS

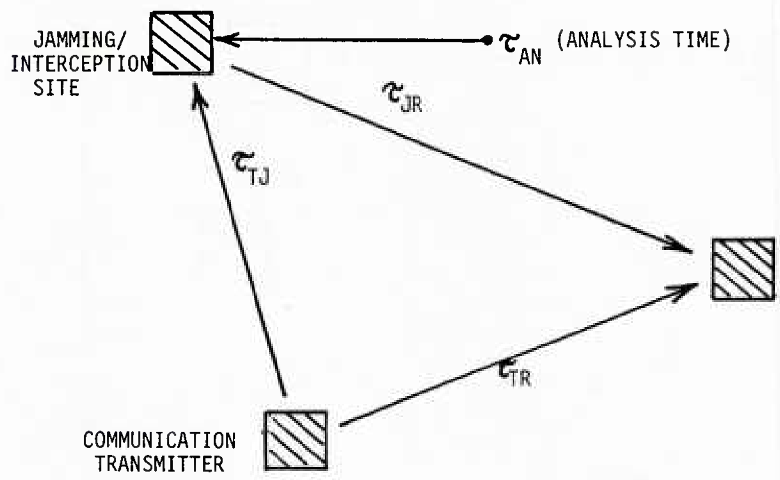


FIG. 7 PROPAGATION AND ANALYSIS TIME LIMITATION

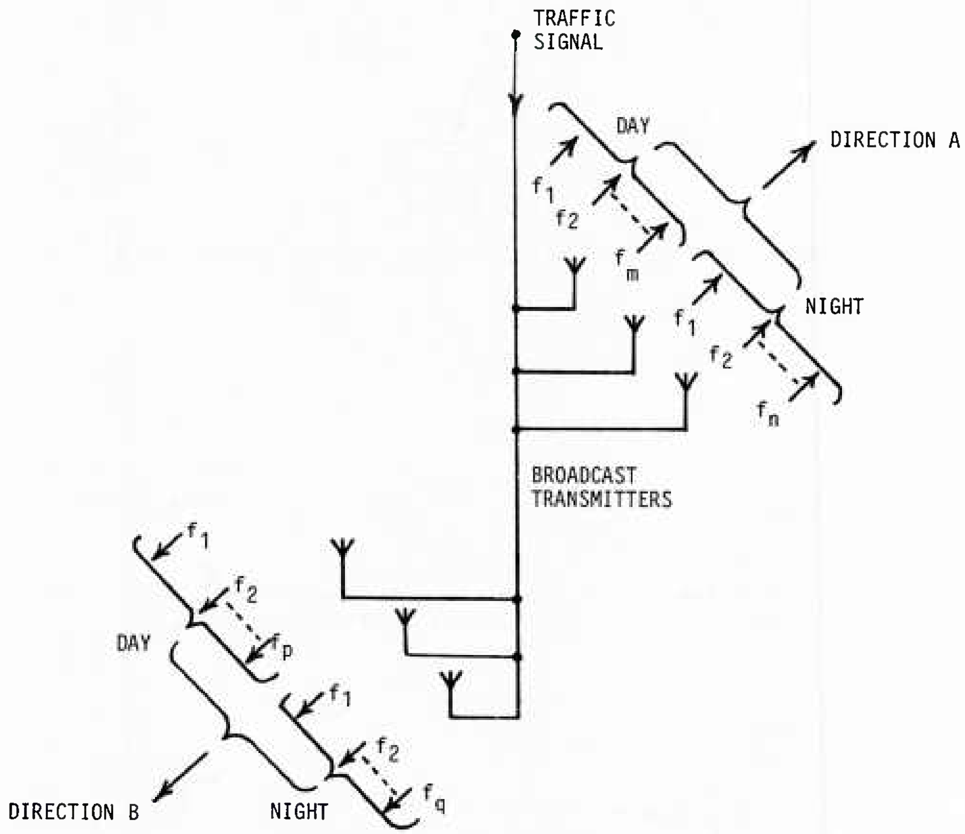


FIG. 8 FIXED-TO-MOBILE BROADCAST SYSTEM

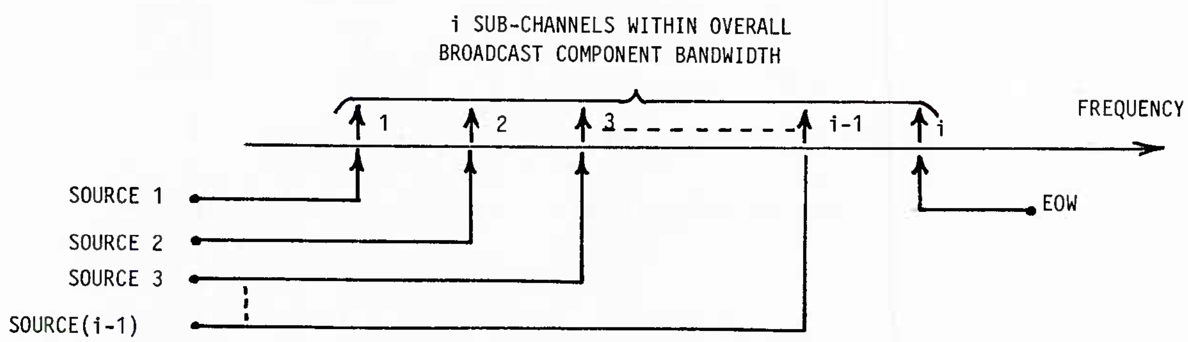


FIG. 9 BROADCAST SUB-CHANNEL RASTER

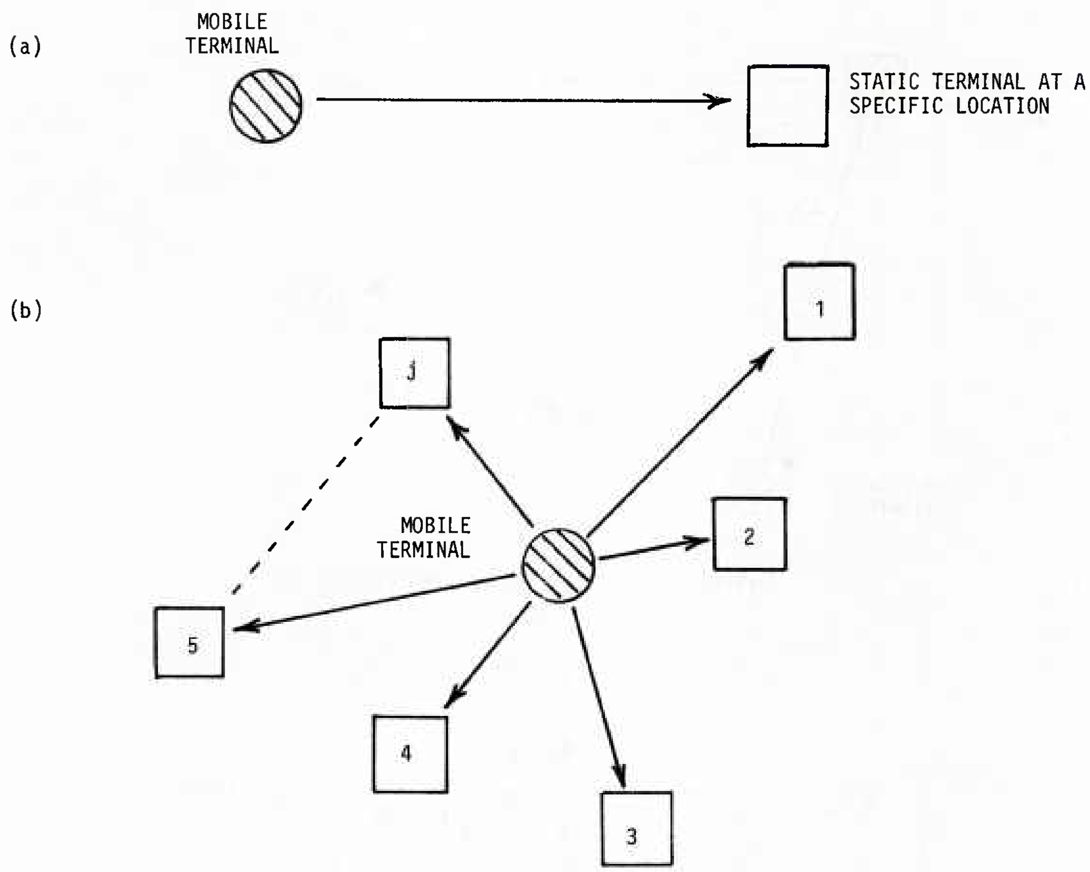


FIG. 10 OPTIONS FOR MOBILE TERMINAL-TO-STATIC TERMINAL COMMUNICATION

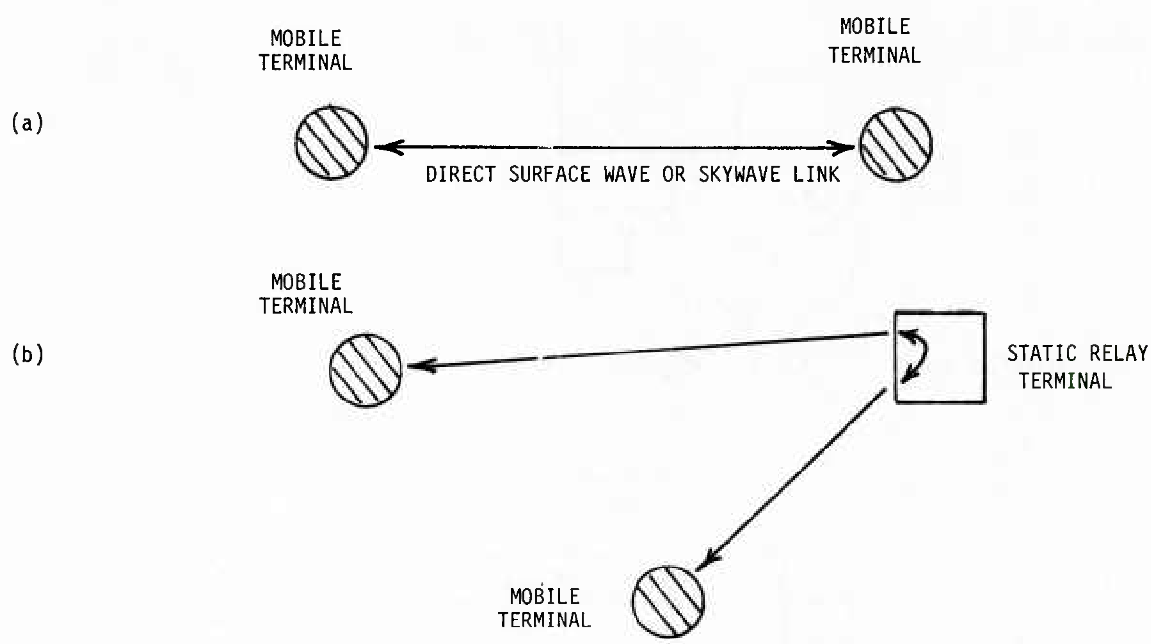


FIG. 11 OPTIONS FOR MOBILE TERMINAL-TO-MOBILE TERMINAL COMMUNICATION

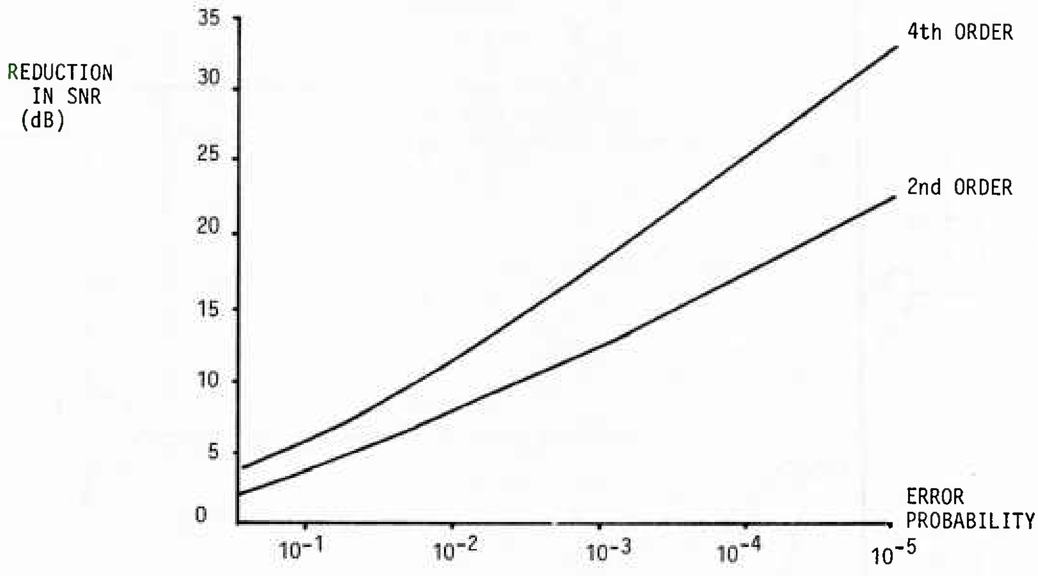


FIG. 12 EFFECT OF 2nd AND 4th ORDER DIVERSITY COMBINING

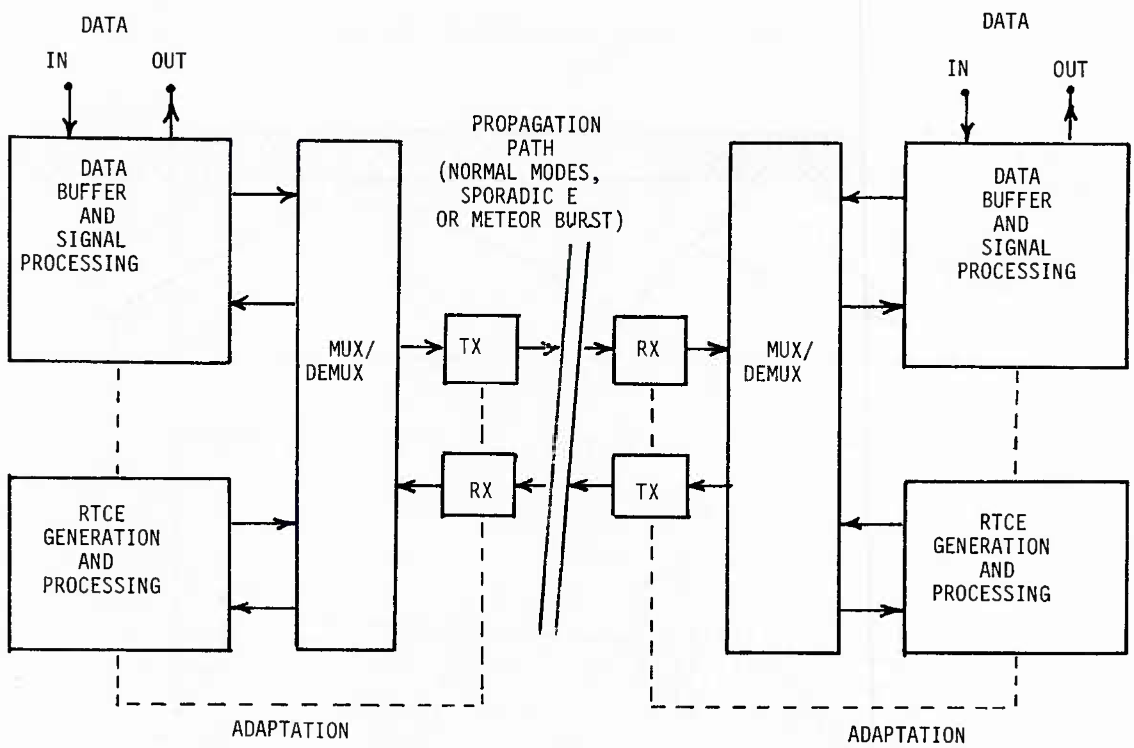


FIG. 13 SIMPLIFIED BLOCK DIAGRAM OF IONOSPHERIC COMMUNICATION SYSTEM

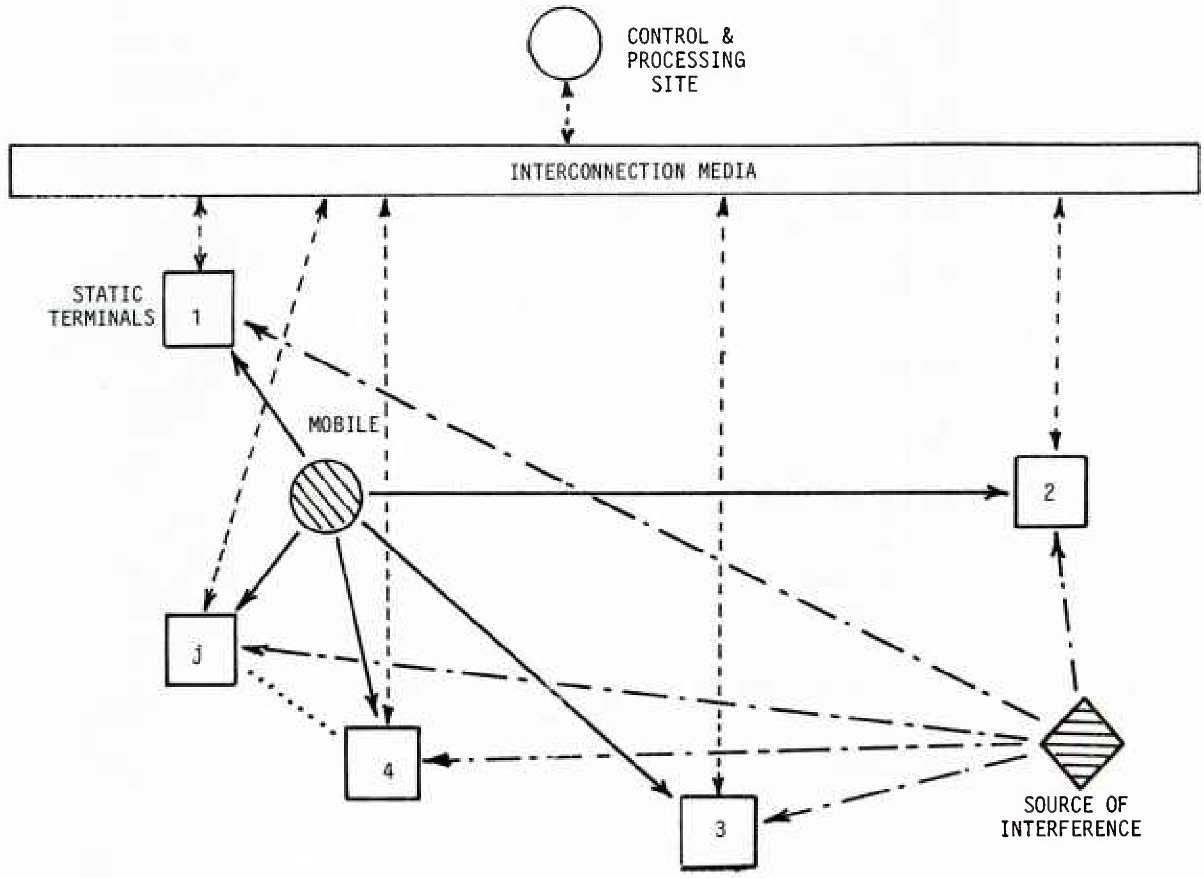


FIG. 14 SCHEMATIC DIAGRAM OF GEOGRAPHICAL DIVERSITY SYSTEM

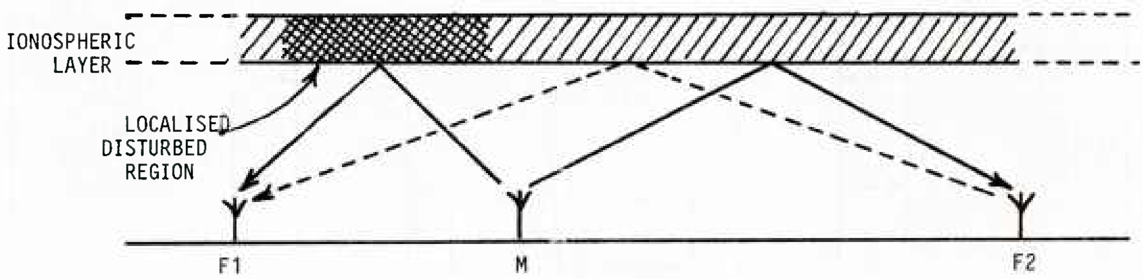


FIG. 15 AVOIDANCE OF A REGION OF LOCALISED IONOSPHERIC DISTURBANCE

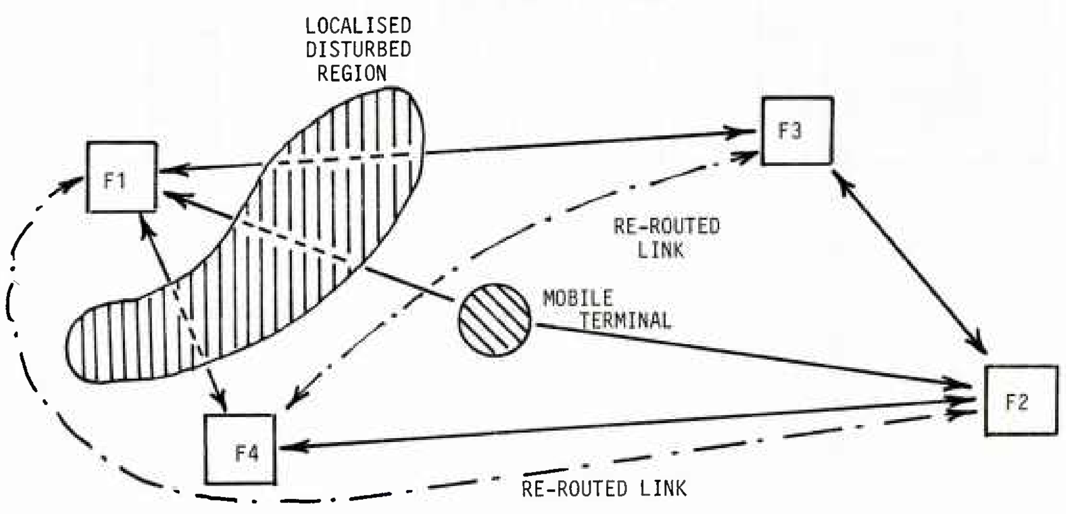


FIG. 16 RECONFIGURATION OF A GEOGRAPHICAL DIVERSITY NETWORK

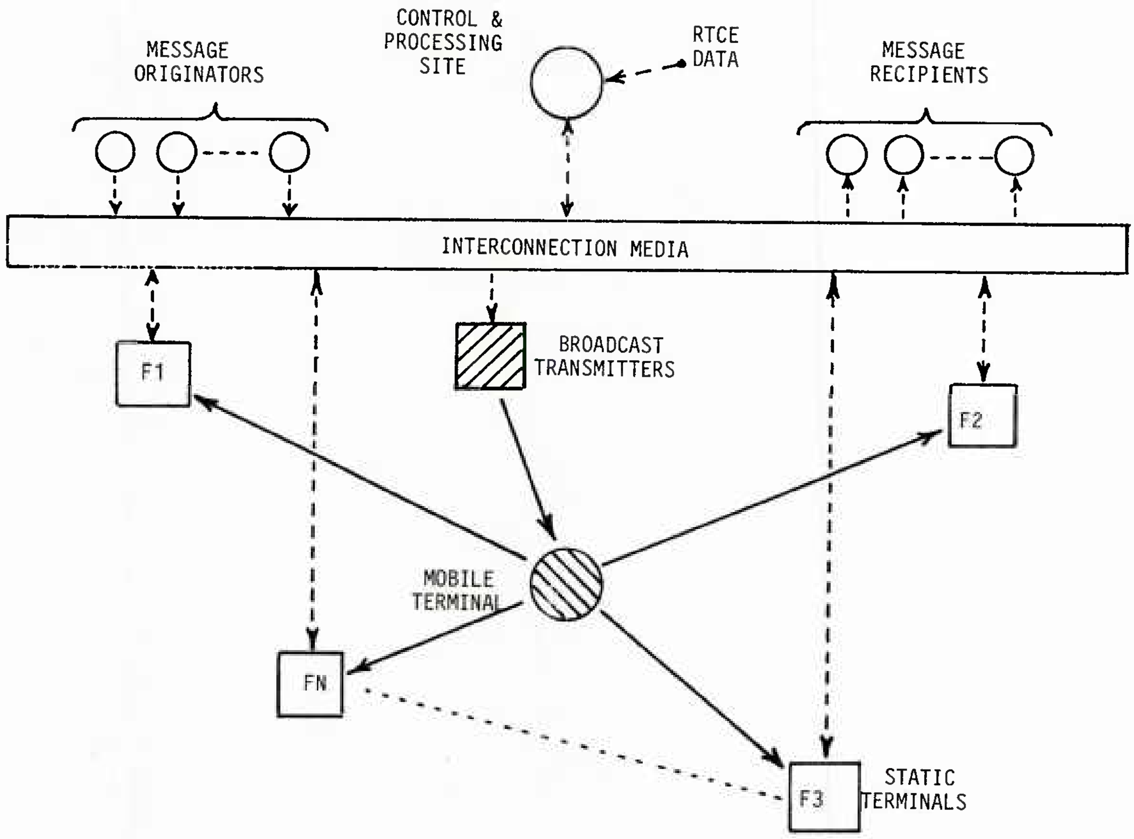


FIG. 17 COMMUNICATION SYSTEM FORMAT 1: CENTRALISED BROADCAST

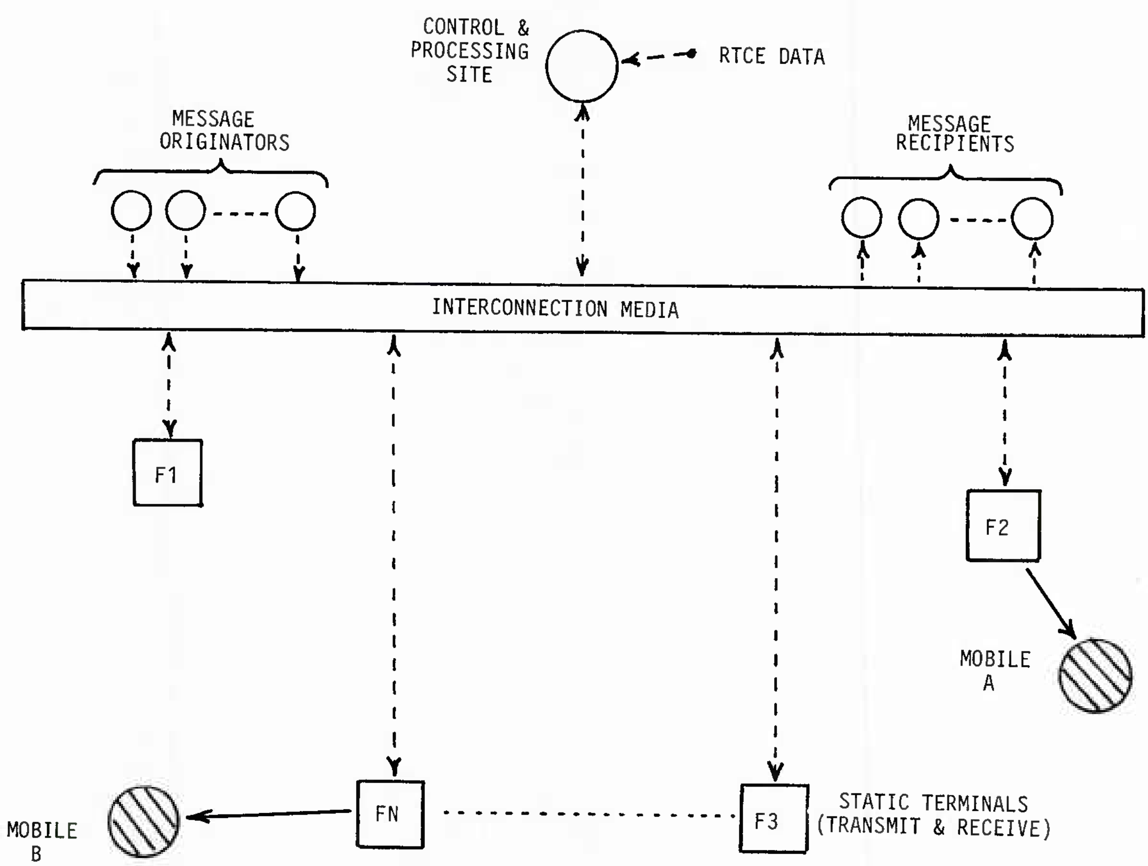


FIG. 18 COMMUNICATION SYSTEM FORMAT 11: DISTRIBUTED BROADCAST

# PROPAGATION I. State of the art of modelling and prediction in HF propagation

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ABSTRACT

The lecture reviews the state of the art in HF propagation modelling and describes the principles of radio frequency predictions. As a basis for the discussions, a brief introduction is given to the ionospheric parameters of importance to HF-propagation. Current methods for frequency prediction are semi-empirical, that is they depend upon a large data base of ionospheric observations, combined with physical models of the ionosphere and of radio wave propagation through the medium. In addition models of the noise and interference environment must be included. The lecture discusses the principles on which the methods are based, as well as their limitations. Examples are given of the use of predictions in system planning and communications. The relative importance of skywave and ground wave communications in the HF-band is discussed.

## 1. INTRODUCTION AND OUTLINE

In spite of the development of satellite communications, microwave links and other communication techniques, interest in the high frequency band is still strong. The HF-band (3-30 MHz) represents a valuable natural resource and should be exploited as efficiently as possible. The radio waves in this band can travel along the earth's surface, but only for limited distances, depending upon the conductivity of the ground or sea. Efficient reflection of the waves from the ionized layers in the upper atmosphere makes HF radio transmissions over long distances possible. A prediction of the performance of a communication circuit must of necessity be based upon a model of the propagation medium, as well as of the propagation process itself. The purpose of the present lecture is to review the state of the art of such models, their usefulness and accuracy. First a brief introduction to the most important principles of ionospheric radio propagation will given. The ionospheric models have been developed from physical principles governing the formation of the ionosphere and a large data base acquired through many years of measurements. Both the data base and the models will be discussed.

Furthermore the principle of the prediction techniques will be described, and how these techniques include models of the electromagnetic noise and interference environment. Weaknesses in present prediction methods and possibilities for future improvement will be pointed out, and finally examples of the use of predictions in system planning and communications will be given.

## 2. A BRIEF INTRODUCTION TO IONOSPHERIC PARAMETERS OF IMPORTANCE TO HF-PROPAGATION

Ultraviolet radiation, X-rays and energetic particles from the sun and interplanetary space interact with the earth's atmosphere to form the ionospheric layers of free electrons and ions in the height range 50-500 km above the earth's surface. The free electrons in the weakly ionized ionospheric plasma absorb, refract and reflect electromagnetic waves over a wide frequency range. The physics of radio wave propagation in this medium is now well understood. Propagation can be described by the magneto-ionic theory developed by Appleton and Hartree (Budden 1960). In the principle, if the properties of the medium is known precisely, the refractive index of an incident radio wave may be computed and the phase and amplitude of the wave may be found at any point in space. The presence of the earth's magnetic field makes the ionosphere birefringent, and the spatial and temporal structure of the ionospheric layers may be very complex. Therefore, accurate solutions to be the wave equations require complex and time consuming calculations. Modern high speed computers have made it possible gradually to introduce more sophisticated and precise raytracing methods, but in practice the prediction methods are still based upon simplified models of the propagation process as well as of the medium. In this section some ionospheric parameters of importance to HF propagation will be introduced.

### 2.1 Formation of the ionosphere

Figure 1 shows a typical daytime height distribution $N_e(h)$ of the ionospheric electron density and indicate how a spectrum of energetic electromagnetic radiation from the sun interacts with the atmospheric gases to form the D-, E- and F-regions in the undisturbed ionosphere. The figure also indicates the occasional presence of a thin dense "Sporadic E-layer". The ions and electrons formed by the solar radiation are lost through recombination, and a simple form of the continuity equation which determines the ionization balance is:

$$\frac{dN_e}{dt} = q(h,\chi)-\alpha_{eff}N_eN_+ \tag{1}$$

where $N_+$ is the positive ion density and $\alpha_{eff}$ is a constant, the effective recombination rate, $\chi$ is the solar zenith angle. The height and intensity of each layer varies in a systematic manner with solar elevation. A first approximation to this variation is shown for a simple model "Chapman" layer in Figure 2. In the real atmosphere many complex processes may cause deviations from this simple model.

The two ionospheric parameters of prime importance for radio wave propagation are the concentration $N_e$ of free electrons and the average frequency $\nu$ with which an electron collides with neutral molecules and ions. The collision frequency $\nu$ is proportional to atmospheric pressure. Typical height variations of $N_e$, $\nu$ and their product $N_e\nu$ are shown in Figure 3.

### 2.2 Radio wave propagation in the ionosphere

The first principles of high frequency propagation may be understood by considering a simple case in which the earth's curvature and magnetic field are neglected and the ionospheric layers are horizontally stratified. The ray path of a radio wave may then be described by Snell's law of refraction, as shown in Figure 4, where $\mu$ is the real part of the refractive index and $I_n$ is the angle of incidence upon the layer labelled n+1. Then

$$\mu_0 \sin I_0 = \mu_1 \sin I_1 = \mu_2 \sin I_2 \ldots = \text{constant} \tag{2}$$

This means that as the ray travels through the slabs the ray direction changes in such a way that $\mu \sin I$ is constant. A continuously varying medium may be approximated by many thin slabs. If the wave is incident upon the ionosphere from free space so that $\mu_0 = 1$, we have $\mu \sin I = \sin I_0$. Reflection occurs where $I = 90^0$, so that $\mu = \sin I_0$. At vertical incidence $I_0 = 0$ and the reflection condition becomes $\mu = 0$. Magneto-ionic theory shows that, in the simple case we are dealing with, the refractive index may be written

$$\mu = \sqrt{1 - \frac{N_e e^2}{m_e f^2 4\pi^2 \varepsilon_0}} \tag{3}$$

where $N_e$ is the number density of free electrons, $m_e$ the electron mass, f the wave frequency and $\varepsilon_0$ the permittivity of free space. The plasma frequency of the propagation medium is defined as

$$f_N^2 = \frac{N_e e^2}{m_e 4\pi \varepsilon_0} \qquad \text{so that} \qquad \mu = \sqrt{1 - \frac{f_N^2}{f^2}}$$

Insertion of numerical values yields $f_N = 9\sqrt{N_e}$ {Hz} where $f_N$ is measured in Hz and $N_e$ is the number of free electrons per $m^3$.

Note that $\mu = 0$ when the wave frequency $f = f_N$, and therefore a wave vertically incident upon the ionosphere is reflected at the height where $f = 9\sqrt{N_e}$. The highest frequency which can be reflected vertically from the layer is called the <u>critical frequency</u>

$$f_o = 9\sqrt{N_{e_{max}}} \tag{5}$$

where $N_{e_{max}}$ is the maximum electron density. Frequencies $f > f_o$ will penetrate the layer. A wave incident obliquely at an angle $I_o$ will be reflected at a height where

$$\mu = \sin I_0 = \sqrt{1 - \frac{81 N_e}{f^2}} \tag{6}$$

and the highest frequency which can be reflected is $f_{max} = 9\sqrt{N_{e_{max}}}/\cos I_o = f_o/\cos I_o$. This frequency is, of course, always greater than the critical frequency $f_o$.

Figure 5 illustrates schematically how the electron density and the refractive index may vary in a simple ionospheric layer, and the ray path through the layer. The wave incident at an angle $I_o$ is reflected back to the ground at a distance d. The maximum frequency which can reach this distance is $f_{max} = f_o/\cos I_o$. If there are no denser ionospheric layers at greater heights, this frequency is the <u>Maximum Useable Frequency</u> (MUF) for the circuit with pathlength d along the earth's surface. Frequencies greater than the MUF will penetrate the layer at the angle of incidence $I_o$. It is possible to show that the wave will behave as if it had travelled along the triangular path indicated in the figure and had been reflected at a virtual height h'. In principle we understand that the MUF for a particular circuit may be derived from simple considerations, if the critical frequency of the layer at the reflection point is known.

Figure 6 shows the results of raytracing for three frequencies in a model layer with critical frequency $f_o$ = 4 MHz and $N_{e_{max}}$ at a height of 300 km. The figure clearly demonstrates that for a frequency $f > f_o$ the wave cannot be reflected back to the ground at distances shorter than a minimum range, called the skip distance. Near this distance there is a focusing effect because waves with different penetration depths into the ionosphere converge.

So far we have only discussed wave refraction, but the ionospheric plasma also acts as an absorbing medium. The electrons oscillating in the incident wavefield reradiates the energy transferred to them from the field, unless their oscillatory motion is disturbed by collisions with the much heavier neutral molecules and positive ions. Such collisions will convert electromagnetic wave energy into thermal energy in the gas, and the wavefield will be attenuated. It is possible to show that the absorption of a high frequency radio wave, measured in desibels, is

$$A \propto \frac{1}{f^2} \int_s \frac{1}{\mu} N_e \nu ds \quad \text{db} \tag{7}$$

where $\nu$ is the collision frequency of an electron, $\mu$ is the refractive index, and integration is along the ray path. We note that A is inversely proportional to the square of the wave frequency, and that the contribution of the integral to A is large where $\mu$ is small and where $N_e\nu$ is large. Figure 3 shows that $N_e$ has a maximum in the D-region where, normally, $\mu \approx 1$. This contribution to the absorption is called the non-deviative absorption. Absorption occurring near the reflection point where $\mu \ll 1$ is called the deviative absorption. For oblique propagation the non-deviative term usually dominates.

The simple model of propagation sketched above may be extended to include the effects of the earth's magnetic field, the earth's curvature and the presence of several reflecting layers. The earth's magnetic field makes the ionosphere birefringent, that is an incident radio wave splits up into two magneto-ionic components, called the ordinary and extraordinary waves. The ordinary wave corresponds most closely to the no-field case, whereas the extraordinary wave has a larger critical frequency and is reflected at a different level. The absorption for the two components may estimated by Equation (7) replacing the wave frequency f with an effective frequency $f \pm f_L$ where + refers to the ordinary and - to the extraordinary wave. $f_L = f_H \cos I$ where $f_H$ is the gyro frequency of the electrons in the earth's field and I is the angle between the direction of propagation and the magnetic field. We note from Equation (7) that the extraordinary wave is more heavily absorbed than the ordinary wave. The ordinary wave will therefore be the important component as far as HF-communications are concerned.

The complex structure of the real ionosphere makes many different modes of propagation possible. Some examples are sketched in figure 7.

Note that the total field at the receiver may be a vector sum of many different modes, including a ground wave which is important at shorter distances. The refractive index of the ground may be expressed as

$$n^2 = \varepsilon - i\,\frac{1800}{f}\,\sigma \tag{8}$$

where $\varepsilon$ is the permitivity, f is given in MHz and $\sigma$ the conductivity in Siemens $m^{-1}$. $\varepsilon$ and $\sigma$ depend upon the properties of the soil, and when n is known the reflecting and attenuating properties of the surface may be derived.

## 3. DATA BASE OF AVAILABLE OBSERVATIONS

Since the presence of ionized layers in the upper atmosphere was experimentally established by Appleton and Barnett (1926) and Breit and Tuve (1926), a large database of ionospheric observations has gradually been compiled to give a global picture of the structure of the ionosphere and its variations. The picture is by no means complete; the problem of mapping the behaviour of the ionosphere is similar to the problem of mapping the weather in the lower atmosphere. More, and more accurate knowledge is always needed, both from a scientific point of view, and from the communicators point of view as the need arises for ever more efficient use of the frequency band available for ionospheric communications.

Under the auspices of the International Council for Scientific Unions (ICSU), several World Data Centers for the collection and exchange of geophysical data have been estalished. The two most important are WDCA in the USA and WDCB in the USSR, but in addition a number of specialized centers exist in other countries. These latter centers emphasize special analysis of data relevant to different diciplines (Piggott & Rawer 1972).

Two international bodies, URSI and CCIR, are of particular importance in the work on collection and evaluation of ionospheric data for communication purposes.

URSI (International Scientific Radio Union) deals with the scientific aspects of radio wave propagation, and is responsible for the Ionosonde Network Advisory Group (INAG) which coordinates ionosonde observations (see below) on an international basis. CCIR (International Radio Consultative Committee) has the International Telecommunication Union (ITU) as its parent body, and promotes international standardization of models for the propagation medium and radio noise environment. Much international cooperation is also being stimulated by the Inter-Union Commission for Solar-Terrestrial Physics (IUCSTP) which promotes intensive studies of particular phenomena.

### 3.1 Observational techniques

The longest time series of ionospheric data stems from ground based observations. The most important techniques used are:

a) The ionosonde, which in its common form is a vertically directed pulsed radar, measures the delay time of the reflected signals in the frequency range 1-20 MHz. This delay time is converted to a virtual height h' of the ionosphere. Figure 8 shows an example of an h'(f) record, or ionogram. The heights and critical frequencies of the E- and F-regions can be read directly from the ionogram, and by means of fairly complex computations, the h'(f) record may be converted to an electron density profile $N_e(h)$.

b) Absorption measurements
The A1 method measures the amplitude of pulsed HF signals reflected vertically from the layers. The A2 method measures the amplitude, at the ground, of VHF radio noise incident upon the earth from the galaxy. The receiving instrument is called a riometer (Relative Ionospheric Opacity meter). Radio noise above the critical frequency of the F-layer will penetrate the ionosphere, as the earth rotates and the receiving antenna sweeps across the sky, a "quiet day" variation of signal intensity will be recorded. The noise will suffer absorption in the lower ionosphere, and a disturbance in the D-region will be recorded as a deviation from the quiet day curve. The riometer is a particularly useful instrument in the disturbed high latitude ionosphere. The A3 method measures the signal strength of a CW MF or HF signal reflected from the ionosphere at oblique incidence, typically over paths of a few hundred kilometers.

c) The incoherent scatter technique involves the use of very powerful, sensitive and sophisticated VHF or UHF-radars which detect weak reflections from plasma irregularities. There are only a few such installations in operation, but they have provided a wealth of information about ionospheric parameters.

Figure 9 shows a map of the distribution of ionospheric observatories. As might be expected, most of them are located in northern hemisphere, and the large oceans and polar regions are not well covered.

The space age introduced rockets and satellites as platforms for ionospheric observations. Rocket flights can yield valuable detailed "snapshots" of ionospheric conditions, whereas satellites have been successfully used to obtain global coverage of some "top-side" ionospheric parameters. The satellite-borne "top-side" ionosondes have been particularly important in global mapping of F-layer electron densities above the maximum of the layer. Unfortunately the top-side ionograms only rarely yield accurate values of the F-layer critical frequencies. Their usefulness in filling in gaps left open in the ground based network is therefore limited.

3.2 The ionospheric data

The ionosonde data are conveniently summarized in daily "f-plots", an example of which is given in figure 10. An example of the long term variation of critical frequencies is shown in figure 11, which clearly demonstrates the 11 year cycle present in the ionospheric parameters. As will be discussed later, this solar cycle dependence is important for prediction purposes. The data base may be used to construct contour maps of ionospheric parameters. An example figure 12 shows a map of foF2 (the critical frequency of the F-layer) compiled for the east zone from data for 40 stations over two solar cycles (Reddy et al 1979).

After more than 50 years of ionospheric observations it is fair to state that the behaviour of the undisturbed mid-latitude and low-latitude E- and F-regions is adequately mapped for many communication purposes. Then are, however, important gaps in our knowledge of ionospheric behaviour during disturbances, and in general the high latitude ionospheres are not well understood. There is also a lack of data, on a global scale, on the structure and behaviour of the lowest part of the ionosphere, the D-region.

3.3 The conducting properties of the earth's surface

The conductivity of the earth's surface is of importance for the antenna gain at the transmitter and receiver sites, and for ground reflections in multihop propagation. There is therefore a need for global conductivity maps, and for some applications there is also a need for terrain modelling. Present models distinguish between land and sea conductivities, using average values of the reflection coefficients, but the inclusion of more detailed maps of ground conductivity would improve propagation predictions. (See section 9)

3.4 The radio noise environment

An HF-signal must be detected against a background of radio noise from natural and man-made sources. The first compilation of radio noise data was carried out by Bailey & Kojan (1943). CCIR (Report 322 1963) has prepared an atlas of radio noise intensities based upon measurements obtained mainly from 16 stations throughout the world during an international cooperative programme. Data were collected from these stations from 1957 to 1961 and the noise power versus frequency are presented in diagrams such as figure 13.

The natural radio noise has two important sources, atmospheric lightning discharges, and galactic cosmic noise. A stroke of lightning produces noise over a wide frequency band, with maximum intensity near 10 kHz. The noise from the thunderstorm activity throughout the world, in particular from the thunder-storm centers in Equatorial Africa, Central America and the East Indies, will propagate to great distances through multiple reflections between the earth and the ionosphere. The galactic noise penetrates the ionosphere from above at frequencies above the critical frequency of the F-layer, and is in general the dominating noise above about 20 MHz. The man-made HF-noise may be important in industrial or densily populated areas, and may have large local variations.

Contour maps of radio noise intensity, such as figure 14 are available for different times of day and season.

# 4. IONOSPHERIC MODELS

Ionospheric modelling for HF-communication purposes aims at a description of the ionosphere and its variations in time and space, which allows prediction of propagation parameters. The model must therefore be simple enough to be practical, and sufficiently complete and accurate to be useful. The degree of complexity and sophistication in a model depends upon the requirements and resources of the user, and the practical solution is always a compromise between the needs for simplicity and for accuracy. There are two different approaches to the modelling of ionospheric circuits. The first is to fit empirical equations to measurements of signal characteristics for different times and paths, the other is to estimate these characteristics in terms of a number of separate factors known to influence the signal (Bradley 1979), such as critical frequencies, layer heights, absorption etc. When a large data base exist for a particular circuit, the former may be useful, but it lacks generality. The second approach has the advantage that a limited data base can be combined with knowledge of physical principles to guide a description of the behaviour of the ionospheric layers.

The first approach has been successfully applied to medium frequency (MF) propagation, whereas it is generally agreed that the second method is the most efficient one for HF propagation.

## 4.1 Modelling of the ionospheric layers

If the ionospheric electron density height profile is known at every point along a propagation path, the signal characteristics at the receiver may be calculated using some form of raytracing procedure. A useful approximation is to neglect variations along the path and assume that the profile at the reflection point (or points if there are more than one hop) may be used in the raytracing. The problem is then to describe this profile in terms of a few measurable parameters, so that the raytracing through the simplified model ionosphere yields realistic signal characteristics, for vertical incidence raytracing should reproduce an ionogram typical for the time and location of the reflection point.

Figure 15 shows four electron density models in order of increasing complexity. The model in Figure 15a includes the E and F-layer only, as simple parabolic layers with specified critical frequencies, heights and thicknesses. This model is used in the first CCIR prediction procedure CCIR (1970) and is at present still recommended by CCIR. A second CCIR procedure is in preparation (CCIR 1978) and the electron density model is shown in Figure 15b. As will be seen, the gap between the E- and F-layer has been filled in, so that the electron density increases linearly with height in this intermediate region. The two CCIR models have been discussed in some detail by Bradley (1979).

Figure 15c shows a model used in a recent method "Ionospheric Analysis and Prediction" (IONCAP) developed at the Institute for Telecommunication Sciencies (Lloyd et al 1982). The region between the E- and F-layer maxima is now even more complex, including a "valley" and two linear segments. This figure also shows a virtual height profile, and demonstrates that the model can reproduce the most important features of an ionogram, including an F1-layer. Note that the IONCAP includes a simple model of the D-region. All three models make some provision for the presence of sporadic E-layers.

Finally Figure 15d shows the electron density profile adopted in the International Reference Ionosphere (IRI, Rawer 1981).

The IRI has been developed by a task group chaired by K Rawer, under the auspices of URSI and COSPAR (Committee for Space Research). The International Reference Ionosphere is described by Rawer (1981). It represents a compendium of height profiles through the ionosphere of the four main plasma parameters: plasma density, plasma temperature of electrons and ions, and ion composition. These parameters are generated from reliable data including both ground based, rocket and satellite data, and the IRI is thus primarily an experimental, not a theoretical model. A computer code generates height profiles for any time of day, position and sunspot number.

The IRI electron density profile models all ionospheric layers and is quite complex. It has therefore not yet been adopted or recommended by CCIR for HF-communication predictions. Raytracing through such a sophisticated profile would require complex computer codes, and it remains to be tested whether the added complexity yields real improvements of predicition accuracy.

## 4.2 Global modelling of key factors

Once a model of the ionospheric layers, such as those described in the preceding section, has been chosen, a model of the temporal and spatial variations of the key factors describing the ionospheric structure must be decided upon. The diurnal, seasonal and sunspot cycle variations of peak plasma densities and heights have been modelled by CCIR (1967), based upon original publications by Jones and Gallet (1960, 1962). The model or atlas is based upon ground based measurements exclusively, and is available as a computer tape. As an example of such modelling, Figure 16 shows measured values of noon critical frequencies versus sunspot number. The sunspot cycle dependence is modelled by linear interpolation between values of critical frequencies at smoothed sunspot numbers $R_{12} = 0$ and $R_{12} = 100$. The diurnal and seasonal variations are modelled in terms of the solar zenith angle $\chi$, which, of course, is readily calculated.

The original Jones and Gallet model gave a description in terms of geographic coordinates, but it was found that inclusion of the earth's magnetic field in terms of a "modified dip coordinate" (Rawer 1963) improved the consistency of the charts. A model of the earth's magnetic field is therefore also essential in a description of the ionospheric layers. The CCIR model is also used as a basis for the International Reference Ionosphere in its global mapping of peak plasma densities and heights. As mentioned above, satellite data have not yet been included for these parameters, although the IRI uses such data to model the shape of the electron density profile above the F-layer peak. A spesific effort by CCIR to include satellite data is in progress.

### 4.3 Modelling of ionospheric absorption

We have shown that most of the ionospheric absorption of HF-waves occurs in the D-region. D-region electron densities are, however, small and difficult to measure, and the ionospheric loss is therefore normally modelled by means of empirical equations based upon absorption measurements. The absorption of the ordinary wave may be written (Budden 1966)

$$L = \text{const} \int_s \frac{\frac{1}{\mu} N_e \nu ds}{(f+f_L^2) + (\frac{\nu}{2\pi})^2} \tag{9}$$

(see section 2.3) where integration is along the ray path. For vertical incidence we may write

$$L(f) = \frac{A(f)}{(f+f_L)^2+(\frac{\nu}{2\pi})^2} \qquad \{dB\} \tag{10}$$

where A(f) is a constant when $\mu \approx 1$ (non-deviative absorption) but has a frequency dependence near the reflection point where $\mu \ll 1$. The models are based upon estimates of A(f) from measurements of absorption. The CCIR absorption equation is based upon studies by Laitinen and Haydon (1950) and Lucas and Haydon (1966), and the empirical equation for oblique incidence is

$$L(f) = \frac{677.2 \text{ I} \sec \phi_o}{(f+f_L)^{1.98}+10.2} \qquad \{dB\} \tag{11}$$

where $\phi_o$ is the angle of incidence at 100 km, and the frequencies are given in MHz

$$I = -0.4+e^{-2.937+0.8445 \, f_oE}$$

The factor I thus contains the solar zenith angle and sunspot cycle dependence through the E-layer critical frequency $f_oE$. The term A(f) is represented by an average value $\bar{A}$ = 677.2 I. An important effort to improve the absorption model has been made by George (1971) who includes a frequency dependence of A(f). This method has been extended and used for absorption modelling in the IONCAP prediction method (Lloyd et al 1981). Models of ionospheric absorption such as those mentioned above have proved useful in low and middle latitudes, but becomes inadequate in latitudes beyond about 60°, where the absorption is large and varies rapidly. An "excess system loss" factor has been included in the models to account for the latitudinal variation in high latitudes.

### 4.4 Statistical description of the ionosphere

The ionosphere has significant day-to-day variations. The data base provides input to the models in terms of monthly median values of the ionospheric parameters and a statistical distribution around the medians, for example as quartile and decile values. These statistical parameters have been included in the prediction models to provide estimates of the probability distribution of signal characteristics.

## 5. THE PRINCIPLE OF HF-PREDICTIONS

The principle of all physical predictions is, of course, extrapolation of past experience into the future. In the preceding section we have described models of the propagation medium which allows estimates to be made of the state of the medium at a certain location, solar activity as indicated by the sunspot number and solar elevation. By predicting the sunspot number at a future time, we may therefore predict the state of the ionosphere. The state of the sun is being monitored continuously from solar observations throughout the world, and reasonably reliable data are available for a period of more than 200 years. Modern observatories issue regular long-term (months, years) predictions of solar activity as measured by the sunspot number. These predictions of monthly values are based upon extrapolation of past "smoothed" sunspot numbers, allowing for the well established 11 year cycle in solar activity. Short term predictions of the development of Solar disturbances are also issued for periods of hours and days. These allow predictions of "effective" sunspot numbers which can be used in HF-prediction modelling. (See comparinion lecture).

In this section we shall not discuss solar physics and prediction of solar conditions, but rather review the procedures used to estimate the reliability of a communication circuit once the probable state of the ionosphere has been established by predictions. We shall only deal with long term predictions here.

A frequency prediction service aims at providing the user with estimates of the circuit reliability as a function of frequency and time. The useful range of frequencies is limited at the high frequency end by the path MUF, that is the highest maximum useable frequency of any possible propagation mode, and at the low frequency end (lowest useable frequency LUF) by the presence of noise and possible screening by the E-layer. The signal to noise ratio must be evaluated at each frequency within this range. The procedure may be summarized as follows:

a) Determination of the path MUF
First the points along the circuit must be found for which ionospheric information is needed. For path lengths less than 4000 km the path midpoint is used. For longer paths two "control points", 2000 km from either end of the circuit are found, and ionospheric conditions determined at these points. The "control point method" has no rigorously proved basis, but seems to work well for many applications. Given the ionospheric model, the MUF may be determined by raytracing, or by using MUF factors (M(D)). These factors are defined as:

$$\text{MUF (single hop, distance D)} = f_o \cdot M(D) \tag{12}$$

Such factors may be derived from ionograms and are tabulated for the different layers by CCIR (1967). Figure 17 shows an example of MUF factors.

The path MUF is the highest of the MUF's derived for the separate layers, and the lowest of the MUFs for the two control points.

The more sophisticated raytracing procedures such as that used in IONCAP allows determination of an area coverage, that is the area of the earth's surface illuminated via the ionosphere at each frequency. (See Figure 4).

b) Determination of signal strength
When the modes that can exist are determined, the power received at the receiver location for each mode is

$$P_r = P_t + G_R - L_b \tag{13}$$

where $P_t$ is the transmitted power, $G_t$ and $G_R$ are the gains of the transmitting and receiving antennas respectively at the elevation angle corresponding to the mode in question, and $L_b$ is the transmission loss. $L_b$ may include many terms, such as free space loss, focusing effects, ionospheric absorption (see section 4.3) and ground reflection loss. The ground reflection loss is modelled by CCIR by the conductivity of the surface at the ground reflection point. The conductivity depends upon frequency, and five reference values for fresh water, sea water, wet ground, medium dry ground and dry ground or ice have been chosen. Loss due to polarization fading, sporadic E and over-the-MUF reflections may also be added. The rms skywave field strength E (dB above 1 µV m$^{-1}$) is given in terms of $P_r$ by

$$E = P_r + 20 \log f + 107.2 \tag{14}$$

when f is the wave frequency in MHz.

c) Determination of signal to noise ratio
The noise charts and a knowledge of manmade noise conditions at the receiver site yield noise power as a function of frequency. For a given band width b of the receiver, the noise field strength $E_n$ may be determined

$$E_n = F_a - 65.5 + 20 \log f + 10 \log b \tag{15}$$

where $F_a$ is the noise figure given in CCIR report 322.

The signal to noise ratio in desibels is then SNR = $E - E_n$.

d) When a required signal-to-noise ratio is specified, the statistical properties of the ionospheric model and the noise model may be used to determine the mean reliability of the circuit for a given month.

## 5.1 Some available prediction methods

We have discussed methods for long term (months, years) predictions of HF propagation conditions, and several such methods, implemented on digital computers, have received recent international attention. It may be useful to mention some of the most important here. These are 252-2, SUP 252, IONCAP, MUFLUF, FTZ, HFM, YLE and LIL 252.

The CCIR recognized method is contained in CCIR Report 252-2 and is recommended by the CCIR for use until SUP 252, supplement to report 252-2, is completed in computerized form and tested. Both utilize a two-parabola ionospheric model and an ionospheric loss equation derived from the world data center data base.

The IONCAP program forecasts for the distribution of the signal-to-noise ratio at frequencies from 2 to 55 MHz. The model considers the E, Sporadic E, F1 and F2 layers, using an explicit electron density profile. The basic ionospheric loss equation is the CCIR 252-2 supplemented by the E layer, ES layer and over-the-MUF considerations. A separate method is included for very long distances.

The MUFLUF, FTZ (Damboldt 1975) and LIL 252 programs were submitted to the IWP (Interim Working Party) 6/12 of the CCIR for consideration of use by international HF broadcasters (CCIR, ITU, Geneva). They are considerably smaller programs than 252-2, SUP 252 and IONCAP, hence are specialized in application.

The HFM YLE is based on CCIR 252-2 with simplifications made in the field strength probability calculations.

A comparison of computer core size requirement and computational time for a sample run of 1 month, 12 hours, 6 circuits and 11 frequencies has been made on a CDC CYBER-750 computer.

| | SIZE | TIME (SECONDS) |
|---|---|---|
| 252-2 | | 27.4 |
| SUP252 | 65 K | 155.4 |
| IONCAP | 56 K | 27.4 |
| MUFLUF | 24 K | 6.5 |
| FTZ | 30 K | 0.5 |
| LIL 252 | 14 K | 3.4 |

NOTE: No direct comparison of HFM YLE was available at this writing; however comparison made by Oy Yleisradio Ab, Helsinki on a Cyper 175 Computer showed HFM YLE required 24.9 seconds of run time compared to 500.1 seconds for SUP 252, method 3.

Table 1 Comparison of some available prediction programs

The Applab III program (Bradley 1975) has been produced by the Appleton Laboratory in the UK. It is similar in size and performance to the SUP 252, but is not identical to this procedure.

The above table illustrates the difference in complexity of the different available models. Their sizes and running times are determined, essentially, by the questions they are developed to answer, and the sophistication of the methods used. The choice of method will depend upon the requirements and resources of the user.

## 6. LIMITATIONS OF THE PREDICTION METHODS

The accuracy of the predictions is limited for several reasons. Firstly the database gives inadequate coverage of many parts of the globe. Examples are the polar regions and the large oceans. This may have the consequence that important spatial structures are neglected. Secondly the ionosphere is highly variable in time, and monthly medians of propagation parameters may not always be useful to the communicator. Again this problem is particularly difficult in high latitudes. Thirdly our knowledge of the physical processes that govern ionospheric behaviour are not adequate for guiding the modelling of propagation through the medium. Nevertheless, tests show that during undisturbed conditions schemes give reasonably accurate results. The performance deteriorates markedly at latitudes above 60$^{o}$.

We have shown that the state of the ionosphere depends upon solar activity, and that most prediction methods use smoothed sunspot number as a driving function to determine the paramaters of the ionospheric layers. The sunspot number is an empirical index of solar activity, and it is not always a good measure of the effects of the sun and the interplanetary medium upon the ionosphere. Our understanding of the physical processes that determine this complex interaction is not sufficiently advanced to make reliable predictions for longer periods (years).

## 7. THE USE OF PREDICTIONS IN COMMUNICATION AND SYSTEM PLANNING

In this section we shall use the IONCAP prediction method to illustrate its application for communication and system planning purposes.

Let us first consider the problems of a user of an established circuit who needs to choose the optimum frequency from hour to hour and day to day. He or she knows the basic performance parameter of the system, such as transmitter power, antenna gains, required signal to noise ratio etc. A monthly prediction may then be issued in the simple form of a table of reliabilities as a function of frequency and time of day. Table 2 shows an example of such predictions. The path MUFs are given separately, with their corresponding reliabilities. The predictions may also be given in the form of a graph, such as in Figure 18, where MUF, LUF and FOT are plotted versus time of day. LUF and FOT (Frequence Optimum de Travail) are the lowest and highest frequencies respectively with an availability of 0.9. From such graphs the available frequency range is readily found for any time of day.

The system planner will need more information than is readily available from the simple outputs just described. He will need to know which modes are dominant at different times for different frequencies and the corresponding elevation angles of the rays, so that efficient antennas may be chosen. Propagation delay times and the relative signal strengths of different modes are also useful parameters allowing estimates of multipath interference to be made. The planner will also need to know the range of losses to be expected so that the minimum transmitter power needed to obtain a certain required reliability can be estimated. Propagation estimates must therefore be made for a range of conditions covering diurnal, seasonal and solar cycle changes.

The IONCAP will provide such users with computer outputs of the form shown in Table 3.

## 8. THE USE OF THE HF SPECTRUM

The HF-spectrum is a valuable natural resource which must be shared amongst many users with very different needs and technical capabilities. Interference from other users of the HF-band is one of the major problems in HF communications. Modern technology offers many possibilities of improving the efficiency of HF-communication systems. Some of these will be discussed in other lectures in this series. The basic principles must be to radiate the energy in the optimum direction, to radiate as little energy as possible, and to choose an efficient modulation technique in order to minimize the band width or the transmission duration. Automatic transmitter power control, antenna steering in azimuth and elevation, frequency sharing in time multiplex amongst several users, are all possible ways to go in future developments of HF-techniques. It seems obvious that improved prediction techniques will be valuable tools in improving the overall efficiency of HF-communications.

## 9. HF-COMMUNICATIONS VIA GROUND WAVE

HF-communication via ground wave is important in many areas, particularly over sea and flat land with high conductivities, where reliable circuits may be established up to distances of several hundred kilometers. The conductivity of the surface is strongly frequency dependent with rapid attenuation at the higher frequencies. In the past CCIR has published a set of curves of ground wave field strength versus distance. An example is shown in Figure 19. CCIR (1978) is in the course of implementing a computer program to estimate ground wave field strengths. Ground wave propagation may be quite complex, particularly over rough terrain and over mixed land-sea paths. There is a need for better charts of ground conductivity, and in some cases terrain modelling may be useful and important. Large topographical features such as mountain ranges and glaciers may cause reflections and strong attenuation, and vegetation, soil humidity and snow cover also influence the propagation characteristics.

## 10. CONCLUSIONS

A review has been made of the state of the art of HF propagation modelling and prediction. Although present day models have reached a high degree of sophistication and complexity there is room for major improvements in many areas. The development of the data base of ionospheric parameters has been slow, in particular in the inclusion of satellite data in the models. As a consequence there are large "white" areas in the world map of the ionosphere, i e the large oceans and the polar regions. The detailed modelling of an important parameter such as the available bandwidth of an ionospheric propagation channel has not been given much attention. This parameter is certainly of interest for the use of modern modulation techniques, such as spread spectrum modulation.

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                              METHOD 24    IONCAP 78.03

          MAR     1983            SSN =   77.
OSLO TO LONDON                                   AZIMUTHS          N. MI.      KM
59.92 N   10.75 E - 51.50 N      .08 W    220.52    31.54          622.6    1153.0
                                  MINIMUM ANGLE    3.0  DEGREES
ITS- 1 ANTENNA PACKAGE
XMTR   2.0  TO  30.0  HORZ. DIPOLE H     -.25 L     -.50 A     0.0  OFF AZ     0.0
RCVR   2.0  TO  30.0  HORZ. DIPOLE H     -.25 L     -.50 A     0.0  OFF AZ     0.0
POWER =   1.000 KW  3 MHZ NOISE = -148.0 DBW   REQ. REL = .90  REQ. SNR = 55.0

                          FREQUENCY / RELIABILITY
```

| GMT | LMT | MUF | 2.0 | 3.0 | 5.0 | 7.5 | 10.0 | 12.5 | 15.0 | 17.5 | 20.0 | 25.0 | 30.0 | MUF |
|---|---|---|---|---|---|---|---|---|---|---|---|---|---|---|
| 2.0 | 2.7 | 6.2 | .99 | .99 | .96 | .52 | .00 | .00 | .00 | .00 | .00 | .00 | .00 | .78 |
| 4.0 | 4.7 | 5.5 | .25 | .94 | .87 | .21 | .00 | .00 | .00 | .00 | .00 | .00 | .00 | .80 |
| 6.0 | 6.7 | 7.5 | .52 | .91 | .92 | .78 | .32 | .00 | .00 | .00 | .00 | .00 | .00 | .79 |
| 8.0 | 8.7 | 11.1 | .00 | .01 | .75 | .83 | .83 | .61 | .23 | .01 | .00 | .00 | .00 | .81 |
| 10.0 | 10.7 | 13.4 | .00 | .00 | .32 | .79 | .91 | .86 | .69 | .44 | .14 | .00 | .00 | .84 |
| 12.0 | 12.7 | 14.4 | .00 | .00 | .15 | .79 | .85 | .84 | .73 | .58 | .30 | .01 | .00 | .75 |
| 14.0 | 14.7 | 14.3 | .00 | .00 | .31 | .79 | .82 | .82 | .70 | .45 | .11 | .00 | .00 | .81 |
| 16.0 | 16.7 | 14.[illegible] | .00 | .13 | .70 | .79 | .74 | .74 | .67 | .38 | .07 | .00 | .00 | .80 |
| 18.0 | 18.7 | 12.3 | .48 | .76 | .83 | .81 | .75 | .70 | .59 | .37 | .12 | .00 | .00 | .79 |
| 20.0 | 20.7 | 9.2 | .93 | .97 | .95 | .90 | .74 | .49 | .16 | .01 | .00 | .00 | .00 | .88 |
| 22.0 | 22.7 | 7.2 | .97 | .98 | .94 | .83 | .57 | .22 | .03 | .00 | .00 | .00 | .00 | .85 |
| 24.0 | .7 | 6.5 | .98 | .97 | .93 | .77 | .46 | .10 | .01 | .00 | .00 | .00 | .00 | .85 |

Table 2 Reliability versus frequency and time of day for March 1983 for the circuit Oslo-London. The reliability is defined as the probability that the signal-to-noise ratio SNR exceeds the required SNR for the service. The sunspot number is SSN = 77. The minimum angle of elevation for the antenna is specified as $3^{o}$. Horizontal half wave dipoles are specified for both transmitter and receiver with heights above ground 1/4 wavelength. ($H = 0.25\ \lambda$, $L = 0.5\ \lambda$). Transmitter power is 1 kW, and the man-made noise level at the receiver site is taken as -148 dBW, corresponding to "rural" environment. The required signal-to-noise ratio, Req SNR, is the ratio in decibels of the hourly median signal power in the occupied bandwidth to the hourly median noise in a 1 Hz bandwidth. It is here specified as SNR = 55 dB, corresponding to a signal-to-noise ratio of 25 dB for a channel with 1 kHz bandwidth. The term Req Rel is used for LUF calculations.

```
                                     METHOD 23   IONCAP 78.03

           MAR     1983                 SSN =  77.
OSLO TO LONDON                                  AZIMUTHS               N. MI.      KM
59.92 N    10.75 E - 51.50 N       .08 W     220.52    31.54          622.6    1153.0
                                     MINIMUM ANGLE   3.0  DEGREES
ITS- 1 ANTENNA PACKAGE
XMTR    2.0   TO   30.0   HORZ. DIPOLE H     -.25 L     -.50 A     0.0  OFF AZ    0.0
RCVR    2.0   TO   30.0   HORZ. DIPOLE H     -.25 L     -.50 A     0.0  OFF AZ    0.0
POWER =    1.000 KW   3 MHZ NOISE = -148.0 DBW    REQ. REL = .90  REQ. SNR = 55.0

 UT  MUF

12.0 14.4  2.0  3.0  5.0  7.5 10.0 12.5 15.0 17.5 20.0 25.0 30.0 FREQ
      1F2  1 E  1 E  2F1  1F2  1F2  1F2  1F2  1F2  1F2  1F2  1F2 MODE
     29.7  5.4  6.0 40.4 31.4 25.0 23.8 29.7 29.7 29.7 29.7 29.7 ANGLE
     375.  82.  88. 262. 401. 309. 292. 375. 375. 375. 375. 375. V HITE
      69. -36. -21.  45.  65.  68.  69.  68.  59.  44.   3. -15. SNR
      12. 103.  88.  27.   5.   3.   4.  13.  22.  37.  78.  82. RPWRG
      .75  .00  .00  .15  .79  .85  .84  .73  .58  .30  .01  .00 REL

24.0  6.5  2.0  3.0  5.0  7.5 10.0 12.5 15.0 17.5 20.0 25.0 30.0 FREQ
      1F2  2F2  1F2  1F2  1F2  1F2  1F2  1F2  1F2  1F2  1F2  1F2 MODE
     34.4 45.7 25.6 27.1 34.4 34.4 34.4 34.4 34.4 34.4 34.4 34.4 ANGLE
     450. 317. 313. 337. 450. 450. 450. 450. 450. 450. 450. 450. V HITE
      70.  70.  74.  71.  68.  53.  29.   2.  -1.  -1.  -0.   0. SNR
       3.  -5.  -6.  -2.   9.  28.  52.  67.  67.  67.  66.  66. RPWRG
      .85  .98  .97  .93  .77  .46  .10  .01  .00  .00  .00  .00 REL
```

Table 3 Output option useful for system planning, showing possible ionospheric modes with corresponding elevation angles. The virtual height of the reflecting layer and the signal-to-noise ratio for each mode are given. The term RPWRG is the required combination of transmitter power and antenna gain, in dB, needed to achieve the required reliability. For explanation of other terms, see Table 2.

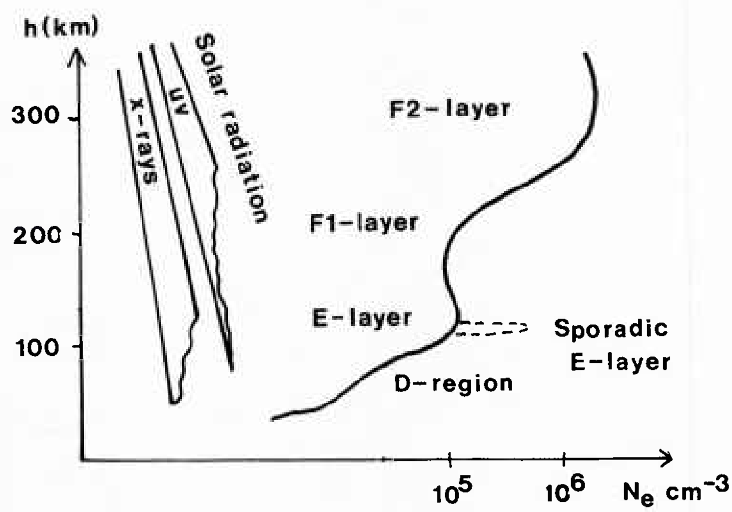



Figure 1 A typical undisturbed daytime electron density distribution. To the right the figure indicates which wavelength bands are absorbed at different heights. The E- and F-regions normally show clear maxima, whereas the F1 region is a ledge in the profile.

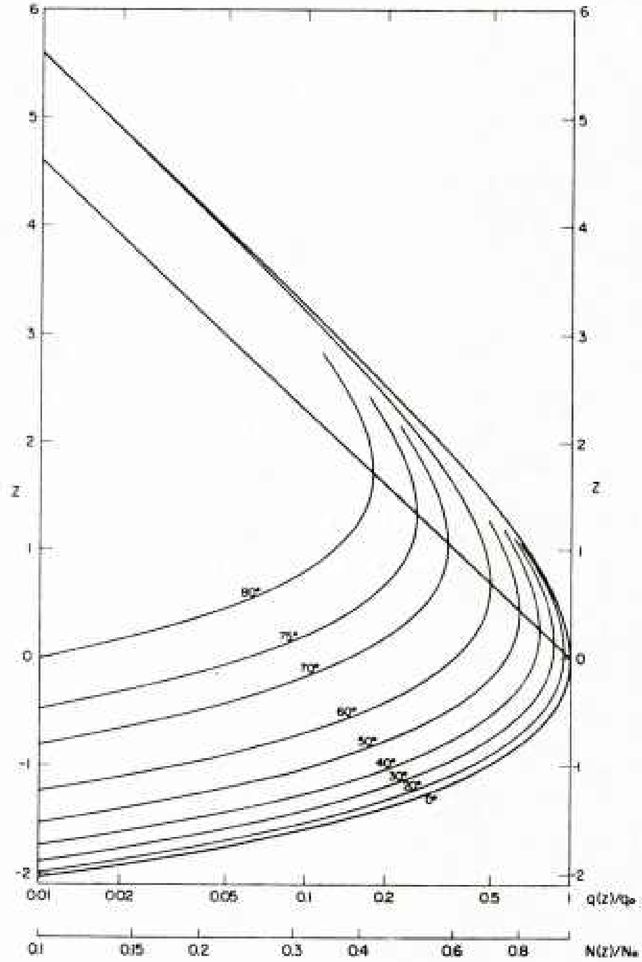



Figure 2 Normalized photoionization rate and electron density in a Chapman layer as a function of normalized height z and solar zenith angle $\chi$. $z = \frac{h-h_o}{H}$ where $h_o$ is a reference height and H is the atmospheric scale height.

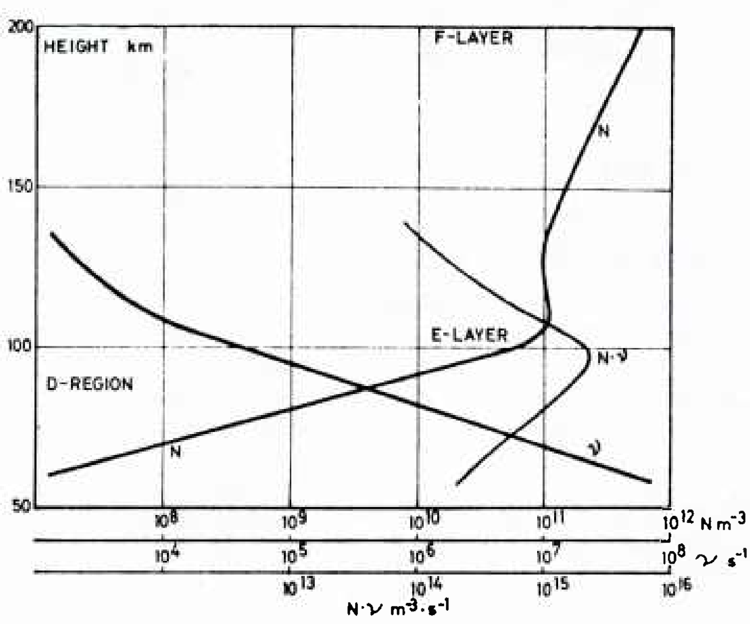



Figure 3 Typical height distributions of $N_e$, $\nu$ and $N_e\nu$ for daytime conditions.

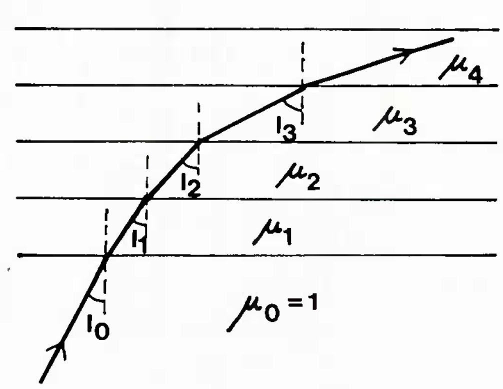



Figure 4 Refraction in a series of slabs according to Snell's law.

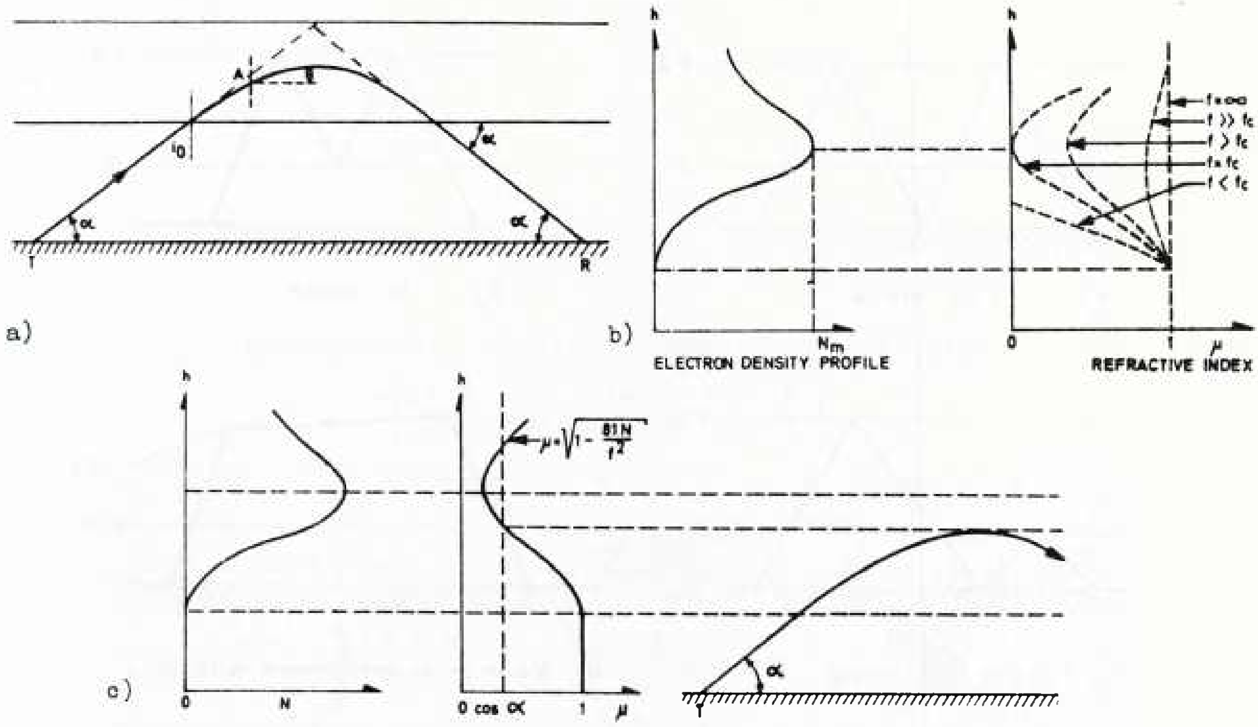



Figure 5 Illustrating a) the path of a ray through the ionosphere, b) the electron density and refractive index variation with height, c) how the ray path is determined by the refractive index variation.

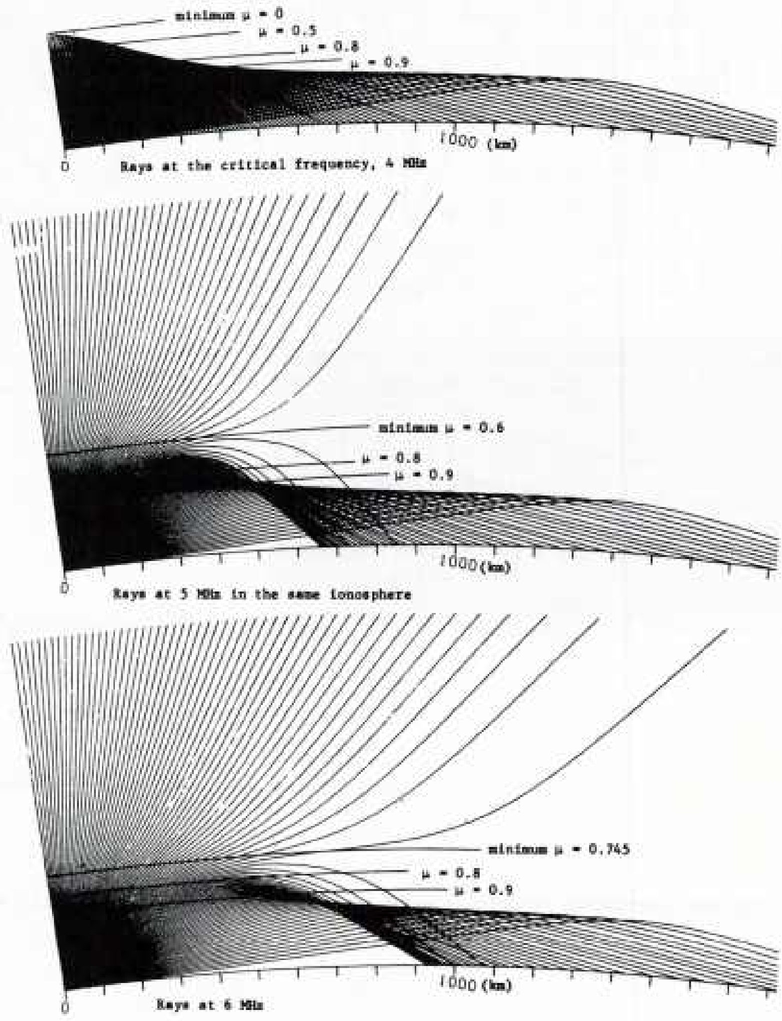



Figure 6 Raypaths for propagation at three different frequencies via a single Chapman model ionosphere of critical frequency 4 MHz, height of maximum electron concentration 300 km and atmospheric scale height 100 km (Croft 1969). Curves indicate heights of selected values of refractive index $\mu$.

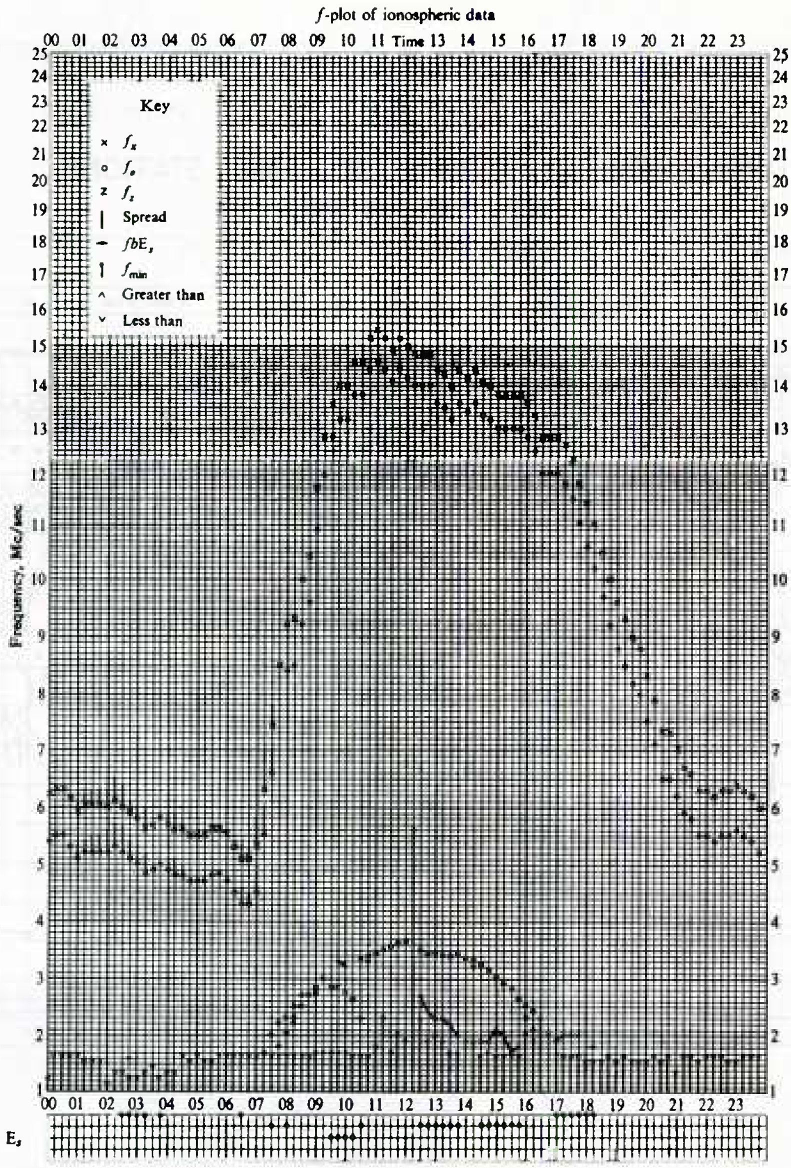



Figure 10 F plot of critical frequencies for E- and F-layer as a function of time of day (Davies 1969).

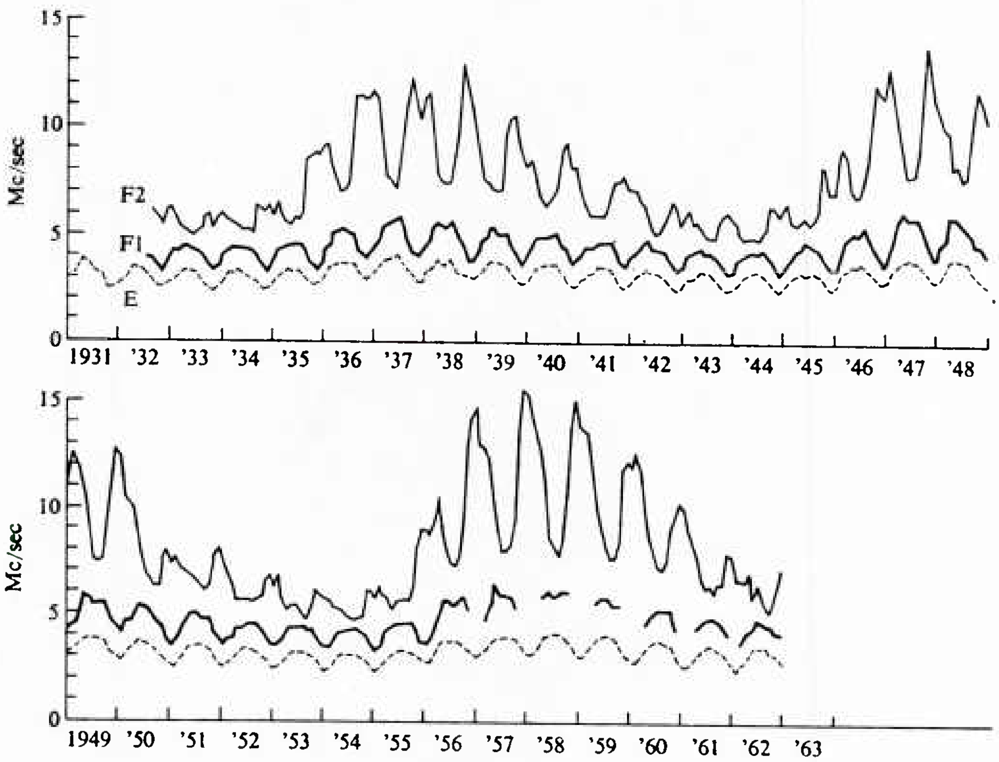



Figure 11 Long term variations of critical frequencies at Slough, England (Davies 1969).

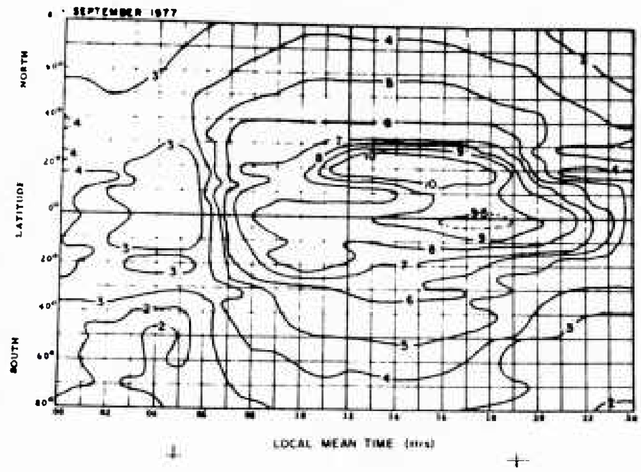



Figure 12 A contour map of foF2 in MHz for the east zone (50°E - 170°E) for September 1977 (Reddy et al 1979).

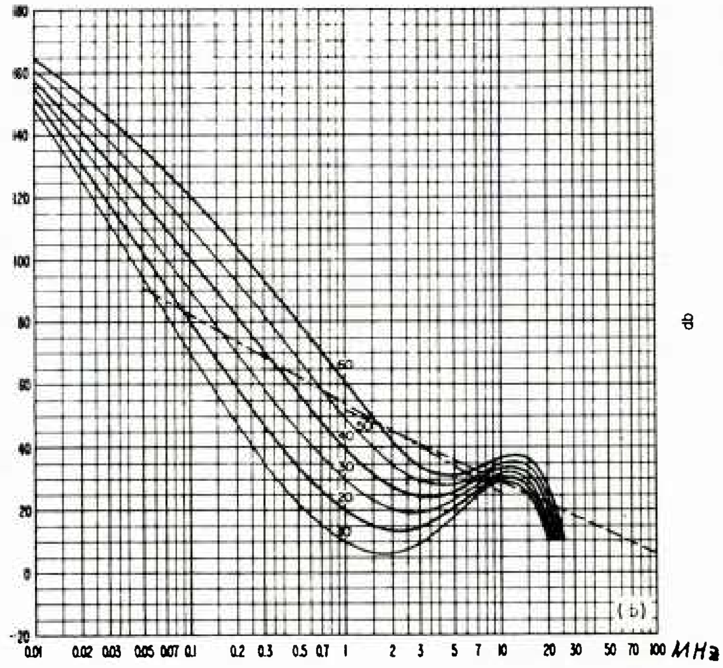



Figure 13 Variation of radio noise parameter $F_a$ with frequency. $F_a$ is the power available from a lossless antenna in dB over $kT_0$. $T_0$ is a reference temp = 288.39 K.

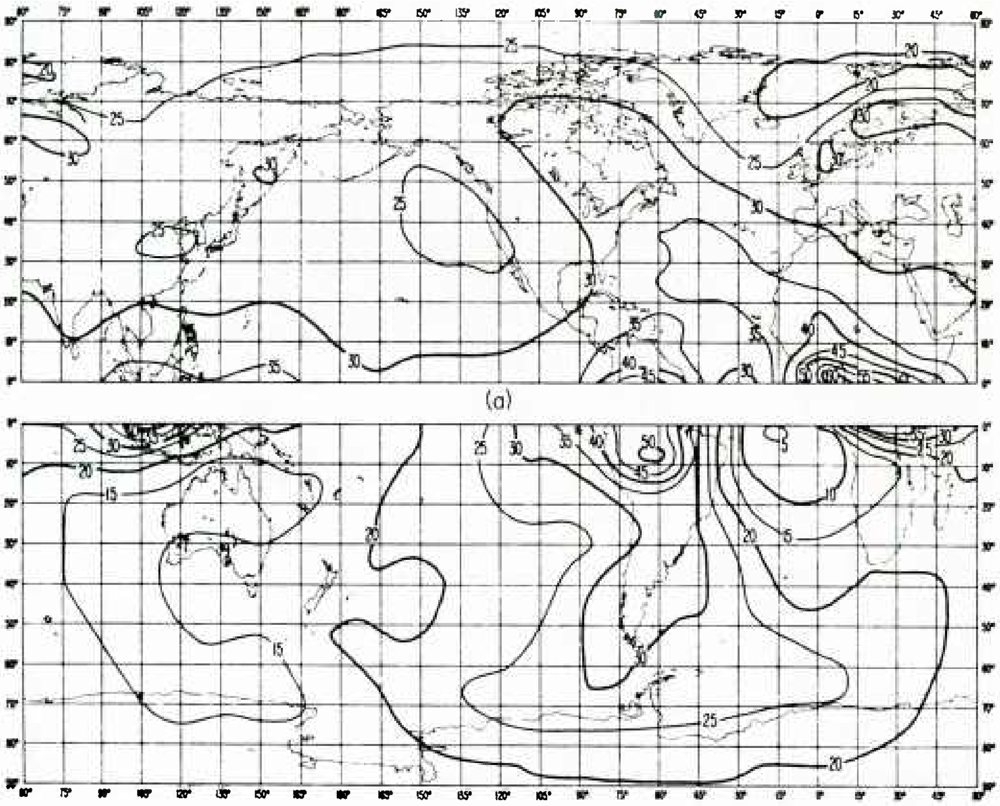



Figure 14 Contour map of $F_a$ at 1 MHz for Winter 0800-1200 Local time.

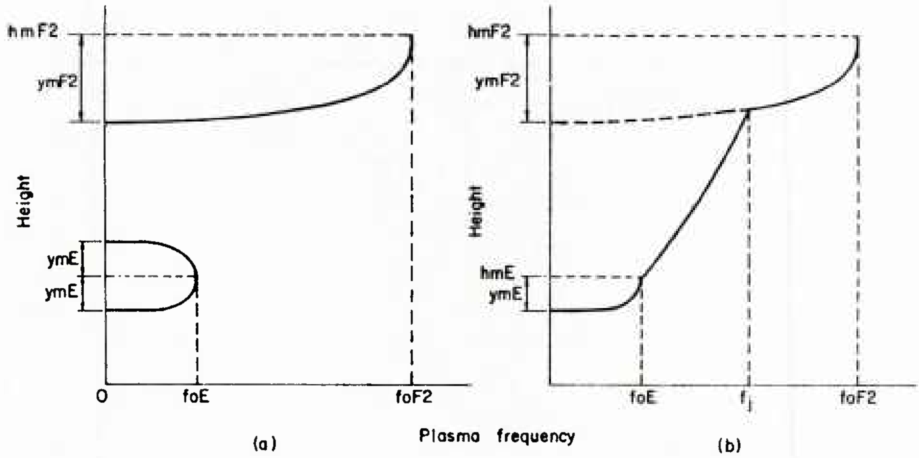
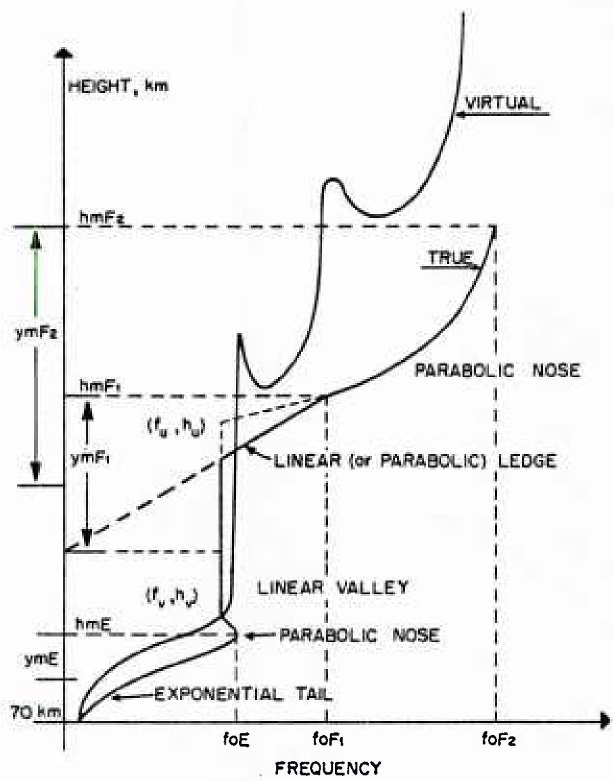



Figure 15a) $N_e(h)$ model for CCIR first procedure

Figure 15b) $N_e(h)$ model for CCIR second procedure.



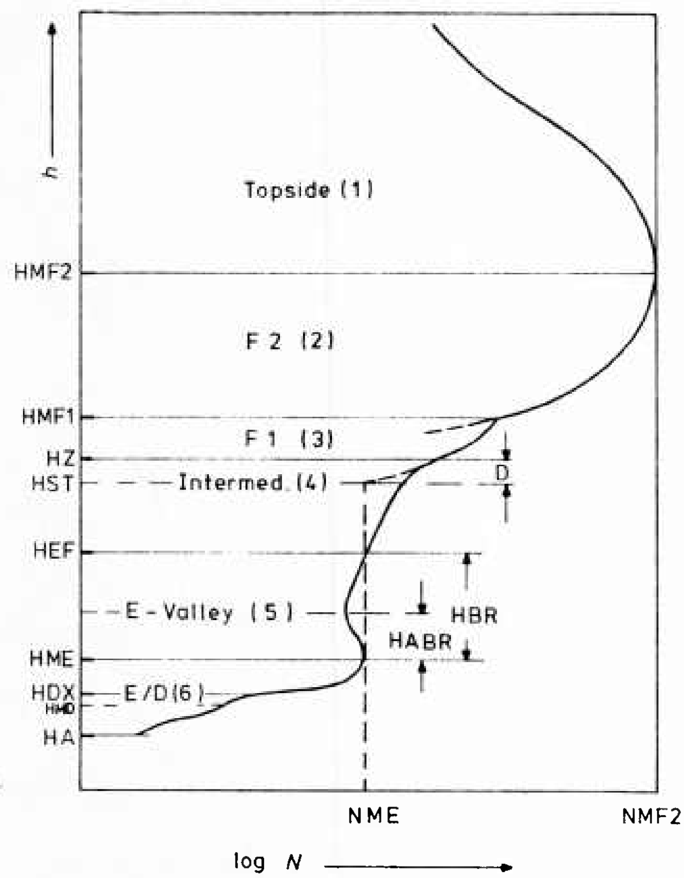




Figure 15c) $N_e(h)$ model used in the IONCAP procedure

Figure 15d) $N_e(h)$ model used in the International Reference Ionosphere.

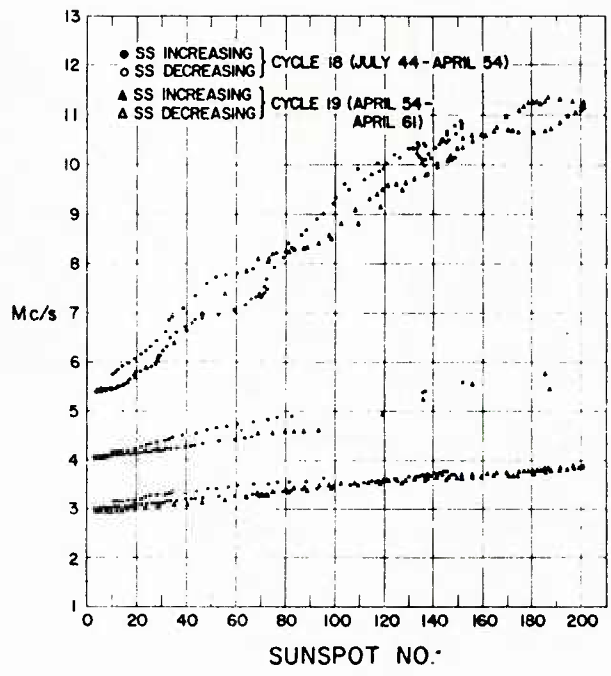



Figure 16 Critical frequencies of ionospheric layers versus smoothed sunspot number (Davies 1965).

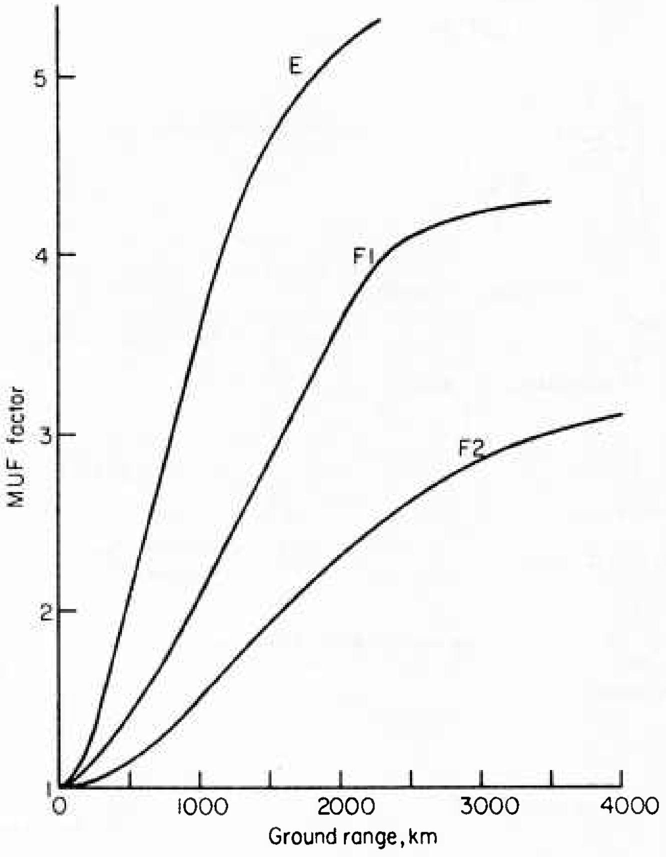
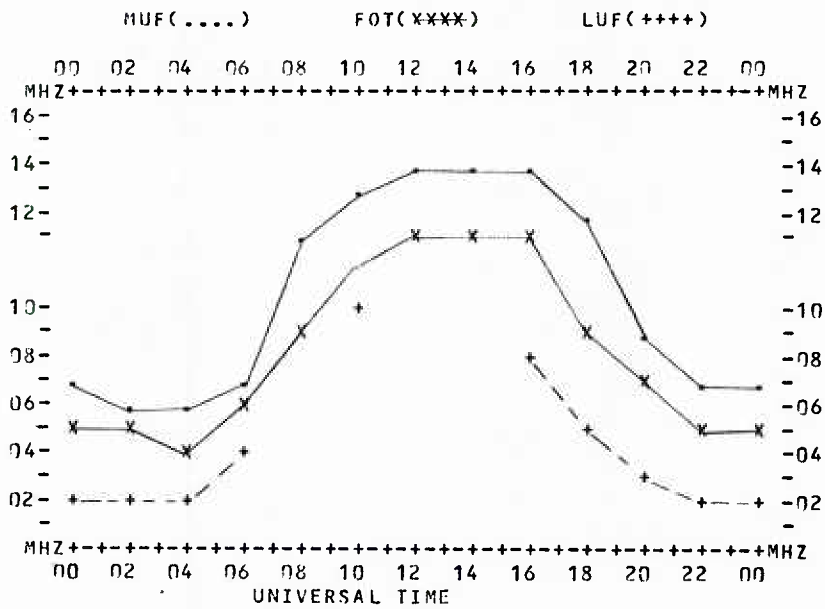



Figure 17 MUF factors for single hop E, F1 and F2 reflections (Bradley 1979).

```
                                  METHOD 28   IONCAP 78.03

           MAR      1983               SSN =   77.
OSLO TO LONDON                                   AZIMUTHS           N. MI.       KM
59.92 N    10.75 E - 51.50 N      .08 W     220.52    31.54      622.6    1153.0
                                  MINIMUM ANGLE    3.0  DEGREES
ITS- 1 ANTENNA PACKAGE
XMTR    2.0   TO   30.0   HORZ. DIPOLE H     -.25 L     -.50 A     0.0  OFF AZ     0.0
RCVR    2.0   TO   30.0   HORZ. DIPOLE H     -.25 L     -.50 A     0.0  OFF AZ     0.0
POWER =    1.000 KW   3 MHZ NOISE = -148.0 DBW    REQ. REL = .90  REQ. SNR = 55.0
```




| GMT | MUF | FOT | LUF |
|---|---|---|---|
| 2.0 | 6.2 | 4.6 | 2.0 |
| 4.0 | 5.5 | 4.1 | 2.0 |
| 6.0 | 7.5 | 6.0 | 3.6 |
| 8.0 | 11.1 | 8.9 | - |
| 10.0 | 13.4 | 10.2 | 9.6 |
| 12.0 | 14.4 | 10.8 | - |
| 14.0 | 14.3 | 11.1 | - |
| 16.0 | 14.0 | 10.9 | 7.9 |
| 18.0 | 12.3 | 9.0 | 4.8 |
| 20.0 | 9.2 | 6.7 | 2.7 |
| 22.0 | 7.2 | 5.1 | 2.1 |
| 24.0 | 6.5 | 4.6 | 2.0 |

Figure 18 MUF, LUF and FOT from IONCAP predictions for the circuit Oslo-London.

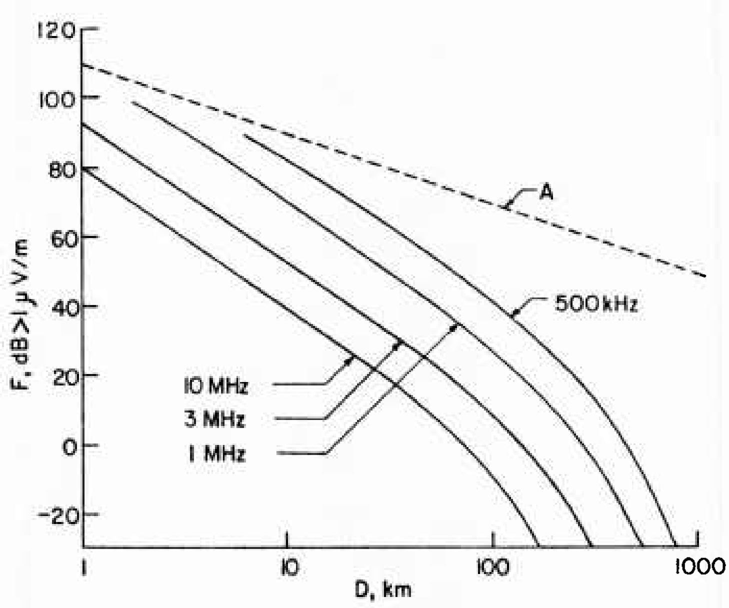



Figure 19 Ground wave field strength versus distance for $\varepsilon = 4$, and $\sigma = 10^{-3}$ 8 $m^{-1}$ (Bradley 1979).

# EMBEDDED REAL-TIME CHANNEL EVALUATION TECHNIQUES

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SUMMARY

In a previous lecture entitled "Real-time channel evaluation" presented as part of AGARD Lecture Series 127 in 1983, the basic principles and types of real-time channel evaluation (RTCE) were described. This lecture deals specifically with the class of RTCE techniques which offers most promise of economic implementation in practical HF communication systems, ie channel evaluation schemes embedded in the overall system design. Such RTCE schemes make use of essentially the same equipment as is used for communications purposes, with the requirement for additional items of equipment solely for RTCE purposes being minimised.

## 1. INTRODUCTION

In the context of HF communication, real-time channel evaluation (RTCE) has received increasing attention over the past decade because it represents an essential element in the process of HF system automation. RTCE is the name give to the process of propagation path parameter identification and modelling which is a pre-requisite for optimal system control. As described in a previous lecture series (Darnell, 1983a) there are many forms of RTCE, the majority requiring equipment additional to that of the communication system itself. From the viewpoint of economics, this has inhibited the widespread adoption of RTCE - based systems in that the RTCE equipment cost is comparable with, or exceeds, the cost of the communications equipment. This situation has led to increased work in the field of "embedded" RTCE techniques, ie techniques which can make use of the same basic RF units as are employed for passing communications traffic. The problem with this latter approach to RTCE is that it must be incorporated at the design stage of a communication system; in contrast, RTCE systems involving dedicated RF equipment can be used with communication systems which are already in service.

### 1.1 System Control via Off-Line Propagation Analysis

The classical method of controlling HF circuits involves the use of off-line propagation analysis coupled with operator experience; to this end, sophisticated propagation analysis programs have been developed to proviide the basic data required for HF system control (Haydon et al, 1976). Although such programs can predict long-term path parameters, say on a monthly median basis, with reasonable accuracy, their short-term precision is limited. Thus, they are better suited to providing data for system planning and frequency assignment where analyses over complete seasonal and sunspot cycles are required.

The aim of off-line propagation analysis procedures is to provide frequency selection data which will give the communicator a 90% probabiility of satisfactory communication at any time, assuming that the basic characteristics of the communications system, eg transmitter power, antenna types, etc, have been correctly specified in the system design. Typically, the frequency selection data would be provided in the form of an optimum working frequency (OWF) on an hour-by-hour basis. Although propagation analysis programs are being continuously refined, they have certain fundamental limitations which lead to off-line OWF predictions being adequate, rather tha optimum, for many purposes; the most important of these shortcomings are:

(a) The effects of interference from other spectrum users are not included in the analysis and prediction model;

(b) The propagation data base used for computation of predicted circuit performance is limited;

(c) The significant effects of perturbations such as sudden ionospheric disturbances (SID's), ionospheric storms and polar cap events (PCE's) cannot, by their very nature, be taken into account in the analysis:

(d) The effects of relatively transient propagation phenomena, such as sporadic E-layer refraction, can only be described approximately;

(e) The "confidence level" for the OWF predictions is normally only 90%.

To overcome some of the above limitations in long-term ionospheric forecasting techniques, short-term forecasting techniques have also been developed. These involve real-time observations of solar and ionospheric parameters, together with feedback concerning which frequencies are propagating at a given time on selected circuits. Clearly, these procedures will tend to overcome some of the data base restrictions of the long-term forecasts; however, the following points should be noted:

(a) Correction data can only be provided on the basis of sampled real-time conditions and therefore will not be uniformly accurate for all links;

(b) There are logistic and economic problems associated with the timely dissemination of the correction data;

(c) In general, the corrections do not indicate which of a set of assigned channels is likely to provide the best grade of service;

(d) As with long-term forecasting techniques, the effects of manmade interfering signals are not indicated.

For the reasons listed above, off-line propagation analysis cannot normally provide circuit parameter forecasts with a degree of confidence required for HF communication where high reliability and availability is essential. Accordingly, increasing emphasis is being given to methods for characterising HF channels accurately in real-time, ie RTCE. A brief review of RTCE techniques will now be given.

## 2. A REVIEW OF RTCE

The definition of RTCE adopted by CCIR is as follows (Darnell, 1978) (CCIR, 1981):

> "Real-time channel evaluation is the term used to describe the processes of measuring appropriate parameters of a set of communication channels in real-time and of employing the data thus obtained to describe quantitatively the states of those channels and hence their relative capabilities for passing a given class, or classes, of communication traffic".

The following comments should be noted concerning the above definition:

(a) The RTCE process is essentially one of deriving a numerical model for each individual channel in a form which can readily be employed for performance prediction and system control purposes.

(b) A particular RTCE algorithm must generate a channel model in a form appropriate to the class, or classes, of traffic which it is required to transmit over the channel. For example, a channel model required for 75 bits/s telegraphy would normally be expected to be considerably less complex than that for a 2.4 kbits/s digitised speech link.

(c) The term "real-time" implies that the measured channel parameter values are updated at intervals which are less than the overall response time of the communication system to control inputs. If measurements are made more frequently, the information cannot be employed effectively by the communication system and is thus redundant.

(d) The output of the RTCE process, in the form of an estimate of the relative capabilities of a set of channels to pass various forms of traffic, must be expressed in terms which are meaningful to the communicator and system controller, eg a predicted error rate for digital data or a level of intelligibility for analogue speech.

(e) RTCE is not simply concerned with more accurate and timely monitoring of HF propagation conditions, but also with characterising the effects of interference from other spectrum users. This is particularly important because, in many instances, it is interference which is the factor limiting communication system performance, rather than propagation.

(f) Also, as was shown previously (Darnell, 1983b), in addition to providing information on the optimum frequency for transmission, RTCE should ideally give an indication of the optimum start times and durations of transmissions.

(g) In the form defined above, RTCE will not simply select channels propagating via "conventional" ionospheric modes but will also make use of more transient modes, eg sporadic E-layer refraction and meteor burst, if appropriate.

### 2.1 RTCE Procedures: General Forms

Fig. 1 is a schematic diagram illustrating the elements of a generalised RTCE procedure with the four inputs of

(i) propagation data;

(ii) interference/noise data;

(iii) objectives of the communication system;

(iv) characteristics of the equipment comprising the communication system.

In a previous lecture (Darnell, 1983a), three basic types of RTCE procedure were defined:

(a) Those operating anywhere in the HF spectrum, without regard to frequency channel assignments, on the basis of negligible interference to other spectrum users. Pulse and chirp sounders fall into this category (Coll and Storey, 1964) (Probst, 1968) (Barry and Fenwick, 1975).

(b) Those operating within a limited set of channels assigned to a system for communications purposes (Stevens, 1968) (Wynne, 1979).

(c) Those operating within a given assigned channel of say 3kHz bandwidth (Betts and Darnell, 1975) (Darnell, 1979) (Gott and Hillam, 1979).

To date, many different forms of RTCE systems have been developed, making use of a variety of measurable parameters. Examples of specific parameters on which RTCE algorithms have been, or could be, based are:

(i) Signal amplitude;
(ii) Signal frequency;
(iii) Signal phase (absolute or differential);
(iv) Propagation time (absolute or relative);
(v) Noise or interference level;
(vi) Channel impulse response function;
(vii) Signal-to-noise or signal-to-interference ratio;
(viii) Energy distribution within the channel bandwidth;
(ix) Received digital data error rate;
(x) Received speech intelligibility level;
(xi) Telegraph distortion factor;
(xii) Rate of repeat requests in an ARQ system;
etc.

In this lecture, attention will be centred on types of channel evaluation procedure which can be embedded in an overall communication system design. These will tend to be drawn from categories (b) and (c) above.

## 3. AN HF SYSTEM DESIGN WITH EMBEDDED RTCE

In general, an embedded RTCE procedure will normally serve two purposes:

(i) To evaluate the channel currently passing traffic;

(ii) To evaluate alternative assigned channels.

The algorithms employed for (i) and (ii) will not necessarily be the same, as will be explained in subsequent sections.

Figs. 2 and 3 show two alternative configurations for an HF communication system with embedded RTCE. It is assumed that (N + 1) distinct channels are assigned to the system, with the frequency of the current traffic channel being denoted by $f_c$ and those of the alternative channels by $f_{A1}, f_{A2}, \ldots f_{AN}$. In the system of Fig. 2, two RTCE processes are incorporated: the first operates within the channel currently passing traffic, whilst the second operates over the N alternative channels (Dawson and Darnell, 1985). Again, it should be stressed that the RTCE algorithms employed for each of the two RTCE processes need not necessarily be the same.

The key to the operation of the overall system is the availability of a frequency agile transmitter. This transmitter will spend the majority of its time tuned to the traffic channel $f_c$; however, at appropriate intervals, it will be re-tuned to the frequencies of the alternative channels where it transmits relatively short RTCE probing signals prior to returning to the original traffic channel frequency. During the short intervals spent probing the alternative channels, the data on the traffic channel will be buffered until transmission is resumed. In this way, a dossier will be maintained describing the current state of the alternative channels. When the RTCE algorithm for the current channel indicates that its ability to pass traffic has fallen below, or is tending towards, some threshold of acceptability, a frequency change will automatically be initiated so that the best alternative channel becomes the traffic channel; data will again be buffered during this transition interval.

At the receivng site, two RTCE analysis units carry out the analysis of the state of the current channel and the N alternative channels respectively. The scanning RTCE receiver can continue to monitor the N channels passively in the intervals between the transmission of active probing signals, eg to assess noise/interference levels.

Clearly, for the system to operate successfully, both ends of the link must possess identical information concerning the state of the current and alternative channels at any time. This can be accomplished by means of an engineering order wire (EOW) facility, operating in the reverse sense, whose function is to pass system conrol information, rather than traffic. It is also necessary for both terminals to operate in accordance with the same time/frequency schedules for RTCE probing of alternative channels; this type of synchronism is also valuable where communications need to be re-established after a propagation outage or system close-down period.

In practice, a complete communication system would normally operate in a bi-directional mode, ie the equipment shown in Fig. 2 would be duplicated in the reverse sense. In this way, traffic and system control information can be exchanged in a multiplexed format in each direction.

The system of Fig. 3 is essentially the same as that of Fig. 2, except that a dedicated scanning transmitter is now provided specifically for the purpose of alternative channel RTCE. This auxiliary RTCE transmitter may have relatively low power rating in comparison with the traffic transmitter. The major advantage of this second type of system is that there is now no requirement for the traffic transmitter to be retuned to allow evaluation of alternative channels. Also, the alternative channels may be evaluated more frequently if it is no longer necessary to buffer the traffic signal. If the current channel and certain of the alternative channels have a relatively small frequency separation, EMC constraints may dictate that the traffic channel transmission should be inhibited during probing of those alternative channels; otherwise, there may be unwanted interaction between the output stages of the two transmitters.

Again, in a practical system, the equipment illustrated in Fig. 3 would also be replicated in the reverse sense, with EOW data multiplexed with the traffic signal.

In the following sections, options for RTCE algorithms to allow both current channel and alternative channel probing in systems of the general form shown in Figs. 2 and 3 will be described.

## 4. RTCE ALGORITHMS

In selecting an RTCE algorithm for current channel evaluation, the key considerations are:

(a) Can the necessary RTCE information be derived from the normal operating signals associated with the communications traffic?

(b) If specific RTCE signals have to be multiplexed with the normal communications traffic, what proportion of the transmitter power, transmission time, transmission bandwidth, etc will these RTCE signals occupy?

(c) To what extent can passive, rather than active RTCE techniques be employed to minimise spectral intrusion?

For alternative channel RTCE, the major requirement is to minimise the time taken to carry out the evaluation to a given level of confidence.

RTCE procedures capable of being simply embedded within an overall HF communication system design will now be described. Each of the techniques has been impelemented to at least a prototype stage.

### 4.1 Channel Evaluation & Calling (CHEC) System

The CHEC system was developed in Canada to improve the reliability of communication between long-range maritime patrol aircraft and ground stations, with the emphasis being placed upon the air-to-ground link (Stevens, 1968). CHEC was designed for a situation where one or more mobiles are required to pass traffic to a base station. On each of the m assigned channels, where m would normally be < 20, the CHEC base transmitter radiates in sequence a probing signal of 5 seconds duration comprising a selective calling code, data on the average noise level at the base station in that channel, together with a CW section, as shown in Fig. 4. At the remote receiver alerted by the selective calling code, the base station average noise levels

$$\overline{n(t)}\Big|_{f_i} \qquad 1 \leqslant i \leqslant m \qquad [1]$$

for the subset of k channels actually propagating to the mobile are decoded to give

$$\overline{n(t)}\Big|_{f_j} \qquad \text{where } j \text{ can take any } k \text{ distinct values in the range 1 to } m \text{ and } k \leqslant m \qquad [2]$$

The subset of corresponding average received signal levels at the mobile

$$\overline{A(t)}\Big|_{f_j} \qquad [3]$$

are evaluated using the CW sections of the base transmissions. Thus, by assuming propagation reciprocity and also making allowance for differences in antenna gains and transmitter powers between base and mobile, a processor at the mobile computes a predicted average signal-to-noise ratio for its own transmissions propagating to the base in each of the k channels. Therefore

$$\mathrm{SNR}(t)\ \mathrm{base}\ \Big|_{f_j} = \left[ G \cdot \frac{\overline{A(t)}}{\overline{n(t)}} \right] \Big|_{f_j} \qquad [4]$$

where G is a channel dependent factor to compensate for the differences in antenna and transmitter characteristics between base and mobile. The optimum channel for mobile-to-base communication is then given by the value of j for which the SNR is a maximum. In experimental form, CHEC was shown to give significant improvements in channel availability and reliability.

Other systems, similar in concept to CHEC but applicable to different operational requirements, have been devised, eg a ship-shore system employed in the UK (Wynne, 1979) and the Canadian radio telephone with automatic channel evaluation (RACE) system (Chow et al, 1981).

A more modern system, similar in nature to CHEC, is termed the "advanced link quality analyser" (ALQA) (Bliss, 1985). ALQA carries out a multi-parameter analysis of a set of channels, based upon the following time averages:

(a) Signal-to-noise density ratio;
(b) RMS multipath time-delay spread;
(c) Frequency spread;
(d) Fading rate;
(e) Fading depth;
(f) Fading power spectrum;
(g) Noise/interference time, frequency and amplitude characteristics.

The overall channel evaluation is based upon an appropriately weighted combination of parameter measurements.

### 4.2 RTCE by Pilot Tone Phase Measurements

In this method, the RTCE probing signal is a simple CW pilot tone inserted at a suitable position in the current channel bandwidth (Betts & Darnell, 1975). The basis of the evaluation procedure is that, after detection of the pilot tone in a narrow bandpass filter at the receiver, its phase variations are analysed and used to infer the suitability of the channel for the transmission of various types of traffic by making use of analytical relationships between phase instability and data error rate.

In the experimental system, the phase of the received pilot tone is compared with that of a locally-generated reference phase source. This phase difference is sampled at regular intervals, typically 10 ms, and the phase difference at the current sampling instant, $\theta_n$, compared with the phase difference measured and stored at the previous sampling instant, $\theta_{n-1}$. Ideally, this phase difference should be zero but for practical channels will normally be non-zero; if the difference in phase between the two samples exceeds a certain preset threshold value, $\theta_t$, a "phase error" is counted, ie

$$\left| \theta_n - \theta_{n-1} \right| > \theta_t \qquad [5]$$

for a phase error.

Clearly, it is necessary that the sampling interval should be an integral multiple of the pilot tone period in order that sampling takes place at the same point in the pilot tone cycle under ideal conditions.

The parameter selected to indicate the state of the channel is the number of phase errors occurring in a predetermined measurement interval, typically 100 to 200 seconds. For practical tests of the system, a low-level pilot tone was frequency-multiplexed with a 2-tone, frequency-exchange keyed (FEK) 50 bit/s binary data signal, as illustrated in Fig. 5. By appropriate calibration, the number of pilot tone phase errors can be related to the number of data bit errors over the same measurement interval. The theoretical relationships for steady signal, flat fading and frequency-selective fading are shown as solid lines in Fig. 6. The points superimposed upon these theoretical plots represent measured values and indicate the typical scatter obtained during an experimental run.

The main conclusion which could be drawn from a comprehensive series of tests was that, for the great majority of channel conditions encountered, the data error rate which would be experienced using a given transmission scheme over a particular path could be predicted with reasonable accuracy via simple phase measurements on low-level CW pilot tones. Thus, this latter parameter could be used directly to establish a quantitative measure of channel acceptability.

Pilot tone RTCE could be used in a scenario where a mobile requires to communicate with a base station; pilot tones could be radiated by a single wideband base station transmitter at low level (typically a few watts) simultaneously on all channels assigned for mobile-to-base transmission which were clear of interference at the base. Hence, as shown in Fig. 7, the mobile would be able to make phase error rate measurements on all channels propagating to it from the base: in the same way as for CHEC, propagation reciprocity would be assumed and allowance made for the different transmitter and antenna characteristics at the two sites in order to predict the channel likely to yield the maximum SNR at the base. The disadvantage of the long evaluation time for the pilot tone method could be offset in some situations by its extreme simplicity of implementation.

### 4.3 RTCE by Error Counting

A simple form of RTCE is to probe the m channels to be evaluated by means of a test signal having essentially the same format as the traffic signal to be passed over the path. It is convenient practically if the RTCE signal is digital so that errors can be counted, rather than having to make more subjective assessments of quantities such as speech intelligibility. The essential requirement is again one of providing transmission and reception systems synchronised in both time and frequency, although the accuracy of synchronisation neccessary is somewhat less than that for say an ionospheric sounding system.

For digital traffic, the assigned channels can be evaluated in sequence using exactly the same modulation format as employed by the traffic transmission and the corresponding error rate measured at the receiver; the channel acceptability is then related directly to the measured error rates. This procedure is equally valid for data or digitised speech traffic since the error rate for the latter can also be interpreted in terms of speech intelligibility - as shown by the empirical model for 1.2 kbits/s digitised speech given in Fig. 8.

A similar, but less precise, relationship exists between analogue speech intelligibility and data error rate. Fig. 9 shows a baseband spectrum in which LINCOMPEX-processed speech (Awcock, 1968) is frequency multiplexed with low-rate binary FEK telegraphy. Fig. 10 is an empirical model showing the relationship between LINCOMPEX speech intelligibility and FEK data error rate for representative HF paths; again, the speech quality could be predicted with reasonable accuracy from error rate measurements on a simple digital RTCE signal.

Therefore, for all common forms of HF traffic, it appears feasible to carry out RTCE via a simple error counting procedure. The main disadvantage of the method is the time taken to accumulate the necessary error count in the case of low-rate data.

Practical trials have been carried out using a basic error counting RTCE system (Darnell, 1978). Two types of traffic signal were used:

(a) 75 bits/s FEK telegraphy;

(b) 1.2 kbits/s digital data.

Path lengths of 700 km and 1100 km were used in the tests, with 24-hour operation.

The classical method of controlling an HF circuit using off-line propagation analysis data is to select one daytime operating frequency and one night-time frequency, ie 2-frequency working as illustrated in Fig. 11. The RTCE error counting trials compared the circuit availability using this form of 2-frequency working with that obtained by employing the RTCE data for frequency selection. On average, it was found that the use of RTCE increased the circuit availability by approximately 45%. It was evident from the results that the factor limiting circuit performance was, in most cases, manmade interference. The value of the RTCE process lay chiefly in its ability to enable the communicator to avoid interfering signals, rather than to track propagation changes.

## 5. RTCE DERIVED FROM THE NORMAL OPERATING SIGNALS OF A COMMUNICATION SYSTEM

All the RTCE techniques described in the previous section have been applicable to situations in which the communicator has available for his use a number of assigned channels, ie for alternative channel evaluation. In general, the active probing techniques, with the possible exception of pilot tone RTCE, are somewhat inconvenient for current channel assessment. Here, it is advantageous if the RTCE information can be derived from the normal operating signals of a communication system.

In addition to assessing the state of the current channel relative to the alternative channels, current channel RTCE may be required

(a) To examine the baseband spectrum of a channel to determine where, within that baseband, a narrowband traffic signal should be placed for minimum error rate;

(b) To determine the optimum signal processing procedures to be applied to a traffic transmission within the channel by making use of an appropriate RTCE signal.

Techniques for current channel RTCE from normal operating signals will now be described.

## 5.1 In Band RTCE

The term "in-band RTCE" refers to a technique designed specifically for the evaluation of sub-channels within a nominal 3 kHz assigned channel bandwidth. At the receiving site, a real-time spectrum analyser monitors the distribution of noise/interference energy for all sub-channels within the bandwidth using a set of bandpass filters. Low-energy regions are identified and indicated to the transmitter site by means of a low-rate EOW; this allows a narrowband traffic spectrum (< 3 kHz) to be adjusted so that the majority of its energy falls in the low noise sub-channels (Darnell, 1979). Studies of narrowband HF interference have indicated that its characteristics can only be expected to be relatively static for periods of a few minutes (Gott & Hillam, 1979); thus it may be necessary to make frequency changes relatively often. If in-band RTCE can be used to select different parts of an assigned channel as the narrowband interference patterns change, the need to change the frequency of the transmitter and receiver is avoided, thus improving the efficiency of spectrum utilisation.

In an FSK/FEK system, it is possible to monitor the noise/interference level in a given tone sub-channel during the intervals when the other tone(s) are keyed on. Thus, a running assessment of the SNR in the current sub-channels can also be made in addition to monitoring the state of the alternative sub-channel (Humphrey and Shearman, 1985).

## 5.2 RTCE Using Soft Decision Information

The term "soft-decision" relates to the confidence level associated with a "hard" digital decision. For example, soft decision information could be obtained from:

(a) Amplitude values of a received signal;

(b) Phase margin between a phase reference and phase detected by a receiver.

Any information which can be extracted from a received signal and subsequently used to quantify a detection decision confidence level can, in principle, be used for RTCE purposes.

In a DPSK modem such as KINEPLEX (Mosier & Clabaugh, 1958), the phase margin between the received signal phase, $\theta_r(t)$, and the locally generated reference phase, $\theta_o(t)$, could be used as the basis of the evaluation ie

$$\text{RTCE Assessment} = F\{ \left| \theta_r(t) - \theta_o(t) \right| \} \qquad [6]$$

Alternatively, the evaluation could be a function of both the amplitude of the received signal and its phase margin. Soft-decision data of this type has been incorporated into an HF modem known as CODEM (Chase, 1973) to enhance transmission reliability and to enable the modem to reject data blocks not meeting the required confidence criteria.

## 5.3 RTCE in ARQ Systems

An ARQ communication system typically formats the data to be transmitted into fixed-length blocks which are then individually labelled. These blocks are transmitted sequentially until control data derived from soft-decision processing or error control decoding indicates that a given block has been decoded erroneously at the receiver. An ARQ signal is then passed to the transmitter site via a feedback link, or EOW, requesting a repeat transmission of the corrupted block. Evidently, the number of block repeats requested in a given time interval will be a measure of channel quality and can be used for RTCE purposes.

If the fixed-length data blocks are encoded using an error control code, it is possible to vary the power of the code in response to the rate of ARQ requests, ie "variable redundancy coding" (Goodman and Farrell, 1975). Thus, as the BER of the forward path increases, so new codes are selected at predetermined BER threshold values. In this way, the throughput is adjusted to match the prevailing channel conditions, whilst maintaining a certain maximum BER at the output of the ARQ system (Hellen, 1985). Fig. 12 illustrates a typical throughput characteristic for such a system.

## 5.4 RTCE by Traffic Signal Modification

In some situations, it may be impossible to obtain the required RTCE data directly from the traffic signal, possibly because the soft-decision parameters are not accessible or as a result of the traffic being encrypted. In the latter case, it is possible to modify the format of the traffic by the introduction of additional signal generation and processing functions which will facilitate the extraction of RTCE data.

Possibly the simplest method of accomplishing the necessary modification of the traffic signal would be to insert an auxiliary, low-level pilot tone at a suitable null in the baseband spectrum of the traffic signal. Analysis of the pilot tone phase error rate, as described in Section 4.2, would then allow the data error rate for the traffic channel to be estimated with reasonable precision.

In certain forms of encrypted data transmission systems, a special error detection and correction (EDC) process can be introduced in order to yield RTCE information to assist in overall system control; Fig. 13 shows such an arrangement. It is assumed that security considerations limit access to the elements of the communication system except for the region shown. If an auxiliary EDC system, shown hatched, is introduced into this region, it can be used to format the encrypted traffic into arbitrary codewords prior to transmission. At the receiver, the received codewords will be decoded to yield the original encrypted traffic stream; however, the EDC algorithm can be implemented in such a way that the number of errors being detected and corrected can be continuously monitored, thus indicating the state of the channel for RTCE purposes.

5.5 RTCE by Assessment of Synchronisation Quality

In a communication system requiring synchronisation at the start of each message, it is possible to design a synchronisation preamble in such a way that RTCE information can be extracted from it.

Figs. 14(a) - (e) illustrate the performance of a synchronisation preamble comprising a 77-bit code formed from 7-bit and 11-bit Barker sequences (Barker, 1953) in Gaussian white noise; the traces show the output of a detection filter matched to the synchronisation sequence for the error rates shown. It is seen that the peak-to-sidelobe ratio of the preamble deteriorates as the BER increases. Thus, the value of peak-to-sidelobe ratio can be related to channel quality.

5.6 RTCE by Pseudo-Error Counting

A technique termed "pseudo-error" counting has been proposed to overcome the problem the excessive time required for error rate monitoring at low BER values (Leon, 1973); here the error rate is artificially amplifed by th use of an over-sensitive detection method so that the rate measured by the RTCE system is substantially greater than that which would be experienced by the traffic transmission, thus allowing the required error count to be accumulated more rapidly. Because of the inherently high error rates associated with HF links, and also the rapidly time-varying nature of the received signal, it may well be difficult to apply pseudo-error counting to HF links due to inaccuracy of calibration.

As an alternative to this, current channel RTCE might be accomplished by the insertion of short intervals of a modulation technique requiring a higher SNR for successful reception than that actually used for data transmission; for example short bursts of 1.2 kbits/s serial DPSK might be inserted in a low-rate FSK data modulation stream.

Another option for increasing the sensitivity of the error counting process is to use the concept of a "soft" error. Fig. 15 illustrates the principle: a "soft error band" is defined as shown, and if the average demodulation output during any symbol interval is within this band, a soft error is registered. By widening the soft error band, the number of errors counted over any given time period can be increased.

5.7 Noise and Interference Characterisation

In previous sections of this lecture, the importance of noise and manmade interference in determining HF communication system performance has been stressed. In areas of high spectral congestion, eg the central region of Europe, it is normally manmade interference which limits system performance, rather than propagation, which is relatively predictable. Simiarly, with the off-line propagation analysis programs discussed in Section 1, one of their major limitations stems from the lack of an adequate model for interference.

It is clear, therefore, that considerable effort must be put into the measurement and characterisation of interference, both from the point of view of RTCE and of off-line analysis. In-band RTCE systems of the type discussed in Section 5.1 could form the basis of interference assessment systems for incorporation into RTCE procedures. Other systems can also be used (Cottrell, 1979) (Barry & Fenwick, 1975) (Gott et al, 1983).

In some cases, the interference assessment will be explicit; in others, such as the error counting technique, the RTCE process evaluates the combined effects of both propagation and interference in a single measurement process

6. CONCLUDING REMARKS

This lecture has attempted to describe the manner in which embedded RTCE techniques can be integrated into an HF communication system, and also to indicate, by means of representative examples, the range of those techniques currently available.

The value of employing some form of RTCE to enhance the performance of HF communications is well established: however, the progression must now be to incorporate such techniques into the design of a communication system at the earliest possible stage, rather than treating them as an optional extra once the system is in service. It is only by embedding the RTCE in the system and using predominantly the same RF units as are used for communications purposes, that economic implementation can be achieved.

In essence, all RTCE techniques try to predict the system performance by direct or indirect measurements of effective SNR. With the advent of various forms of programmable signal processor, it has become feasible to implement channel encoding/decoding (modulation and error control coding) functions via software. This development allows greater flexibility in the extraction of RTCE data from the normal operating signals of the communication system in that parameters such as amplitude and phase margins can be made readily accessible for continuous monitoring. It also becomes feasible to adapt the design of say a modem to facilitate the availability of RTCE information.

The use of what might be termed "co-lateral" RTCE data should not be neglected, if it can be incorporated into the communications control procedures in a systematic way. In this category, for example, could be placed the monitoring of known transmissions situated at fixed points on the earth's surface to determine the general quality of ionospheric propagation.

In the context of the overall HF communication system, RTCE data could, in principle, be employed as a source of control information to assist in the adaptation of the following parameters:

- Transmitter power level;
- Frequency of operation;
- Bandwidth;
- Information rate;
- EDC algorithm;
- Modulation type and spectral format;
- Start time and duration of transmission;
- Antenna characteristics, eg null positions;
- Diversity combining algorithm

  etc.

The potential advantages to the HF communicator arising from the use of RTCE can be summarised as:

(a) Off-line propagation analysis requirements can be eliminated for operational purposes; however, this form of analysis will still be valuable for system planning purposes.

(b) The effects of manmade interference can be measured and specified quantitatively, thus eliminating the major cause of operational uncertainty.

(c) Relatively transient propagation modes, such as sporadic E layer refraction, can be identified and used for high quality communication; the presence of these modes can increase the available spectrum by as much as 2 or 3 times.

(d) RTCE facilitates the selection of channels higher in frequency than would have been suggested by off-line propagation analysis, hence reducing spectrum congestion.

(e) RTCE provides a means of automatically selecting an optimum transmission channel and of ranking stand-by channels in order of preference - an essential pre-requisite for an automatic HF system.

(f) Radiated power and spectral occupancy can be minimised, consistent with meeting a received signal fidelity criterion, thus reducing spectral pollution.

(g) RTCE provides the basic data required for adaptation of communication system parameters other than frequency, eg signal processing algorithms, antenna characteristics, etc.

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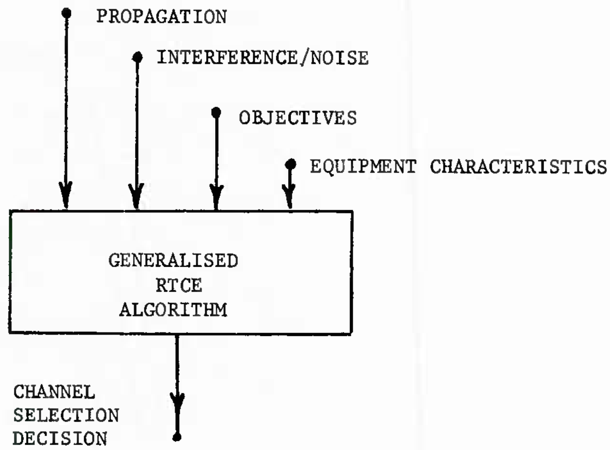


FIG 1 INPUTS REQUIRED FOR GENERALISED RTCE PROCEDURE

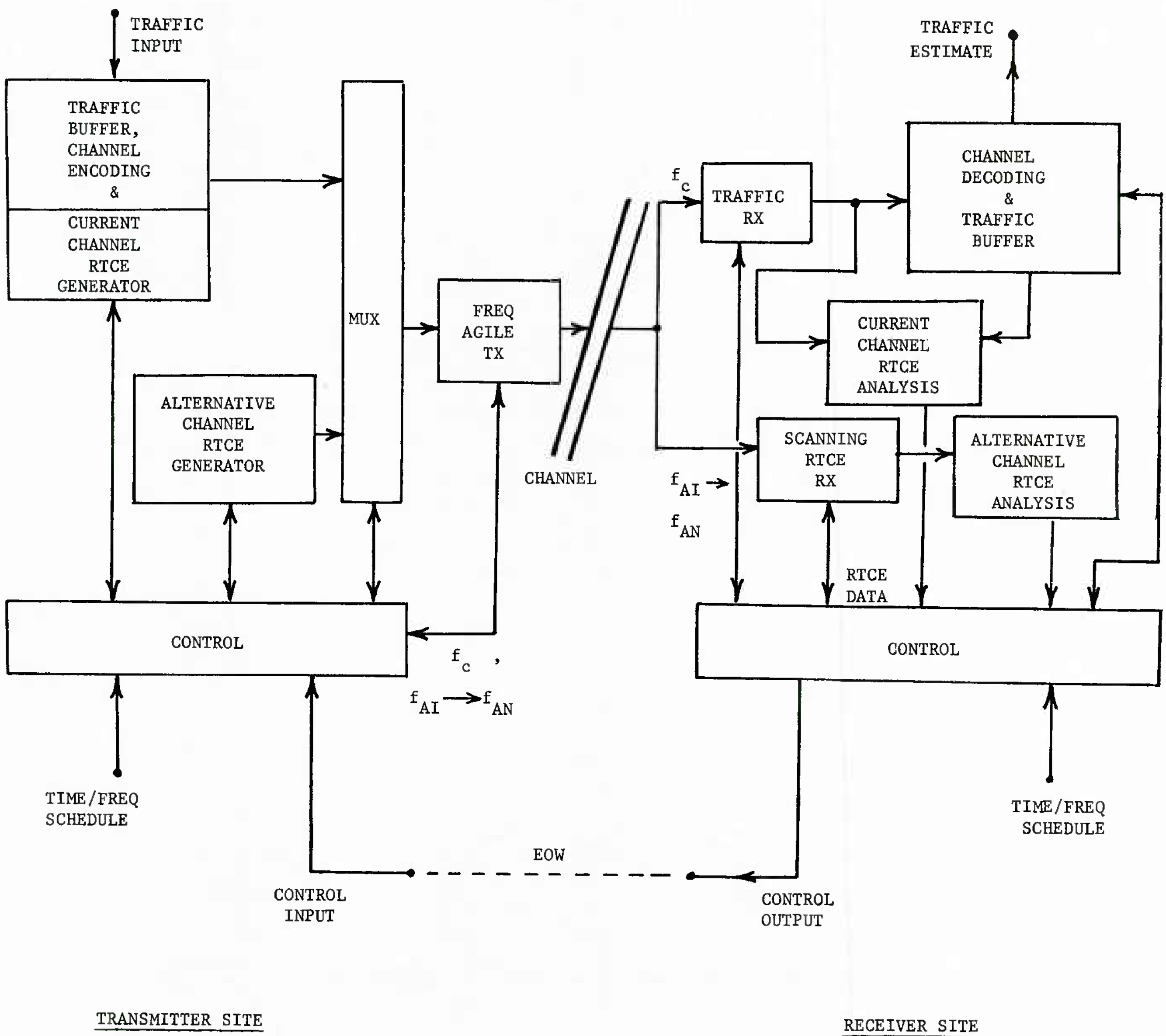


FIG 2 HF COMMUNICATION SYSTEM WITH EMBEDDED RTCE - TYPE 1

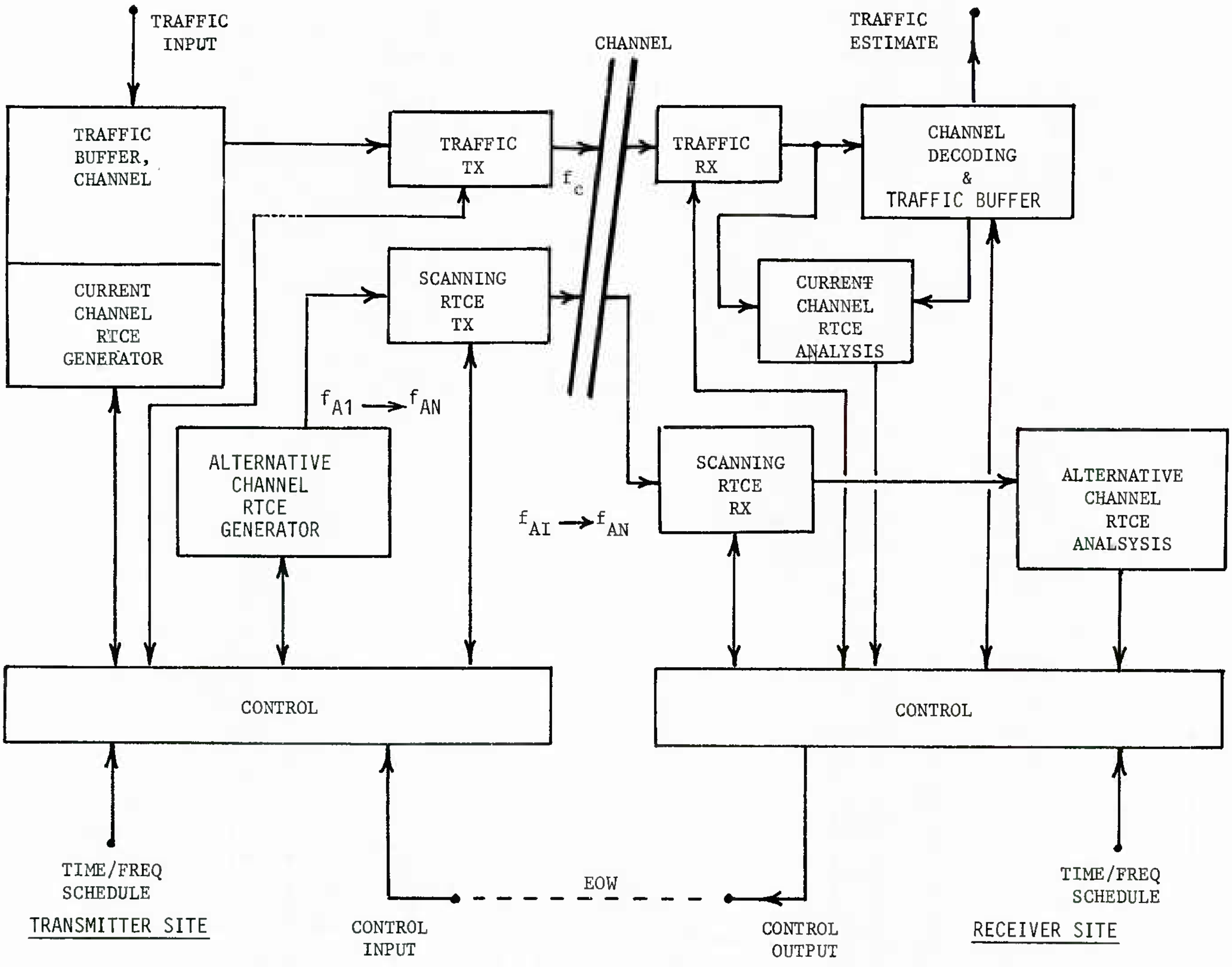


FIG. 3 HF COMMUNICATION SYSTEM WITH EMBEDDED RTCE - TYPE 2

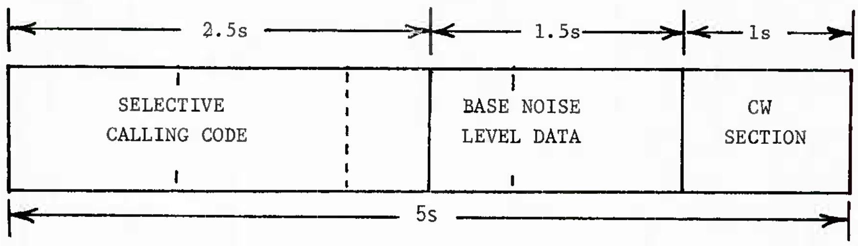


FIG 4 CHEC SIGNAL FORMAT

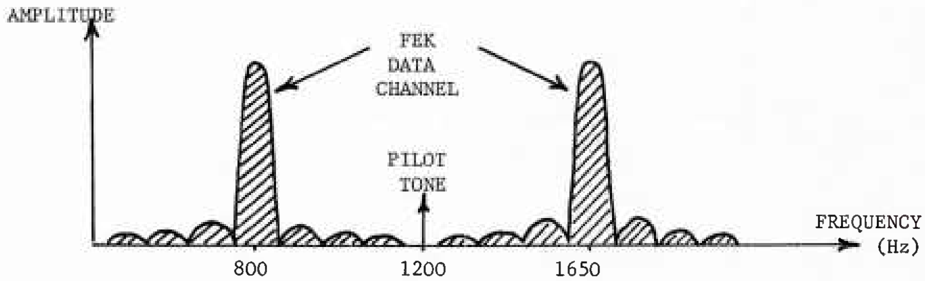


FIG 5 BASEBAND SPECTRUM OF FEK DATA CHANNEL WITH PILOT TONE INSERTED

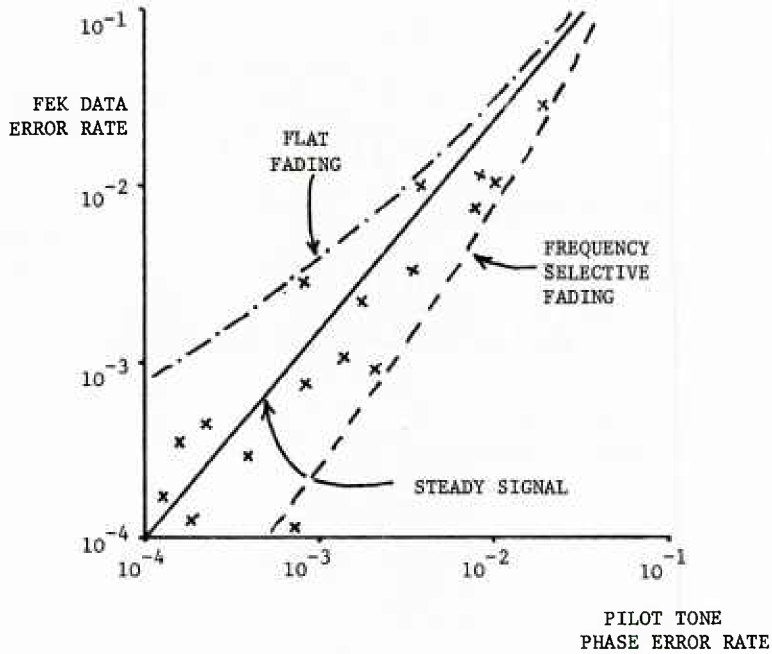


FIG 6 THEORETICAL AND EMPIRICAL RELATIONSHIPS BETWEEN PILOT TONE PHASE ERROR RATE AND FEK DATA ERROR RATE

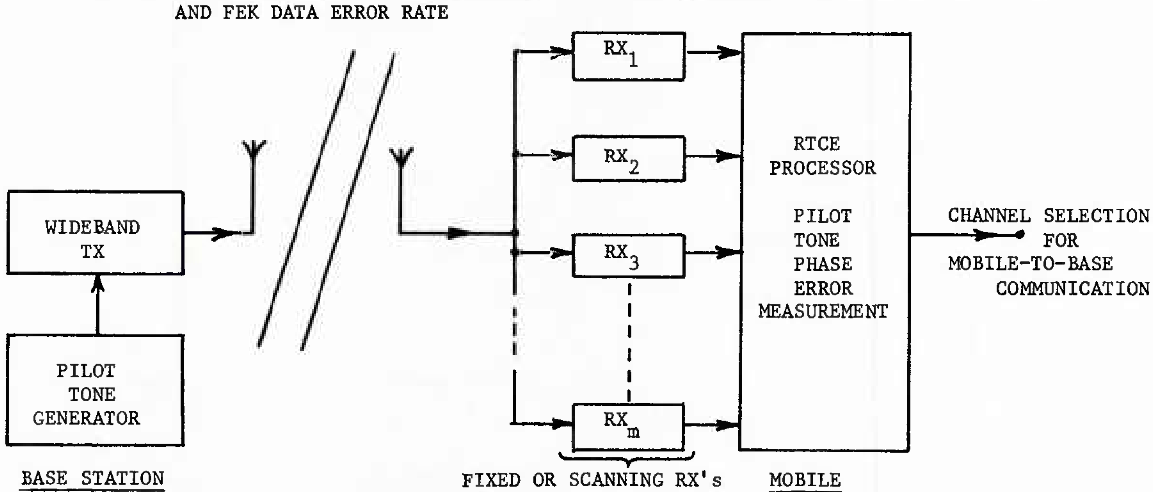


FIG 7 APPLICATION OF PILOT TONE RTCE

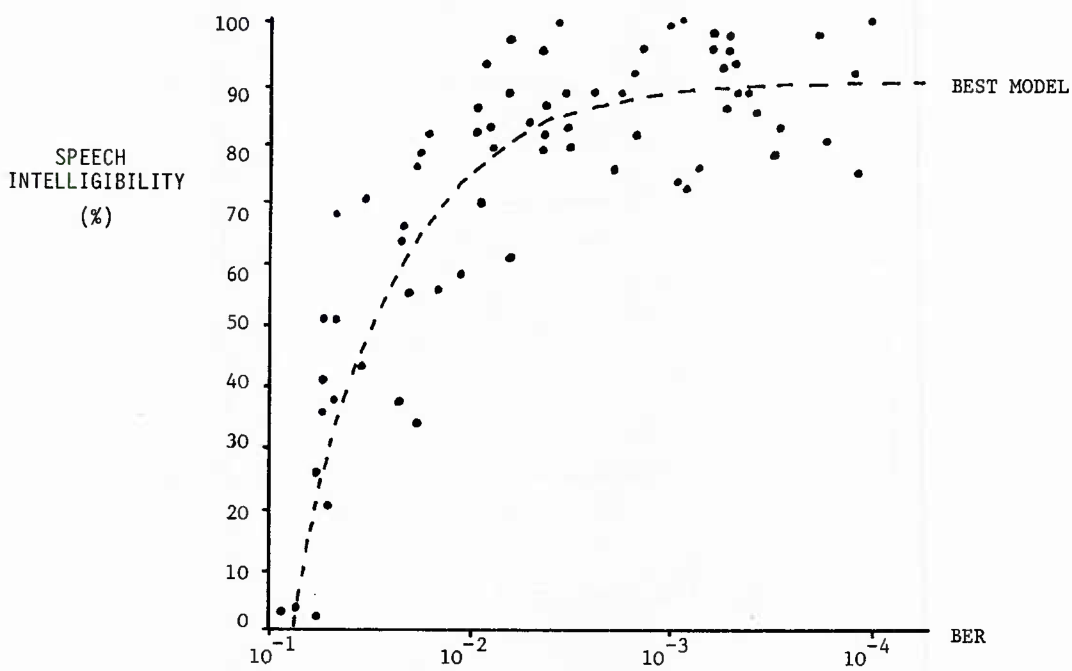


FIG 8 EMPIRICAL RELATIONSHIP BETWEEN DIGITISED SPEECH INTELLIGIBILITY & BER AT 1200 BITS/S

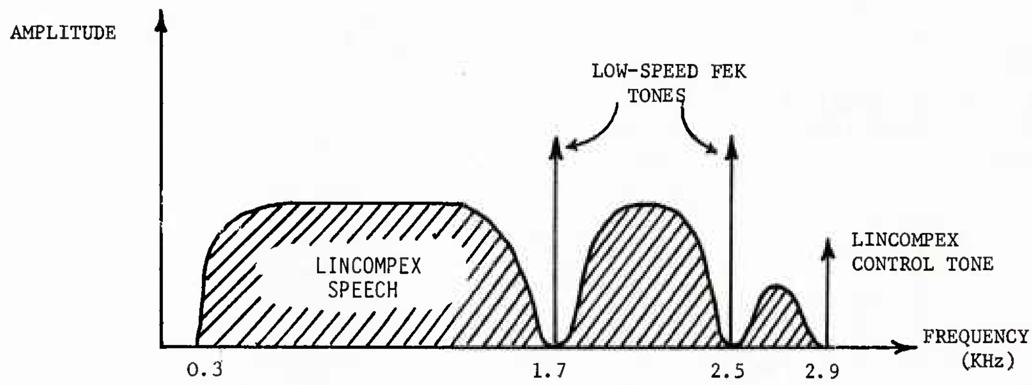


FIG 9 COMPOSITE BASEBAND SPECTRUM OF LINCOMPEX-PROCESSED SPEECH AND LOW-SPEED FEK TELEGRAPHY

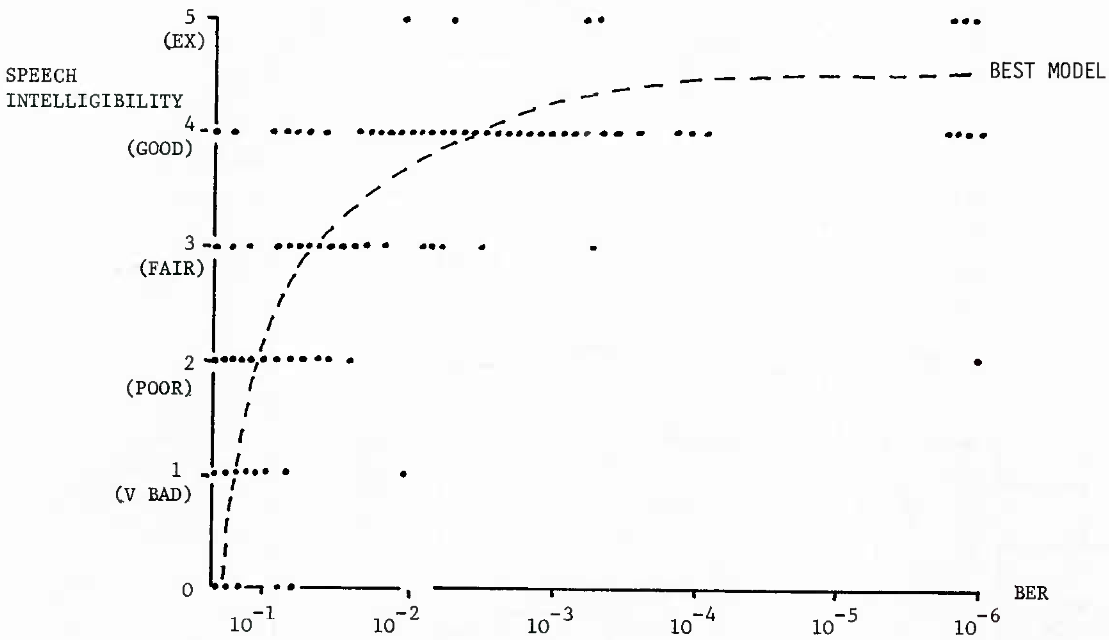


FIG 10 EMPIRICAL RELATIONSHIP BETWEEN LINCOMPEX-PROCESSED SPEECH INTELLIGIBILITY AND LOW-SPEED FEK TELEGRAPHY BER

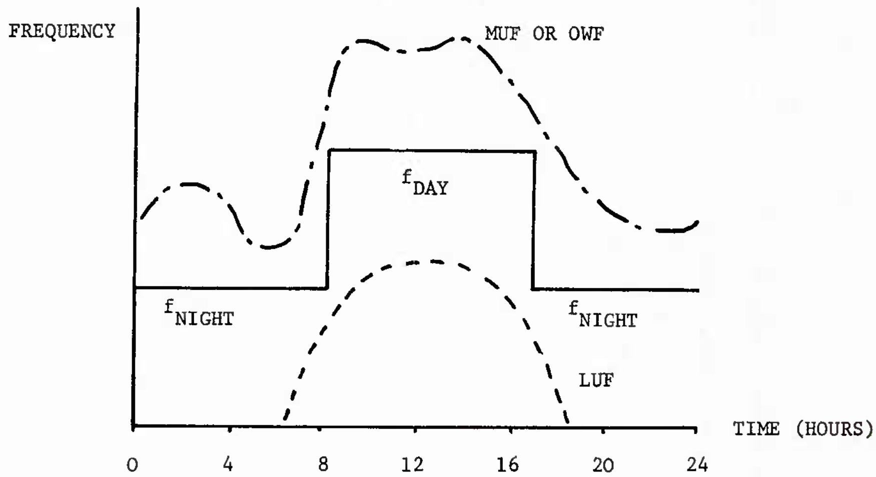


FIG 11 EXAMPLE OF 2-FREQUENCY OPERATION

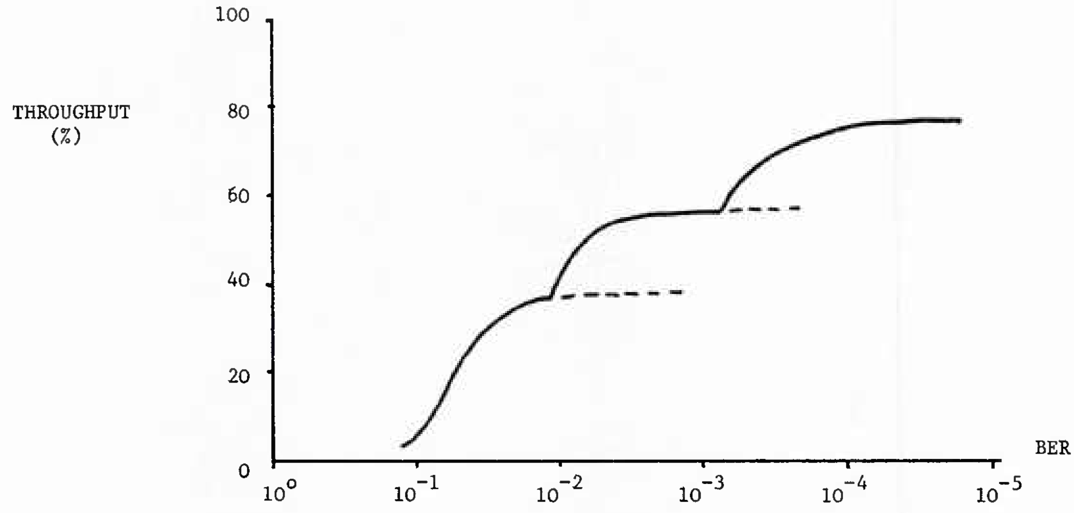


FIG 12 VARIABLE REDUNDANCY CODING IN AN ARQ SYSTEM

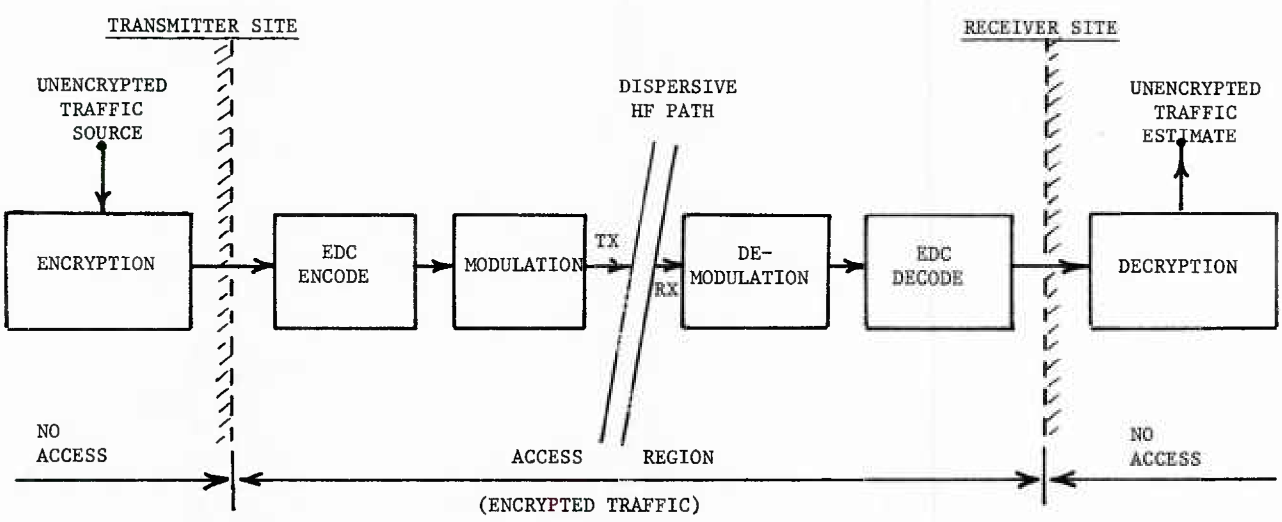


FIG 13 THE USE OF AN AUXILIARY EDC SYSTEM FOR RTCE WITH ENCRYPTED TRAFFIC

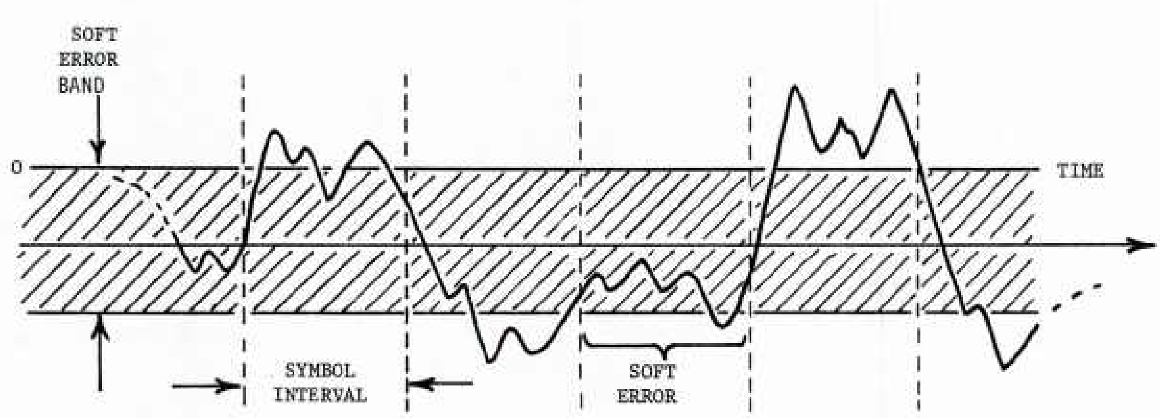


FIG 15 PRINCIPLE OF SOFT ERROR COUNTING

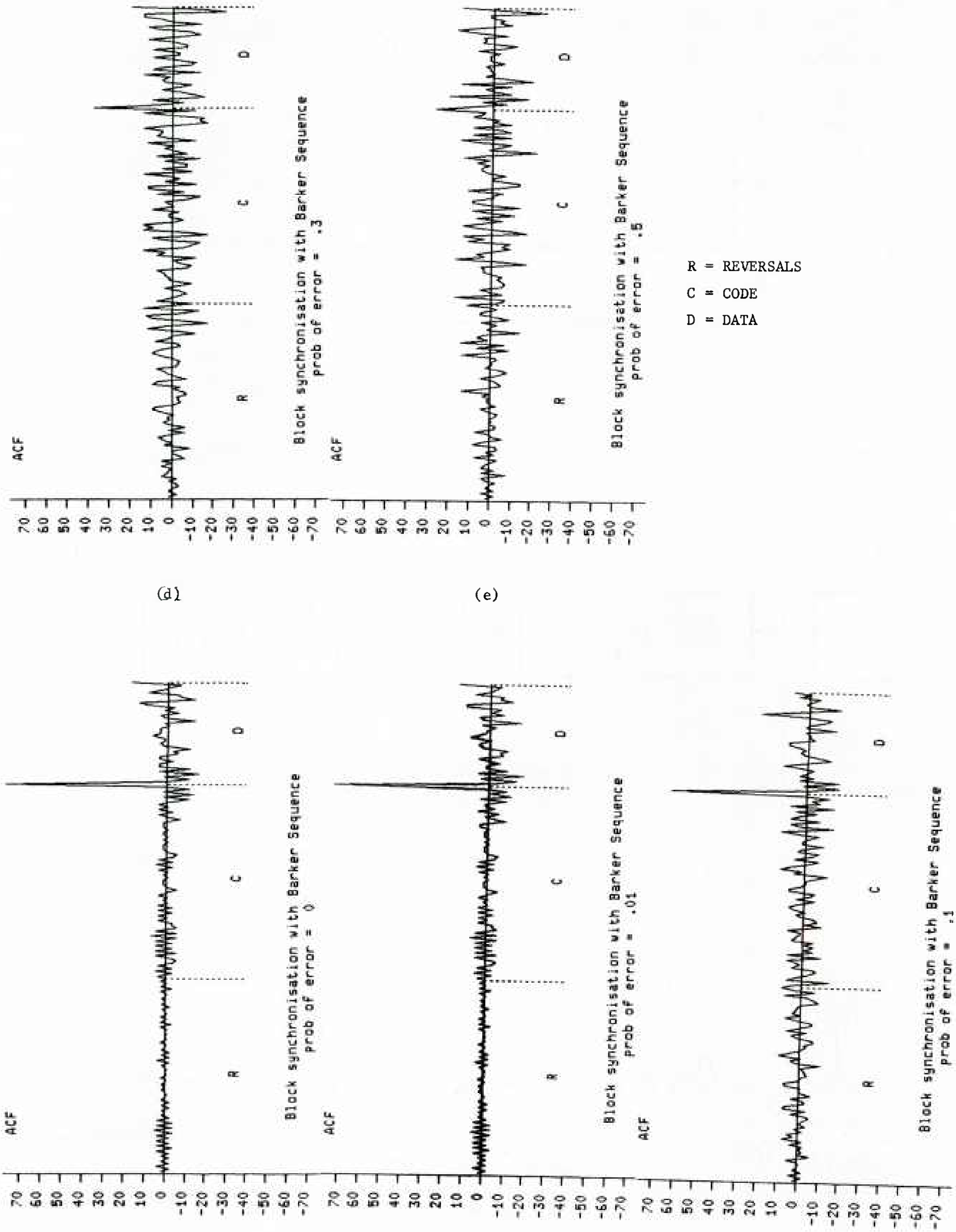


FIG 14 OUTPUT OF CORRELATOR IN RESPONSE TO SYNCHRONISATION PREAMBLE FOR A RANGE OF ERROR PROBABILITIES

# AN INTRODUCTION TO ERROR-CONTROL CODING WITH APPLICATION TO HF COMMUNICATIONS

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Summary

In these past forty years, there has been an uninterrupted trend toward the digitization of communications, with far reaching consequences in terms of improved reliability, increased operational speed, reduced equipment size, freedom from calibration problems, improved ability to mechanize complicated signal processing algorithms, etc. The resulting increase of the demand for reliable, low-error-rate, high-speed digital data transmission has created the need for the adoption of coding schemes. This holds true in general, for all types of communication channels, in particular for HF links. As it is appreciated by the practicing communicator, a problem in high-speed data transmission is the occurrence of errors: codes provide an effective means for their control. What is perhaps less appreciated is that coding represents also a way to boost the data rate in a channel toward the theoretical limit established by Shannon for that channel. Linear block codes (of which cyclic codes are a subclass) and convolutional codes are the main categories of codes of interest to HF communications. They are capable of correcting random errors due to white Gaussian noise, as well as burst errors due to impulse noise. In block codes, a block of information bits is followed by a group of check bits. The latter verify the presence of errors in the former. In convolutional codes, check bits are continuously interleaved with information bits, and they check the presence of errors not only in the block immediately preceding them, but in other blocks as well. For the various coding schemes reviewed in this lecture, several numerical examples are given, to help in the quantitative appraisal of the merits of the code, versus required equipment complexity.

## A. INTRODUCTION

### 1. General

The first half of the twentieth century brought about the development of radio communications, characterized by the transmission of messages, speech and television that was mostly in analog form (and with the HF ionospheric links figuring prominently in the handling of the intercontinental traffic). The second half has seen an uninterrupted trend toward the digitization of communications, with far reaching consequences in terms of improved reliability, increased operational speed, reduced equipment size, freedom from calibration problems, improved ability to mechanize complicated signal processing algorithms, etc. [Golomb, 1964, (1); Viterbi and Omura, 1979, (2); Wozencraft and Jacobs, 1965, (6); Gallager, 1968, (7)].

Digital communications were given impetus by several driving needs, most prominently by the ever growing demand for data exchanges between computers and remote terminals. There is, however, an additional aspect that has a great importance and that directly relates to the topic of this lecture: the ability of digital techniques to make it feasible and practical to approach the theoretical efficiency limit of a communication channel. It is here, in fact, that coding finds an application of general use, as an approach to optimize communications on a given channel (in our case, ionospheric HF channels) rather than providing a way to achieve secrecy in military communications.

In order to better illustrate the point above, we must backtrack several decades, and go back to the work of Hartley in the late '20s, and to the publications of Shannon, Wiener, Fano and other pioneers of digital communications, in the mid '40s. These authors, all of great theoretical strength, developed methods for the computation of the efficiency of a communications system, and established the theoretical maximum that, for

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each type of system, this efficiency can attain. Shannon had the intuition that achieving error-free digital communications on noisy channels, and performing the most efficient conversion of an analog signal into coded digital form, were two facets of the same problem, having a common solution. Shannon's main result is actually that, as long as the input rate to a channel encoder is less than a quantity called the channel capacity C, encoding and decoding approaches do exist, that, asymptotically and for arbitrarily long sequences, lead to the error-free reconstruction, at the receiving terminal of the link, of the input sequence. The capacity C (bits/sec) can be easily computed from the receiver's bandwidth W (Hz) and from the Signal/Noise ratio (power ratio) S/N:

$$C = W \log_2 \left(1 + \frac{S}{N}\right)$$

What coding ultimately does is to maximize the likelihood of correct interpretation, at the receiving end of the link, of the incoming waveforms, and to push the data rate toward the theoretical limit C established by Shannon . On the strength of these conceptual developments, a wealth of codes were developed, such as the Shannon-Fano-Hoffman codes for a discrete channel, the Hamming codes for a discrete channel with discrete noise, the Bose-Chaudhuri-Hocquenghen codes, that have found use in HF military links and that resulted from contributions due to Reed, Muller, Golay, Slepian, and others. For a channel with white Gaussian noise, the search for high-efficiency codes translated in the search for waveforms that exhibit the smallest possible mutual correlation.

For several decades, the design of efficient codes was exclusively a theoretical exercise, with no opportunity to reach the stage of engineering implementation. Technology was not yet on a par with the complicated hardware that they required. However, recent advances in technology, especially the development of large scale, integrated-circuits building blocks, have changed all this. It finally became feasible and practical to mechanize coders and decoders that are known from theoretical work to be optimum, and several implementations have actually already entered the practice of HF communications. Figure 1 is a typical block diagram of a digital link between two terminals. Usually, the alphabet is binary (coding in digits 1 and 0) and the source may be a computer, whose output is transformed by the source encoder into a (binary) sequence of ones and zeros. The transformation is done in such a way that the amount of bits/sec that represents the source output, is the minimum required by the source frequency content and by the number of discernible levels. Also the transformation is done in such a way that the reconstruction of the source output at terminal B is feasible and adheres to its original. The channel encoder (as well as the decoder in reception) is the most important unit from the standpoint of the topic of this lecture: its function is to transform the binary data sequence at the output of the source encoder into some longer sequence that is called the code word. This longer sequence enters then a modulator (to modulate for instance by FSK, or Frequency Shift Keying, a radio carrier). The block called channel is the medium where the signals propagate, in our case an ionospheric path at HF; while in the channel, the signals are corrupted by noise and interference.

At terminal B, the demodulator makes the decision whether, for each received signal, the transmitted waveform was a 1 or a 0. The channel-decoder, then, by knowing the rules by which the channel encoder did operate, attempts to correct the transmission errors and performs an estimate of the actual code word that was transmitted. The source decoder transforms this code word, that has been reduced to a stream of information bits, into an estimation of the actual source output, and delivers it to the user. If the channel is characterized by low noise, the various estimations performed by the units of terminal B will be very similar to the functions that they are meant to represent. If the channel, on the contrary, is very noisy, substantial differences may arise [Lin, 1970, (3)].

In Figure 1, the channel encoder and the channel decoder perform the function of error control. This is done through a judicious use of redundancy [Shanmugan, 1979, (4)]. The channel encoder adds digits to the bit stream of the source's message. While these additional bits do not convey information in themselves, they make it possible for the channel decoder to detect and correct errors in the received, information-bearing, bit stream, thus reducing the probability of errors. The encoder divides the input message bits into blocks of k message bits, and replaces each k bit message block with an n bit codeword by adding n-k check bits to each message block. The decoder looks at the received version of the original code word, which may occasionally contain errors, and attempts to decode the k message bits. The design of the encoder and decoder consists of selecting rules for generating code words from message blocks and for extracting message blocks from the received version of the codewords, with the fundamental aim of lowering the overall probability of error.

2. Examples of code generation

In order to start with a simple example, let's consider a block with five horizontal lines (or rows) and with seven vertical columns. This block (with n = 5 x 7 = 35) represents schematically the word "Hello" in the teletype 7-unit alphabet:

| | |
|---|---|
| H | 0001011 |
| e | 0100001 |
| l | 0010011 |
| l | 0010011 |
| o | 0000111 |

A code word, aimed at reducing the probability of errors in the transmission of the block above, can be generated as follows: we add to each line and to each column one more symbol, in order to make the overall number of 1s <u>even</u> in each line and in each column. At the end, we add a symbol at the lower right corner, to make <u>even</u> the number of 1s contained in the last line. The new block is as follows:

$$\begin{array}{c|c}
0001011 & 1 \\
0100001 & 0 \\
0010011 & 1 \\
0010011 & 1 \\
0000111 & 1 \\
\hline
0101101 & 0
\end{array}$$

If during the transmission an error occurs, this can be corrected, provided that there is only one of them. Correction is done by checking each line and each column for <u>even parity</u> (an even number of 1s). If there is a single error, one column-check and one line-check will fail, and the error will be identified at the intersection and will be corrected. The 48 symbols of the block above form a code word. This code has a total n = 48 symbols, of which k = 35 symbols are information-carrying symbols. It is referred to as a (48, 35) code. There are n-k = 48 - 35 = 13 check symbols. These are the redundant digits added to the message (by using modulo-2 arithmetic), in order to provide the code word with error-correcting capability.

The mathematical way of generating the code word (48,35) above is as follows. It is an important way because the hardware mechanization of the mathematical scheme (based on matrix calculations) is straightforward, although in some cases it might be cumbersome.

The first step is to multiply the original message matrix (5 x 7) by a first generator matrix (7 x 8), to obtain an intermediate (5 x 8) coded word, using throughout modulo-2 arithmetic.

$$C_1 = D\ G_1 = \begin{bmatrix} 0&0&0&1&0&1&1 \\ 0&1&0&0&0&0&1 \\ 0&0&1&0&0&1&1 \\ 0&0&1&0&0&1&1 \\ 0&0&0&0&1&1&1 \end{bmatrix} \times \begin{bmatrix} 1&0&0&0&0&0&0&0 \\ 0&1&0&0&0&0&0&0 \\ 0&0&1&0&0&0&0&0 \\ 0&0&0&1&0&0&0&0 \\ 0&0&0&0&1&0&0&0 \\ 0&0&0&0&0&1&0&1 \\ 0&0&0&0&0&0&1&0 \end{bmatrix} = \begin{bmatrix} 0&0&0&1&0&1&1&1 \\ 0&1&0&0&0&0&1&0 \\ 0&0&1&0&0&1&1&1 \\ 0&0&1&0&0&1&1&1 \\ 0&0&0&0&1&1&1&1 \end{bmatrix}$$

D (5x7 = 35)

original message

first generator matrix $G_1$ (7x8 = 56)

identity matrix of order 7

first arbitrarily chosen matrix (to implement desired parity-check scheme, line-wise)

inter-mediate code word $C_1$ (5x8 = 40)

The second step consists of making the Transpose of the (5 x 8) intermediate code word, and of the following additional operations: multiply this new (8 x 5) matrix by a second generator matrix (5 x 6), in order to obtain the transpose of the final coded word.

$$C^T = C_1^T \times G_2 = \underbrace{\begin{bmatrix} 0&0&0&0&0 \\ 0&1&0&0&0 \\ 0&0&1&1&0 \\ 1&0&0&0&0 \\ 0&0&0&0&1 \\ 1&0&1&1&1 \\ 1&1&1&1&1 \\ 1&0&1&1&1 \end{bmatrix}}_{\substack{C_1^T \\ (8x5\ =\ 40)}} \times \underbrace{\left[\begin{array}{ccccc|c} 1&0&0&0&0&1 \\ 0&1&0&0&0&1 \\ 0&0&1&0&0&0 \\ 0&0&0&1&0&0 \\ 0&0&0&0&1&1 \end{array}\right]}_{\substack{G_2 \\ (5x6\ =\ 30)}} = \underbrace{\left[\begin{array}{ccccc|c} 0&0&0&0&0&0 \\ 0&1&0&0&0&1 \\ 0&0&1&1&0&0 \\ 1&0&0&0&0&1 \\ 0&0&0&0&0&1 \\ 1&0&1&1&1&0 \\ 1&1&1&1&1&1 \\ 1&0&1&1&1&0 \end{array}\right]}_{\substack{C^T \\ (8x6\ =\ 48)}}$$

identity matrix of order 5

second arbitrary chosen matrix (to implement desired parity-check scheme, column-wise)

The third and last step consists of obtaining the final (6 x 8) coded word by transposing. We have:

$$C = \text{Transpose of } C^T = \underbrace{\left[\begin{array}{ccccccc|c} 0&0&0&1&0&1&1&1 \\ 0&1&0&0&0&0&1&0 \\ 0&0&1&0&0&1&1&1 \\ 0&0&1&0&0&1&1&1 \\ 0&0&0&0&1&1&1&1 \\ \hline 0&1&0&1&1&0&1&0 \end{array}\right]}_{\substack{C \\ (6x8\ =\ 48)}}$$

All the codes illustrated in this lecture are based on ideas similar to the principle that allowed us to generate the code word (48,35) above. The mathematics involved might be of higher caliber, the codes might be more efficient. However, there is a striking fundamental similarity between the simple code introduced above and the more sophisticated ones that we will illustrate later on in this lecture.

Let's see now another example that shows how the probability of error is reduced by the adoption of a coding scheme. Let's assume that we have a HF link with a bandwidth of 3 KHz and a Signal-to-Noise ratio of 13 dB (power ratio = 20). We want to transmit a data rate (from the source), of 1200 bits/sec, with an error rate less than $10^{-6}$. We have available a modem that can operate at the rates of 1200, 2400, 3600, 4800, and 6000 bits/sec, with error probabilities respectively of $2 \times 10^{-4}$, $4 \times 10^{-4}$, $8 \times 10^{-4}$, $1.4 \times 10^{-3}$ and $2.4 \times 10^{-3}$. According to the Shannon theorem, the channel capacity C is, in our case:

$$C = W \log_2 (1 + S/N) = 3 \text{ KHz} \log_2 (1 + 20) = 13 \text{ Kbits/sec}$$

Therefore, since the source bit rate is less than the channel capacity C, we should be able to find a way to transmit the data with an infinitesimally small probability of error. Let's start considering a code obtained simply by tripling the symbols that come out from the source encoder: if it is a 0, we use 000; if it is a 1, we use 111. This are now our codewords. We adopt in reception a logic based on majority ruling: an error occurs when in a word two or more symbols are wrong. The data rate that we must use in the modem is 3600 bits/sec, with corresponding error rate $8 \times 10^{-4}$. The codeword is thus characterized by the following error probability:

$$P_e \text{ (that two or more bits in the triplet are in error)} = \binom{3}{2} q_c^2 (1 - q_c) + \binom{3}{3} q_c^3$$

where $q_c = 8 \times 10^{-4}$, if we signal at the rate in the modem of 3600 bits/sec. Numerically, the equation above yields:

$$P_e = 1.9 \times 10^{-6} > 10^{-6} \text{ ($10^{-6}$ is the required error rate, not to be exceeded)}$$

The results above are therefore not acceptable and we must try a longer codeword, in order to fulfill the error rate requirement. Let's generate in the modem the codeword 00000 for every 0 at the source's output, and 11111 for every 1. Now the modem operates at the rate of 6000 bits/sec (five times the 1200 bits/sec rate of the source) and $q_c = 2.4 \times 10^{-3}$. We still use the majority rule logic: an error will occur in reception when three or more symbols are received wrong. The error probability $P_e$ is now:

$$P_e \text{ (that three or more bits in the quintuplet are in error)} =$$
$$= \binom{5}{3} q_c^3 (1 - q_c)^2 + \binom{5}{4} q_c^4 (1 - q_c) + \binom{5}{5} q_c^5 = 1.376 \times 10^{-7} < 10^{-6}$$

With the quintuplet we meet therefore the error rate requirement.

3. Error detection and error correction approaches

In the examples given in Section 2, we adopted an approach that corrects as best as feasible, the received errors, caused by noise. This approach belongs to the category of forward-error-correcting codes. There is another category of approaches that performs in reception the error detection (detection only, not inclusive of correction) and in which the terminal B in Figure 1 re-transmits back to terminal A the received message, once that an error has been detected, for a repetition of the exchange. There are some obvious inconveniences associated with this method: we must have a two-way link between the two terminals, we must use time message re-transmission, we must send from B to A an acknowledgement even when the message received at B appears correct, etc. However, there is the great advantage that the overall probability of error is now much lower than achievable with the forward-error-correcting codes. We can easily see how this happens with the following example. Let's assume that the receiver/decoder accepts only (f the triplets of the example in Section 2 are adopted) a 0 when what is received is 000, and a 1, only when 111 is received. In any other case, terminal B requests from A a re-transmission. An error now occurs only if all three bits of the triplet are wrong. We have therefore for $P_e$:

$$P_e \text{ (that all three bits are wrong)} = (q_c)^3 = 5.12 \times 10^{-10} \ll 10^{-6}$$

In fact, because the modem now works at 3600 bits/sec, we know that $q_c = 8 \times 10^{-4}$. Therefore, we do not need in this case to use any quintuplet in order to satisfy the overall error probability requirement.

4. Random errors and burst errors, and their control by coding

Generally, two types of noise are encountered in communication channels. The first kind is Gaussian noise, including thermal noise in the equipment, cosmic noise, etc. This noise is often white. In the case of white Gaussian noise, the occurrence of errors during a particular signaling interval, does not affect the performance of the communication system during the subsequent signaling interval. The discrete channel in this case can be modeled by a binary symmetric channel and the errors due to white Gaussian noise are referred to as random errors [Shanmugan, 1979, (4)]. A second type of noise often encountered in a communication channel is the impulse noise, where high intensity noise bursts sporadically appear during long quiet periods. When this happen, several bits in sequence may be affected, and errors occur in bursts.

There are error control schemes that are particularly effective in counteracting random errors. Other schemes have special immunity from burst errors. Error correcting codes are divided in two general categories: block codes (of which a subclass are the cyclic codes) and convolutional codes. In block codes, a block of information bits is followed by a group of check bits that are derived from the former. At the receiving terminal, the check bits are used to verify the information bits in the block immediately preceding the check bits. In convolutional codes, check bits are continuously interleaved with information bits, not only in the block immediately preceding them, but in other blocks as well. The cyclic codes have particular advantages, in as much as they simplify considerably the required encoding and decoding equipment.

B. LINEAR BLOCK CODES

1. The fundamental concept of these codes

We will discuss now the block codes, in which each block of k message bits is encoded into a block of n > k bits, by adding n-k check-bits derived from the k message bits. They are called linear when each of the codewords (n-bit block at the output of the channel encoder) can be expressed as a linear combination of k linearly independent code vectors: linear block codes are the most commonly used block codes and we will limit our discussion to them. In all the operations pertaining the generation of these codes, we must remember to use modulo 2 arithmetic. This is how this arithmetic works: we make additions the regular way (for instance, 1 + 1 = 2). However, we divide then the result by 2 and we take the remainder as the final result of the modulo 2 operation. For instance, for the 1 + 1 = 2 case, we divide 2 by 2, the remainder is zero, and therefore we write 1 + 1 =0. If we sum 0 + 1, and we divide the result by 2, the remainder of the regular division is 1 and thus we conclude 0 + 1 = 1.

Let's see some examples of code generation. An effective way of implementing this generation is the use of matrix representation: C = DG, where D is the message vector, C the final codeword and G is the so-called generator matrix, that embodies the rules adopted for detecting errors in the received codeword. The generator matrix has dimensions k x n, and has the form:

$$G = [I_k \,\vdots\, P]_{k\times n}$$

where $I_k$ is the identity matrix of order k, P is an arbitrary matrix of dimensions kx(n-k). This last matrix, once chosen, fully defines the (n,k) block code completely. Suppose that the generator matrix G of a (6,3) block code is

$$G = \left[\begin{array}{ccc|ccc} 1 & 0 & 0 & 0 & 1 & 1 \\ 0 & 1 & 0 & 1 & 0 & 1 \\ 0 & 0 & 1 & 1 & 1 & 0 \end{array}\right]$$

and we want to find all code vectors of this code. The message block size for this codes is 3 and the overall length of the code vectors is n = 6. The possible 8 messages are: (0,0,0), (0,0,1), (0,1,0), (0,1,1), (1,0,0), (1,1,0), (1,1,1), (1,0,1). The code vector for the message block D = (111) is:

$$C = DG = (111)\begin{bmatrix} 1 & 0 & 0 & 0 & 1 & 1 \\ 0 & 1 & 0 & 1 & 0 & 1 \\ 0 & 0 & 1 & 1 & 1 & 0 \end{bmatrix} = (1\ 1\ 1\ 0\ 0\ 0)$$

The encoder has to store the G matrix (or, at the very least, the submatrix P of G) and must perform binary arithmetic operations to generate the check bits. Associated with each (n,k) block code, there is a parity check matrix H which is defined as:

$$H = [P^T \,\vdots\, I_{n-k}]\ (n-k)xn$$

where $P^T$ is the transpose of the matrix P. The transpose is obtained from the original matrix by changing place of each element of the matrix, following the rule that an element $P_{ij}$ goes to the place $P_{ji}$. In this way, if the original matrix has, for example, 7 lines and 5 columns, its transpose will have 5 lines and 7 columns.

Let's see now how at the receiving terminal, the decoder utilizes the parity check matrix H. This matrix is used to verify whether a codeword C is generated by the matrix $G = [I_k \mid P]$. This verification can be done as follows. C is a codeword in the (n,k) block code generated by G if and only if $CH^T=0$, where $H^T$ is the transpose of the matrix H. If C is a code vector transmitted over a noisy channel and R is the received vector, we have that R is the sum of the original code vector C and of an error vector E. The receiver does not know C and E and its function is to obtain C from R, and the message block D from C. The receiver perform its function by determining a (n-k) vector S defined as

$$S = R\,H^T$$

This vector is called the error syndrome of R, and we can write for it: $S = CH^T + EH^T = EH^T$ because $CH^T = 0$. Thus the syndrome of a received vector is zero if R is a valid code vector. Furthermore, S is related to the error vector E and the decoder uses S to detect and correct errors. As an example, let's consider a (7,4) block code generated by

$$G = \left[\begin{array}{cccc|ccc} 1 & 0 & 0 & 0 & 1 & 1 & 1 \\ 0 & 1 & 0 & 0 & 1 & 1 & 0 \\ 0 & 0 & 1 & 0 & 1 & 0 & 1 \\ \underbrace{0 \quad 0 \quad 0 \quad 1}_{I_4} & & & & \underbrace{0 \quad 1 \quad 1}_{P} & & \end{array}\right]$$

The parity check matrix H for this code is:

$$H = \left[\begin{array}{cccc|ccc} 1 & 1 & 1 & 0 & 1 & 0 & 0 \\ 1 & 1 & 0 & 1 & 0 & 1 & 0 \\ \underbrace{1 \quad 0 \quad 1 \quad 1}_{P^T} & & & & \underbrace{0 \quad 0 \quad 1}_{I_3} & & \end{array}\right]$$

For a message block D = (1011), the code vector C is given by C = DG = (1 0 1 1 ¦ 0 0 1), and the syndrome S is $S = CH^T = (000)$. Now, if the third bit of the code vector C suffered an error in transmission, then the received vector R will be: R = (1 0 0̲ 1 0 0 1) = (1 0 1 1 0 0 1) + (0 0 1 0 0 0 0) = C + E, and the syndrome of R is $S = RH^T = (101) = EH^T$, where E is the error vector (0 0 1 0 0 0 0). Note that the syndrome S for an error in the third bit is the third row of the $H^T$ matrix. It can be verified that, for this code, a single error in the $i^{th}$ bit of C would lead to a syndrome vector that would be identical to the $i^{th}$ row of the matrix $H^T$. Thus single errors can be corrected at the receiver by comparing S with the rows of $H^T$ and correcting the $i^{th}$ received bit if S matches with the $i^{th}$ row of $H^T$. This simple scheme does not work if multiple errors occur.

## 2. Terminology and basic properties

In defining the error correction capability of a linear block code, some new terminology is utilized, such as <u>the weight</u> of a code and its <u>minimum distance</u>. The weight of a code is defined as the number of non-zero components in the codeword. <u>The distance</u> between two code vectors is the number of components in which they differ, while the <u>minimum distance</u> of a block code is the smallest distance between any pair of codewords in the code. An important property to remember is that the minimum distance of a block code is equal to the minimum weight of any non-zero word in the code. An example will clarify these points. Let's consider the (6,3) block code introduced in the previous section. Table I here below gives the weight of the various codewords.

Table I
Codewords' Weight

| Codeword | Weight |
|---|---|
| 000000 | 0 |
| 001110 | 3 |
| 010101 | 3 |
| 011011 | 4 |
| 100011 | 3 |
| 101101 | 4 |
| 110110 | 4 |
| 111000 | 3 |

From the second column in Table I, we see that the minimum distance is 3. No two codewords in this code differ in less than three places. The ability of a linear block code to correct random errors can be specified in terms of the code's minimum distance. The decoder will associate a received vector R with a transmitted code vector C if C is the code vector closest to R in the sense of the Hamming distance. Another property to remember is that a linear block code with a minimum distance $d_{min}$ can correct up to $\left[(d_{min} - 1)/2\right]$ errors and detect up to $d_{min} - 1$ errors in each codeword, where $\left[(d_{min} - 1)/2\right]$ denotes the largest integer no greater than $(d_{min} - 1)/2$. If we call t the errors that the code will correct, we have:

$$t \le \left[(d_{min} - 1)/2\right]$$

We can also show that such a code can detect up to $d_{min} - 1$ errors. We can deduce from the above that for a given n and k, we should design a (n,k) code with minimum distance as large as possible.

3. Error detection and correction

In reception, the matrix H is used to verify whether or not a received code word R was generated by the matrix G. We proceed this way: we compute the Syndrome S of the received vector R: $S = R\,H^T$. If S=0, this verify that correct reception took place. If, on the contrary, $S \neq 0$, an error occurred in the transmission.

How we correct the error ? One way is to "look-up" at a standard array prepared for the code under consideration. As an example, let's consider the (6,3) code introduced in B.1 above. Suppose that the transmitted word was C=(111000) and we receive R = (111 100). We can verify that the Syndrome $S \neq 0$, and we know that an error occurred. The look-out table for the (6,3) code has been worked out by Shanmugan (4), and it is reproduced in Table II.

Table II

Look-out Table for the (6,3) Code

| Syndrome | Co-Set Leader | | | | | | | |
|---|---|---|---|---|---|---|---|---|
| 000 | 000000 | 100110 | 010011 | 001101 | 110101 | 011110 | 101011 | 111000 |
| 110 | 100000 | 000110 | 110011 | 101101 | 010101 | 111110 | 001011 | 011000 |
| 011 | 010000 | 110110 | 000011 | 011101 | 100101 | 001110 | 111011 | 101000 |
| 101 | 001000 | 101110 | 011011 | 000101 | 111101 | 010110 | 100011 | 110000 |
| 100 | 000100 | 100010 | 010111 | 001001 | 110001 | 011010 | 101111 | 111100 |
| 010 | 000010 | 100100 | 010001 | 001111 | 110111 | 011100 | 101001 | 111010 |
| 001 | 000001 | 100111 | 010011 | 001100 | 110100 | 011111 | 101010 | 111001 |
| 111 | 100001 | 000111 | 110010 | 101100 | 010100 | 111111 | 001010 | 011001 |

<u>First Row</u>

The nine columns are: Syndrome, Co-set Leader [the first of the eight legitimate code vectors of the (6,3) code], while the remaining seven columns are the additional seven legitimate vectors.

<u>Second row, and following rows</u>

We choose an hypothpetical syndrome, and we adopt as co-set leader a n-tuple that does not appear in the first row. Once that is done, we complete the row by adding the n-tuple to all codewords of the first row. For instance, the second row is computed as follows:

| | | | |
|---|---|---|---|
| 110 | 100000 | 100000+<br><u>100110=</u><br>000110 | etc. etc. |

Use of the look-out table in decoding

We have assumed in our example in this Section that R = (111 100). Because $S \neq 0$ we know that this is a wrong codeword. We look in Table II and we see that this code word appears in the fifth row of the last column. There is a good probability that the word that had been transmitted (and erroneously received) is 111 000, the word in the first row of the last column.

4. Hamming Codes (a class of linear block codes)

These codes are popular because they require the minimum number of bits for correction of single errors. They are characterized by the following parameters:

$$d_{min} = 3$$
$$n = 2^m - 1 \quad \text{(where m is a positive integer)}$$
$$n - k = m$$
$$k = 2^m - m - 1$$
$$t = 1$$

They always correct single errors; can detect but not correct double errors; multiple errors result, in general, in incorrect decoding.

C. CYCLIC CODES

1. General properties

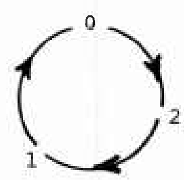
The research conducted on linear block codes has identified a subclass of these codes, called the cyclic codes, that have especially attractive properties, such as ease of implementation; ability to correct large numbers of random errors, long burst of errors and loss of synchronization; etc. These codes have a mathematical structure that helps considerably in the design of error-correction features and furthermore, due to their cyclic nature, require encoding/decoding equipment that is considerably simpler than required by regular block codes. A cyclic (n,k) code has the property that every cyclic shift of a code word is another code word. That is, if

$$C = \left[C_{n-1}, \quad C_{n-2}, \quad \ldots\ldots \; C_0\right]$$

is a code word, so are

$$\begin{matrix} [C_{n-2}, & C_{n-3}, & \ldots\ldots & C_{n-1}] \\ \vdots & \vdots & & \vdots \\ [C_0, & C_{n-1}, & \ldots\ldots & C_1] \end{matrix}$$

Example for n=3




This formulation of cyclic codes suggests treating the elements of each word as coefficient of a polynomial of degree n-1. With this convention, the codeword C can be represented with the polynomial

$$C(x) = C_{n-1}x^{n-1} + C_{n-2}x^{n-2} + \ldots\ldots + C_1x + C_0$$

The condition that every cyclic shift of a code word be another code word can be expressed as follows. That C(x) is a code word implies that $x^iC(x)$ modulo-$(x^n+1)$ is also a code word for all i. It can be verified that multiplication by x-modulo-$(x^n+1)$ results in a cyclic shift and that the coefficients of $x^iC(x)$-modulo-$(x^n+1)$ are, in fact,

$$\left[C_{n-i-1}, \; C_{n-i-2}, \ldots\ldots \; C_1, \; C_0, \; C_{n-1} \ldots\ldots C_{n-i}\right]$$

The encoding and the syndrome calculation of cyclic codes can be simply implemented with shift-registers, having feed-back connections. Also, because cyclic codes have considerable algebraic structure, it is possible to design and implement various simple and efficient decoding methods.

Examples of cyclic codes used in HF communications are:

> Golay Code (23,12), capable of correcting any combination of up to 3 random errors in a block of 23 digits;
>
> Bose-Chaudhuri-Hocquenghen (BCH) codes, capable of correcting large numbers of random errors. With m and t positive integers ($t < 2^{m-1}$), block length $n = 2^m-1$, parity-checks $(n-k)mt$, $d_{min} \geq 2t + 1$.

2. Polynomial Representation of Cyclic Codes

As an example, let's have an 11-bit binary word: (10001 001010). We can state that this word is fully equivalent to the polynomial expression:

$$(x^{10} + x^6 + x^3 + x)$$

There is also full equivalence between a block code's generator matrix and a cyclic code's generator polynomial. For instance, the generator matrix of a (15,11) block code, that also belongs to the cyclic codes subclass:

$$G = [I_k P] = \left|\begin{array}{ccccccccccc|cccc}
1&0&0&0&0&0&0&0&0&0&0&1&0&0&1\\
0&1&0&0&0&0&0&0&0&0&0&1&1&0&1\\
0&0&1&0&0&0&0&0&0&0&0&1&1&1&1\\
0&0&0&1&0&0&0&0&0&0&0&1&1&1&0\\
0&0&0&0&1&0&0&0&0&0&0&0&1&1&1\\
0&0&0&0&0&1&0&0&0&0&0&1&0&1&0\\
0&0&0&0&0&0&1&0&0&0&0&0&1&0&1\\
0&0&0&0&0&0&0&1&0&0&0&1&0&1&1\\
0&0&0&0&0&0&0&0&1&0&0&1&1&0&0\\
0&0&0&0&0&0&0&0&0&1&0&0&1&1&0\\
0&0&0&0&0&0&0&0&0&0&1&0&0&1&1
\end{array}\right|$$

is fully equivalent to the following generator polynomial:

$$g(x) = x^4 + x + 1$$

We have, in fact:

$$G = [I_k P] = \begin{bmatrix}
x^{14}&&&&&&&&&&&x^3&&&1\\
&x^{13}&&&&&&&&&&x^3&x^2&&1\\
&&x^{12}&&&&&&&&&x^3&x^2&x&1\\
&&&x^{11}&&&&&&&&x^3&x^2&x&\\
&&&&x^{10}&&&&&&&&x^2&x&1\\
&&&&&x^9&&&&&&x^3&&x&\\
&&&&&&x^8&&&&&&x^2&&1\\
&&&&&&&x^7&&&&x^3&&x&1\\
&&&&&&&&x^6&&&x^3&x^2&&\\
&&&&&&&&&x^5&&&x^2&x&\\
&&&&&&&&&&x^4&&&x&1
\end{bmatrix} = \begin{bmatrix}
(x^{10}+x^7+x^6+x^4+x^2+x+1)\,g(x)\\
(x^9+x^6+x^5+x^3+x+1)\,g(x)\\
(x^8+x^5+x^4+x^2+1)\,g(x)\\
(x^7+x^4+x^3+x)\,g(x)\\
(x^6+x^3+x^2+1)\,g(x)\\
(x^5+x^2+x)\,g(x)\\
(x^4+x+1)\,g(x)\\
(x^3+1)\,g(x)\\
x^2 g(x)\\
x g(x)\\
g(x)
\end{bmatrix}$$

where

$$g(x) = x^4 + x + 1$$

3. Encoding operation for Cyclic Codes and its mechanization

To encode a non-cyclic block code, we multiply the original message by the known generator matrix. This process is quite lengthy. For a cyclic code, the procedure is much simpler and shorter. In fact, to encode into a n-bit cyclic code a k-bit message word $d(x)$, we simply multiply $d(x)$ by $x^{n-k}$, and we add the remainder of the division:

$$\frac{x^{n-k}\,d(x)}{g(x)}$$

Therefore, we have:

$$C(x) = x^{n-k}\, d(x) + \text{Rem}\left\{\frac{x^{n-k}\, d(x)}{g(x)}\right\}$$

A numerical example will clarify these points:

- We want to encode a (15,11) cyclic code $d(x) = (x^{10} + x^6 + x^3 + x)$ with the generator polynomial $g(x) = x^4 + x + 1$ (we have already encountered this code, previously in this lecture);
- We compute the remainder: $\text{Rem}\left\{\frac{x^{n-k}\, d(x)}{g(x)}\right\} = \text{Rem}\left\{\frac{x^4 (x^{10} + x^6 + x^3 + x)}{x^4 + x + 1}\right\} = x + 1$
- Therefore, we have: $C(x) = x^4\ (x^{10} + x^6 + x^3 + x) + \text{Rem}\ (x) =$ $= x^{14} + x^{10} + x^7 + x^5 + x + 1.$

  In the usual binary notation, this is (10001001010 0011).

- We would have obtained the same results, but with a lengthier process, by multiplying the original message by the matrix that generates the block code (15,11).

The mechanization of the encoding operation consists of performing the mathematical steps involved in the encoding equation

$$\frac{x^{n-k}\, d(x)}{g(x)} = q(x) + \frac{r(x)}{g(x)}$$

The division can be accomplished by the dividing circuit consisting of a feedback shift register, [Shanmugan (4)].

4. Decoding of cyclic codes

The Syndrome is computed by dividing the received word $C(x) + e(x)$ by $g(x)$:

$$S(x) = \text{Rem}\left\{\frac{C(x) + e(x)}{g(x)}\right\}$$

This calculation can be performed by using shift registers with feedback loop. If the Syndrome is not zero, one way to correct errors is to use a look-up table, as we discussed for the case of block codes.

Specific mechanizations of decoders for cyclic codes are the Meggitt Decoder and the Error Trapping Decoder.

## D. CONVOLUTIONAL CODES

### 1. General properties

While in the block codes, a block of n digits depends on the block of the related k input message digits, in a convolutional code, the block of n digits generated by the encoder in a certain time unit depends not only on the block of k message digits within that time unit, but also on the preceding (N-1) blocks of message digits (N>1). Usually the values of n and k are small. Like block codes, convolutional codes can be designed to either detect or correct errors. In practice, they are used mostly for error correction. The analysis of their performance is complicated and is normally done by simulating the encoding and the decoding operations in a digital computer.

In the following sections we will give an introductory level treatment of the convolutional codes, and we will illustrate both the encoding operation and some of the decoding approaches.

### 2. The encoding operation

The code generated by a convolutional encoder is called a (n,k) convolutional code of constraint length nN digits and efficiency k/n. The parameters of the encoder, k, N, and n are in general small integers and the encoder consists of shift registers and modulo-2 adders. An encoder for a (3,1) convolutional code with a constraint length of nine bits is shown in Figure 2. The operation of the encoder proceeds as follows. We assume that the shift register is initially clear. The first bit of the input data is entered into $D_1$. During this message bit interval, the commutator samples the modulo-2 adder outputs $c_1$, $c_2$ and $c_3$. Thus, a single message bit yields three output bits. The next message bit in the input sequence now enters $D_1$, while the content of $D_1$ is shifted into $D_2$ and the commutator again samples the three adder outputs. The process is repeated until the last message bit is shifted into $D_3$. It is important to notice that the convolutional encoder operates on the message stream in a continuous manner, thus it requires very little buffering and storage. Another important point is that each message bit has its influence on N x n digits, where N is the size of the shift register, and n is the number of the commutator segments.

3. The decoding operation

The code tree in Figure 3 that applies to the convolution encoder of Figure 2, is of help in understanding the decoding operation [Shannmugan, 1979, (4)]. The situation depicted in Figure 3 is the one that immediately preceeds the appearance of the $i^{th}$ message $d_i$. We assume $d_{i-1} = d_{i-2} = 0$. The rule for generating the paths in Figure 3 is that we shall diverge upwards from a node of the tree when the input bit is zero. Node A corresponds to the $d_{i-1} = d_{i-2} = 0$. Then, if $d_i = 0$, we move upwards from A and the decoder output is 000. Nodes 1, 2, 3 and 4 can be considered starting nodes for $d_{d+2}$, with $d_i d_{i+1}$ = 00, 01, 10 and 11 respectively. For each starting node, the first message bit influences the code blocks generated from the starting node and the two succeeding ones. Thus, each message bit will affect nine code digits and there are eight distinct 9-bit code blocks associated with each starting node. One way of performing the decoding operation is the so-called Exhaustive Tree-Search method, particularly suitable for decoding convolutional codes when N is small (the decoder, in fact, has to examine $2^N$ branch sections of the code tree). In the case of an ideal transmission channel, free of noise, the received codewords are identical to the transmitted ones. In decoding them, we follow the codewords through the code tree, n bits at a time (n = 3 in the example considered here). The message is reconstructed from the path taken through the tree. If noise is present, we proceed as follows. We make reference to Figures 2, 3, 4, where N = 3 and n = 3, so that $d_i$ influences nN = 9 code digits. Therefore, on one hand, no advantage is gained in examining the codeword beyond the 9 digit block affected by $d_i$, while, on the other hand, we would lose some of the benefits of the code, if we would examine less than nine bits of the received codeword. Assuming that $d_{i-1}$ and $d_{i-2}$ have been correctly decoded, then a starting node is defined on the code tree, and we can identify eight distinct and valid 9-bit code blocks that emerge from this node. By comparing the received code block with the eight valid code blocks, we determine which one of these blocks is closest (Hamming-distance-wise) to the received code block. If the valid code block corresponds to an upward path, $d_i$ is a 0, otherwise, we conclude that $d_i$ is a 1. The procedure is repeated until the entire message is decoded.

Another decoding scheme, suitable for the cases in which N is large, is the so-called Sequential Decoding approach (see Figure 4). In this approach, the decoder, at the arrival of a n-bit code block, compares these bits with the code blocks associated with the two branches diverging from the starting node. The decoder follows an upward (bit decodes as a 0) or a downward path (bit decoded as a 1), depending which of the code blocks exhibits the fewest discrepancies with the received bits. If a received code block contains transmission errors, the decoder might start out on a wrong path in the code tree, and the number of errors will, most likely, grow more numerous than in the case that the decoder follows the correct path. The decoder can be designed to keep a running record of the total number of discrepancies and (if these are particularly numerous), to retrace backwards its path to the node at which the apparent error was made, and then take an alternative branch out of that node, to find eventually a path through N nodes.

The decoder begins retracing when the number of accumulated errors exceeds a threshold, as shown in Figure 4. Since every branch in the code tree is associated with n bits, then, on the average over a long path of j branches, we can expect the total number of bit differences between the decoder path and the corresponding received bit sequence to be $(P_e)^{1}(n)$ (j) even when the correct path is being followed. The number of bit errors accumulated will oscillate about E(j) (see Figure 4), if the decoder is following the correct path (path 1 in Figure 4). The accumulated bit error will, on the contrary, diverge sharply from E(j) soon after a wrong decoder decision is made (see path 2 in Figure 4). When the accumulated errors exceed a discard level (see Figure 4), the decoder decides that an error has been made and retraces its path to the nearest unexplored path and start moving forward again. After some trial and error, an entire N node section of the code tree is retraced and at this point a decision is made about the message bit associated with this N node section. Thus, the sequential decoder operates on short code blocks most of the time, and reverts to trial and error search over long code blocks only when it judges that an error has been made.

## E. BURST ERROR CORRECTION

Random errors occur when each bit is affected independently by the noise. On the contrary, burst errors occur in clusters, due to such causes as the occurrence of deep fades in HF communication links.

A (n,k) linear code is capable of correcting error bursts of length $\ell$, or less, when no burst of length $2\ell$ or less is a code vector. Also, the number of parity-check bits of a linear code must satisfy the condition.

$$n - k \geq 2\ell$$

An effective approach to burst error correction is code interlacing

$$(n,k) \rightarrow (\lambda n,\ \lambda k),$$

where $\lambda$ = interlacing degree. Interlacing is accomplished by arranging code vectors of the (n,k) original code into $\lambda$ rows of a rectangular array, and then transmitting them column by column. Bursts of length $\lambda$ will affect no more than 1 digit in each row.

If the original code corrects single errors, interlaced code corrects single bursts of length $\lambda$ or less. If the original code corrects all combinations of t or fewer errors, the interlaced code corrects any combination of t bursts of length $\lambda$ or less.

There are several codes that have the ability of correcting both random and burst errors, such as product codes, concatenated codes, etc.

## F. CODING APPLICATIONS TO HF COMMUNICATIONS

Most of the coding applications to HF communications have consisted thus far of error detection through the use of parity checks, and of automatic error correction by re-transmission (ARQ). Forward error correction has waited considerably before gaining even partial acceptance, because of the complexity of the required circuitry. However, due to the technological progress, these codes become increasingly practical and their use widespread. Interleaving of message bits over a period longer than the expected fading cycle is an effective way to distribute the bit errors, so that coding techniques can be used to correct them, and to achieve reduction in error rates of several orders of magnitude. This is, of course, at the expense of the information rate. The manner in which coding is used depends upon the nature of the signaling techniques employed by the link. A method that involves the transmission of short packets of information is not well suited to the use of interleaving, since the packets will be in general much shorter than a fade-out. Also, if a high percentage of packets is error-free, it is advantageous to use only error detection with each packet, and to rely on repeats for error correction.

In algebraic encoding, in addition to the information to be transmitted, redundant information is inserted into the channel, as we have seen previously in this lecture. These check bits may be used for forward error correction or for error detection only. Its purpose, in HF communications, is usually to reduce the susceptibility to the deleterious effects of deep fadings. Because these fades can have durations beyond 0.1 seconds, the code must be very long and must have a structure suitable for the correction of error bursts. The most common techniques are to use either very long recurrent codes or short codes with a considerable number of interleaved blocks. The latter approach is used to randomize the effect of burst errors. Decoding of recurrent codes as well as interleaving and de-interleaving cause transmission time delays of about 1 to 2 seconds. Code rates between 1/2 and 4/5 are used if what is required is only error detection. Short block codes (50 bits) may be used under certain conditions. Their effectiveness is however limited because of the high probability that may characterize the number of errors in a code word.

Present state-of-the-art transmission systems use ARQ procedures to increase communication reliability; error detection is based on the recognition of errors in the transmission of a single character. The application of systematic block codes would be already a substantial improvement. Concerning a code's ability to correct both burst errors and random errors, no analytical method for constructing such a code has been developed thus far. However, there are useful results, applicable to specific cases: a) a subclass of cyclic codes has been found that correct both types of errors effectively - however, they either have to be used for low data rates, or be very long; b) random-error correcting codes can be designed that have some simultaneous burst-correction capability, provided that the codes are not used to their fullest random-error correcting capability; c) multiplying a generator polynomial of a cyclic code with d=2t + 1 produces a code which is capable of simultaneously correcting all bursts of length t+1, a result of practical importance only for the cases in which t has small values.

An interesting application of coding to HF communications is the 2.4 Kilobits/sec Linear Prediction Encoding (LPC) used in the HF links of the NATO network. In this code, the system preamble uses a Bose-Chaudhuri-Hocquenghen (BCH) error correction code of length 252 bits (252,128), while each digital voice's word is encoded in a (24,12) block code.

## G. CONCLUSIONS

The use of coding in HF communications is expected to gradually expand from present-day limited application to a more general use, following the progress in microprocessor technology, and the ever greater easiness by which complex circuitry is mechanized.

This introductory lecture was aimed at providing a balanced view of the potential usefulness of codes in HF communications, vis-a-vis the complexity of the required circuitry: this is the real dilemma of these techniques. The material of this lecture, complemented with readings from the bibliographic references given in Section H, should be sufficient introduction to coding, and enough preparation for understanding other lectures of this Series on coding apparatus, and for approaching with increased confidence coding problems of the HF communications practice.

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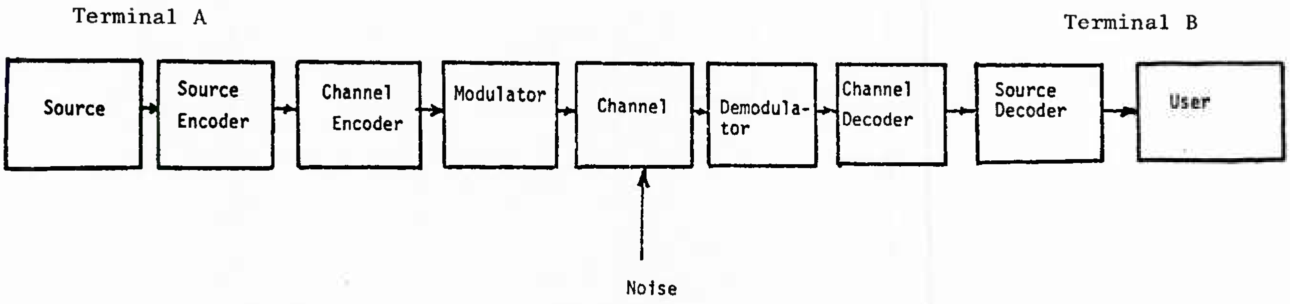


Figure 1 - Block Diagram of a Communications Link that uses Coded Transmission

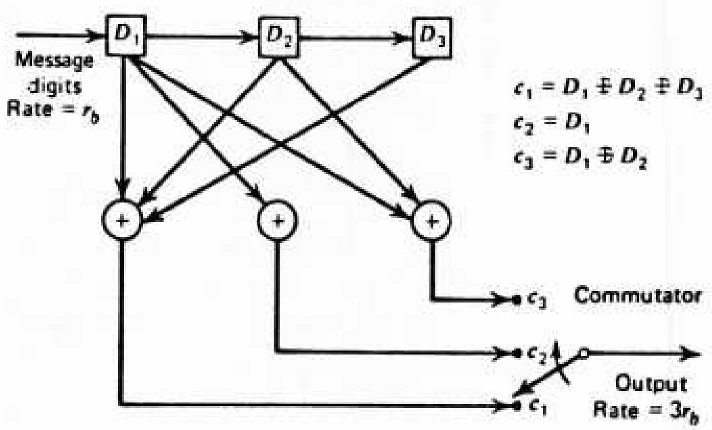


Figure 2 - An example of Convolutional Encoder, with N =3, n=3, k =1 ( from Shanmugan (1979), [4])

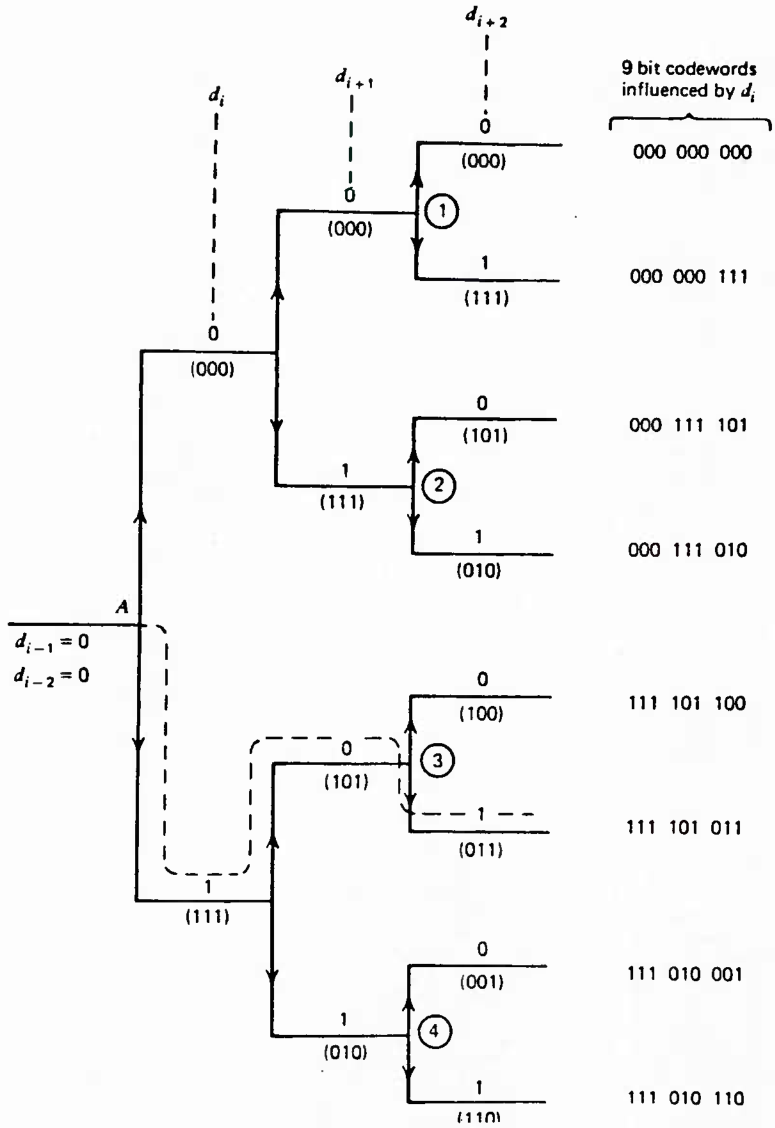


Figure 3 - Code Tree for the Convolutional Encoder of Figure 2 ( from Shanmugan (1979), [4] )

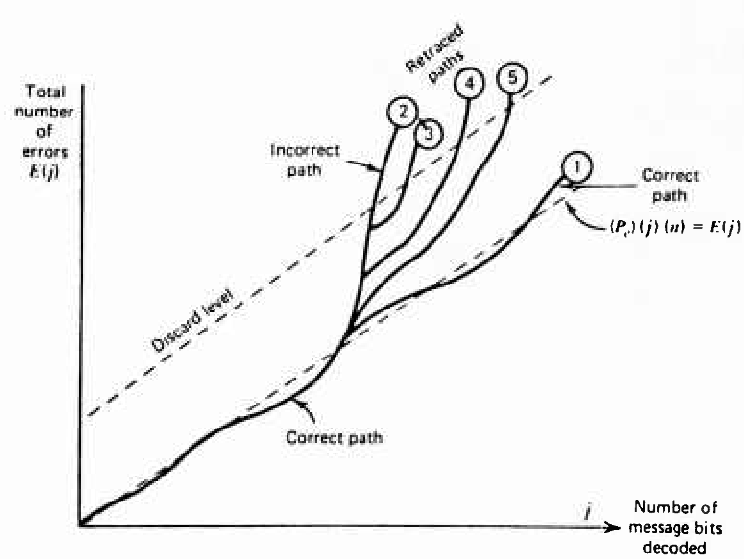


Figure 4 - Threshold Setting in Sequential Decoding ( $P_e$ = Probability that a Received Bit is in error)
( from Shanmugan (1979), [4] )

# MODERN HF COMMUNICATIONS, MODULATION AND CODING

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SUMMARY

Digital communications over High Frequency radio channels are limited by the effects of time varying multipath signals and an impulsive noise characteristic. Modulation techniques which utilize adaptive receiver structures in conjunction with advanced error correction coding concepts can provide quality communication at both low data rates around 100 bps and high data rates around 2400 bps. In this tutorial paper, the multipath channel is examined and two basic constraints are introduced. When the learning constraint is satisfied, it is possible to estimate the channel multipath gain and phase components. The diversity constraint establishes the necessary condition for implicit diversity. For low data rate applications where Intersymbol Interference (ISI) is negligible, adaptive receivers are discussed when both learning and diversity constraints are satisfied. An incoherent adaptive receiver is discussed for applications where the learning constraint is not satisfied. For high data rate applications, both constraints are satisfied in an HF application but ISI effects are severe. Adaptive techniques including equalizers and maximum likelihood sequence estimators are discussed. Experimental results from HF channel simulator tests are presented in a comparison of nonadaptive and adaptive high speed modems. We discuss the use of error correction coding to protect against impulsive noise and show that multipath fading reduces the theoretical performance capability by only 1 to 3 dB when proper coding and interleaving are employed. Performance of practical coding schemes using channel state information are used to show the potential performance on an HF channel.

## 1. INTRODUCTION

HF radio uses frequencies in the range from 2 to 30 MHz. At these frequencies, communications beyond line-of-sight is due to refractive bending of the radio wave in the ionosphere from ionized layers at different elevations. In most cases more than one ionospheric "layer" causes the return of a refracted radio wave to the receiving antenna. The impulse response of such a channel exhibits a discrete multipath structure. The time between the first arrival and the last arrival is the multipath maximum delay difference. Changes in ion density in individual layers due to solar heating cause fluctuations in each multipath return. This time varying multipath characteristic produces alternately destructive and constructive interference. In addition to these channel variations, the received signal includes additive noise which may be impulsive in nature and it may also include interference from other users.

To provide successful communications under these sets of conditions requires sophisticated receivers and the use of error correction coding techniques. This paper discusses adaptive receiver structures for the HF channel and summarizes some of the more important error correction coding considerations. In Section 2 the important characteristics of the fading multipath channel are defined including two constraints which are critical for receiver adaptivity and for extraction of diversity from the multipath structure. In Section 3 adaptive receivers are discussed for both the negligible and significant intersymbol interference conditions. Both coherent and incoherent receivers are also treated. In Section 4 some recent results from experiments on high speed HF modems are summarized. Error correction coding considerations are presented in Section 5.

## 2. FADING MULTIPATH CHANNELS

For digital communication over HF radio links, an attempt is made to maintain transmission linearity, i.e., the receiver output is a linear superposition of the transmitter input plus channel noise. This is accomplished by operation of the power amplifier in a linear region or with saturating power amplifiers using constant envelope modulation techniques. For linear systems, multipath fading can be characterized by a transfer function of the channel $H(f,t)$. This function is the two dimensional random process in frequency $f$ and time $t$ that is observed as carrier modulation at the output of the channel when sine wave excitation at the carrier frequency is applied to the channel input. For any continuous random process, we can determine the minimum separation required to guarantee decorrelation with respect to each argument.

For the time varying transfer function $H(f,t)$ let $t_d$ and $f_d$ be the decorrelation separations in the time and frequency variables, respectively. If $t_d$ is a measure of the time decorrelation (coherence time) in seconds, then

$$\sigma_t = \frac{1}{2\pi t_d} \text{ Hz}$$

is a measure of the fading rate or bandwidth of the random channel. The quantity $\sigma_t$ is often referred to as the Doppler spread because it is a measure of the width of the received spectrum when a single sine wave is transmitted through the channel. The dual relationship for the frequency decorrelation $f_d$ (coherence bandwidth) in Hz suggests that a delay variable pass system of bandwidth B, the sampling theorem states that any linear filter can be represented by parallel TDL filters operating on each quadrature carrier component with a tap spacing of 1/B or less. The optimum receiver can then be realized by a cascade of two such parallel TDL Quadrature filters: one with tap weights adjusted to form the noise filter, the second with tap weights adjusted to form the matched filter. Since the cascade of two bandlimited linear filters is another bandlimited filter, in some applications it is more convenient to employ one TDL to realize R(f) directly. In practice, signals cannot be both time and frequency limited so that these TDL filters can only approximate the ideal solution. One advantage of the TDL filter is the convenience in adjusting the tap weight voltage as a means of tracking the channel and noise spectrum variations.

The optimum receiver requires knowledge of the noise power spectrum K(f) and the channel transfer function H(f). When K(f) is not flat over the band of interest, the input noise process contains correlation which is to be removed by the noise filter. Techniques to reduce correlated noise effects include (1) prediction of future noise values and cancellation of the correlated component, (2) mean square error filtering techniques using an appropriate error criterion, and (3) noise excision techniques where a Fast Fourier Transform is used to identify and excise noise peaks in the frequency domain. The problem of noise filtering is usually important in bandwidth expansion systems because of interference from other users as well as hostile jamming threats.

The realization of the matched filter depends on the amount of time coherence. When the learning constraint $R \gg \sigma_t$ is satisfied, there is enough time to learn the channel and the matched filter structure and subsequent detection process can be accomplished coherently, i.e., with known amplitude and phase for each multipath return. As R decreases relative to $\sigma_t$, a degradation due to noisy estimates increases. When the learning constraint is not satisfied, incoherent detection must be used to avoid this noisy estimate problem. We consider these two conditions separately.

### 3.1.1 Coherent Reception, Learning Constraint Satisfied

When there is enough time coherence to learn the channel characteristics, an important realization of the matched filter, a RAKE TDL filter, using the concepts developed by Price and Green [3], can be used to adaptively derive an approximation to $H^*(f)$. A RAKE filter is so named because it acts to "rake" all the multipath contributions together. This can be accomplished using the TDL filter shown in Figure 2, where the TDL weights are derived from a correlation of the tap voltages with a common test sequence, i.e., S(t). This correlation results in estimates of the equivalent TDL channel tap values. Complex notation is used here in Figure 2 to represent the quadrature components for a bandpass process. The tap values are complex in this representation. By proper time alignment of the test sequence the RAKE filter tap weights become estimates of the channel tap values but in inverse time order as required in a matched filter design. For adaptation of the RAKE filter, the test sequence may be either a known sequence multiplexed in with the modulated information or it may be receiver decisions used in a decision-directed adaptation.

An alternate structure for realizing the matched filter is a re-circulating delay line which forms an average of the received pulses. This structure was proposed as a means of reducing complexity in a RAKE filter design [4] for a frequency-shift-keying system. For a Pulse Amplitude Modulation system the structure would take the form shown in Figure 3. An inverse modulation operation between the input signal and a local replica of the signal modulation is used to strip the signal modulation from the arriving signal. The re-circulating delay line can then form an average of the received pulse which is used in a correlator to produce the matched filter output. This correlation filter is considerably simpler than the RAKE TDL filter shown in Figure 2.

In both the RAKE TDL and correlation filter, an averaging process is used to generate estimates of the received signal pulse. Because this signal pulse is imbedded in receiver noise it is necessary that the measurement process realize sufficient signal-to-noise ratio. This fundamental requirement is the basis for the learning constraint

$$R(b/s) \gg \sigma_t \text{ (Hz)}$$

introduced earlier. If the signal rate R from which adaptation is being accomplished is not much greater than the channel rate of change $\sigma_t$ then the channel will change before the averaging process can build up sufficient signal-to-noise ratio for an accurate measurement. This requirement limits the application of adaptive receiver techniques with implicit frequency diversity gain to slow fading applications relative to the data rate. In many HF channel applications the fade rates are on the order of Hz and data requirements are hundreds of times larger.

### 3.1.2 Incoherent Reception, Learning Constraint Not Satisfied

For data rate applications on the order of tens of bits/ second, the data rate R may not be large enough relative to the Doppler spread $\sigma_t$ to accurately estimate the channel gain and phase values for coherent detection. This condition is even more likely in aircraft communication or auroral HF transmissions both of which have increased Doppler spread values. When the phase information cannot be estimated, the communicator generally selects an orthogonal set of waveforms rather than the coherent choices of antipodal or quadrature antipodal. Although binary orthogonal signaling has a 3 dB poorer distance between transmitted signals than binary antipodal, this distance loss relative to binary antipodal signaling is improved by using more orthogonal waveforms. For M-ary orthogonal signaling in a nonfading channel application, the bit error probability can be made to exponentially tend to zero when $E_b/N_0 > 2(\ln 2)$ [5]. The optimum receiver however requires a matched filter for each orthogonal waveform so there is a significant complexity disadvantage as M increases.

On a fading channel, more waveform choices implies a form of diversity. However in this application there is a performance limit to M because unlike the coherent detection case, performance does not monotonically improve with increased diversity. In a noncoherent system, a square law combiner structure is used to sum the effective diversity signals. Qualitatively when the bit energy per diversity is not somewhat larger than the additive noise, a signal suppression effect results. For a fixed symbol energy, more diversity implies a smaller value of bit energy per diversity. Quantitatively, Pierce [6] has shown that the optimum diversity in an incoherent system is approximately

$$D_{opt} \doteq \frac{E_b}{3N_0}$$

In an HF channel application, the effective diversity order is a function of the channel statistics. In the limit of large signal-to-noise ratios incoherent systems are poorer than coherent systems by about 5.3 dB [6] when the incoherent system operates at the optimum diversity, but for other channel conditions the difference will be larger. For this reason, it is always desirable from a performance viewpoint to use coherent systems when the learning constraint is satisfied.

The optimum incoherent receiver in an M-ary orthogonal signaling scheme requires a noise filter followed by a matched filter for each of the M-waveform possibilities. The outputs of the M matched filters can be used to form a hard decision or as a set of soft decision inputs to the decoder. Again a tapped delay line filter is used in the matched filter to isolate the various multipath returns. Figure 4 illustrates a TDL matched filter for one of the M-waveforms. At each tap of the TDL, correlation is performed with the particular waveform for this matched filter. The correlation operation integrates over the waveform symbol period. The magnitude squared of the correlator output at each multipath delay forms the set of sufficient statistics for this particular waveform hypothesis. These statistics are weighted by attenuator values which are a function of the multipath power vs. delay profile. This profile can be estimated by averaging the correlator output for a particular tap from all matched filters over a period of time for which the multipath profile remains fixed.

### 3.1.3 Summary of Small-ISI Adaptive Receiver

The receiver for this small-ISI example has in general a noise filter to accentuate frequencies where noise power is weakest, and a matched filter structure which coherently recombines the received signal elements to provide the implicit diversity gain. The implicit diversity can be viewed as a frequency diversity because of the decorrelation of received frequencies. The matched filter in this view is a frequency diversity combiner which combines each frequency coherently or incoherently.

An important application of this low data rate system is found in jamming environments where excess bandwidth is used to decrease jamming vulnerability. In more benign environments, however, many communication requirements do not allow for a large bandwidth relative to the data rate and if implicit diversity is to be realized in these applications, the effects of intersymbol interference must be considered.

## 3.2 HIGH DATA RATE RECEIVERS

When the transmitting symbol rate is on the order of the frequency decorrelation interval of the channel, the frequencies in the transmitted pulse will undergo different gain and phase variations resulting in reception of a distorted pulse.

Although there may have been no intersymbol interference (ISI) at the transmitter, the pulse distortion from the channel medium will cause interference between adjacent samples of the received signal. In the time domain, ISI can be viewed as a smearing of the transmitted pulse by the multipath thus causing overlap between successive pulses. The condition for ISI can be expressed in the frequency domain as

$$T^{-1}(\text{Hz}) \gtrsim f_d(\text{Hz})$$

or in terms of the multipath spread

$$T(\text{sec}) \lesssim 2\pi\sigma_f \ (\text{sec}) \quad .$$

Since the bandwidth of a digital modulated signal is at least on the order of the symbol rate $T^{-1}$Hz, there is no need for bandwidth expansion under ISI conditions in order to provide signal occupancy of decorrelated portions of the frequency band for implicit diversity. However it is not obvious whether the presence of the intersymbol interference can wipe out the available implicit diversity gain. Within the last decade it has been established that adaptive receivers can be used which cope with the intersymbol interference and in most practical cases wind up with a net implicit diversity gain. These receiver structures fall into three general classes: correlation filters with time gating, equalizers, and maximum likelihood detectors.

### 3.2.1 Correlation Filters

These filters approximate the matched filter portion of the optimum no ISI receiver. The correlation filter shown in Figure 3 would fail to operate correctly when there is intersymbol interference between received pulses because the averaging process would add overlapped pulses incoherently. When the multipath spread is less than the symbol interval, this condition can be alleviated by transmitting a time gated pulse whose "off" time is approximately equal to the width of the channel multipath. The multipath causes the gated transmitted pulse to be smeared out over the entire symbol duration but with little or no intersymbol interference. The correlation filter can then be used to match the received pulse and provide implicit diversity [7]. In a configuration with both explicit and implicit diversity, moderate intersymbol interference can be tolerated because the diversity combining adds signal components coherently and ISI components incoherently. Because the off-time of the pulse can not exceed 100%, this approach is clearly data rate limited for fixed multipath conditions. Because of this data rate limitation, this type of ISI receiver has had little application in HF systems.

### 3.2.2 Adaptive Equalizers

Adaptive equalizers are linear filter systems with electronically adjustable parameters which are controlled in an attempt to compensate for intersymbol interference. Tapped Delay Line filters are a common choice for the equalizer structure as the tap weights provide a convenient adjustable parameter set. Adaptive equalizers have been widely employed in telephone channel applications [8] to reduce ISI effects due to channel filtering. In a fading multipath channel application, the equalizer can provide three functions simultaneously: noise filtering, matched filtering for explicit and implicit diversity, and removal of ISI. These functions are accomplished by adapting a Tapped Delay Line Equalizer (TDLE) to force some error measure to a minimum. By selecting the error measure such that it contains correlated noise effects, improper diversity combining and filtering effects, and ISI, the TDLE will minimize the combined effects.

A Linear Equalizer (LE) is defined as an equalizer which linearly filters each of the N explicit diversity inputs. An improvement to the LE is realized when an additional filtering is performed upon the detected data decisions. Because it uses decisions in a feedback scheme, this equalizer is known as a Decision-Feedback Equalizer (DFE). The optimality of the DFE under a mean square error criterion and a means of adapting the DFE in a fading channel application have been derived in [9].

The operation of a matched filter receiver, a Linear Equalizer, and a Decision Feedback Equalizer can be compared from examination of the received pulse train example of Figure 5. The binary modulated pulses have been smeared by the channel medium producing pulse distortion and interference from adjacent pulses. Conventional detection without multipath protection would integrate the process over a symbol period and decide a +1 was transmitted if the integrated voltage is positive and -1 if the voltage is negative. The pulse distortion reduces the margin against noise in that integration process. A matched filter correlates the received waveform with the received pulse replica thus increasing the noise margin. The intersymbol interference will arise from both future and past pulses when matched filtering to the channel is employed. The ISI can be compensated for a linear equalizer by using properly weighted time shifted versions of the received signal to cancel future and past interferers. The decision-feedback equalizer uses time shifted versions of the received signal only to reduce the future ISI. The past ISI is cancelled by filtering past detected symbols to produce the correct ISI voltage from these interferers.

A general Decision-Feedback Equalizer receiver is shown in Figure 6. The first two filters, the noise and matched filters, are the small ISI receiver discussed in the previous subsection. The noise filter reduces correlated noise effects and the matched filter coherently combines the multipath returns. The forward filter minimizes the effect of ISI due to future pulses, i.e., pulses which at that instant have not been used to form receiver decisions. After detection, receiver decisions are filtered by a Backward Filter to eliminate intersymbol interference from previous i.e., past, pulses. Because the Backward Filter compensates for this "past" ISI, the Forward Filter need only compensate for "future" ISI.

When error propagation due to detector errors is ignored, the DFE has the same or smaller mean-square-error than the LE for all channels [9]. The error propagation mechanism has been examined by a Markov chain analysis [10] and shown to be negligible in practical fading channel applications.

A significant problem in the use of adaptive equalizers in the HF channel application is the requirement for rapid tracking of the adaptive equalizer weights. Simple estimated gradient techniques sometimes called Least Mean Squares algorithms are not fast enough because of the correlation between signals on the equalizer taps. Because of this correlation when the algorithm attempts to adjust one tap, noise fluctuations result in other taps. This problem can be eliminated through the use of Kalman filtering techniques [11] or Lattice filter structures [12].

### 3.2.3 Maximum Likelihood Detectors

Since the DFE minimizes an analog detector voltage, it is unlikely that it is optimum for all channels with respect to bit error probability. By considering intersymbol interference as a convolutional code defined on the real line (or complex line for bandpass channels), maximum likelihood sequence estimation algorithms have been derived [13,14]. These algorithms provide a decoding procedure for receiver decisions which minimize the probability of sequence error. A Maximum Likelihood Sequence Estimator (MLSE) receiver still in general requires a noise filter and matched filter although these functions can be imbedded in the estimator structure. Figure 7 illustrates the filtering and sampling functions which precede the MLSE. The estimation techniques used to derive the noise and matched filter parameters also provide an estimate of the discrete sample response used by the MLSE decoder to resolve the intersymbol interference. An implementation of an MLSE receiver has been described by Crozier, et. al [15].

The MLSE algorithm works by assigning a state for each intersymbol interference combination. Because of the one-to-one correspondence between the states and the ISI, the maximum likelihood source sequence can be found by determing the trajectory of states.

If some intermediate state is known to be on the optimum path, then the maximum likelihood path originating from that state and ending in the final state will be identical to the optimal path. If at time n, each of the states has associated with it a maximum likelihood path ending in that state, it follows that sufficiently far in the past the path history will not depend on the specific final state to which it belongs. The common path history is the maximum likelihood state trajectory [13].

Since the number of ISI combinations and thus the number of states is an exponential function of the multipath spread, the MLSE algorithm has complexity which grows exponentially with multipath spread. The equalizer structure exhibits a linear growth with multipath spread. In return for this additional complexity, the MLSE receiver results in smaller (sometimes zero) intersymbol interference penalty for channels with isolated and deep frequency selective fades.

## 5. HIGH DATA RATE HF MODEM PERFORMANCE

A series of tests on parallel tone modems by Watterson [16], and on two serial tone modems provide a basis for modem performance comparison for a high data rate HF channel application.

The parallel tone modems divide the data into low rate parallel sub-channels so that nonadaptive techniques can be used and ISI effects can be avoided. The serial tone approaches use PSK transmissions and some form of decision feedback equalization in an adaptive receiver structure.

### 5.1 TEST CASES FOR COMPARISON

There is considerable performance data available for the flat fading (single path) and the dual fading path suggested by Watterson [16]. This latter channel has two "skywave" returns which are spaced by 1 millisecond and they each fade with a $2\sigma$ Doppler spread of 1.0 Hz. In general the more echoes, the easier it is for the equalizer to track because it is less likely that all echoes will simultaneously become small. The Watterson two path channel has thus become a good standardized test case. Watterson tested conventional parallel tone modems on this channel and the results are given in his report [16]. This channel was also used in an evaluation of a Harris Corporation modem under USAF Contract No. F30602-81-C-0093 and a SIGNATRON modem under USAF Contract No. F30602-80-C-0296. The results from the final reports from these contracts provide a comparison of present day serial tone modem technology with the parallel tone modem approaches tested by Watterson.

All of the modems in this comparison are uncoded except for the parallel tone MX-190 modem which used a rate 16/25 block code. Although the simulators used on some of the tests were not exactly the same, no appreciable performance difference is anticipated from this factor. The fairest comparison is on a peak power basis because HF radios are peak power limited. For this reason we use peak bit energy $E_{pb}$ rather than average bit energy $E_b$. The peak-to-average ratios assumed for the modem comparison were

| | | |
|---|---|---|
| SIGNATRON (SIG) Modem: | 1 dB | SERIAL TONE |
| Harris (HRS) Modem: | 1 dB | |
| USC-10: | 7 dB | PARALLEL TONE |
| ACQ-6: | 7 dB | |
| MX-190: | 7 dB | |

### 5.2 MODEM COMPARISONS

The flat fading channel results are shown in Figure 8. In the flat fading tests, the two serial modems compared perform almost the same and both gain a large number of dB over the parallel tone modems. Most of this improvement is due to the peak-to-average ratio. The frequency selective fading results are shown in Figure 9. In these tests the Harris and SIGNATRON modem realize almost exactly the same implicit diversity when the fading is slow. The performance advantage of these modems over the parallel tone modems is very large under these conditions.

In comparison with parallel tone modems under the faster fading conditions, both the SIGNATRON and Harris modem still have a significant performance gain even over the coded MX-190 system. Both the Harris and SIGNATRON modems have not included error correction coding.

These results indicate the large performance advantage that can be realized when adaptive receiver structures are used to exploit the multipath structure of the HF channel. This fact has been well known for some time for low data rate applications where ISI is not too severe but these recent results establish that it holds in high data rate applications as well.

## 6. ERROR CORRECTION CODING

Digital transmission over fading channels is impaired by additive channel noise and a fluctuating received signal level which results in poor signal-to-noise ratio some fraction of the time. When the fading is slow with respect to the data transmission rate, it is usually possible to determine the channel parameters, i.e., instantaneous received signal power and additive noise power. Under these known channel conditions, one can do almost as well as a non fading channel by using interleaving to span the fade intervals. On HF radio channels the additive noise can be impulsive in nature. Since the impulsive noise durations are much shorter than fade intervals, burst error correction coding or relatively short interleaver techniques provide protection against the noise effects. In this review, we will concentrate on the improvement realized against multipath fading effects.

### 6.1 INFINITE INTERLEAVING RESULT, A REVIEW

In a slow fading channel application typical of HF radio communication where it is possible to track the channel state information, the loss in detection efficiency due to the fading is only a few dB provided that the channel is memoryless. Ericson [17] established this result in a comparison of exponent bounds and capacities for the fading and nonfading white Gaussian noise channel. Of course if the fading is slow enough to allow tracking of the channel state information, it will not be true in general that the channel is memoryless, i.e., received symbols will not be independent. If delay in message detection can be tolerated, interleaving of the transmitted symbols and de-interleaving of the received samples and the channel state information can provide a memoryless channel condition with known channel parameters. To show that the detection efficiency under this condition has only a small degradation due to the fading, consider the exponent bound parameter [18] for the discrete memoryless channel which has binary input u and real number output y.

$$R_0 = -\log_2 \sum_x [q(u) \sum_y \sqrt{p(y|u)}]^2 \quad . \tag{1}$$

When the rate r in bits per channel usage is less than $R_0$, the bit error probability is upper bounded by an exponentially decreasing function of the block length for a block code or the constraint length for a convolutional code. For binary sources and Gaussian noise channels, $R_0$ is maximized when the source distribution q(u) is equally likely and the exponent bound parameter becomes

$$R_0 = 1 - \log_2(1 + Z) \tag{2}$$

where Z takes one of two forms depending on whether the channel is fading or not. For nonfading conditions, the y integration in (1) is only over the Gaussian noise density and one obtains the well known result for received energy per information bit $E_b$ and noise spectral density $N_0$ watts/Hz,

$$Z = e^{-rE_b/N_0} \qquad \text{Gaussian Noise, No Fading} \tag{3}$$

For fading conditions, the y integration in (1) also requires integration over the density of fading signal strength if it is assumed that this parameter is known for each received y sample, i.e., de-interleaving of the channel state information is used at the receiver. Thus, one finds

$$Z = \int_0^\infty p(s)e^{-rE_b s/N_0}ds \qquad \text{Gaussian Noise, General Fading} \tag{4}$$

where s is a unit mean random variable for the received power on a fading channel. For a Rayleigh fading received envelope, the power is exponentially distributed which gives the Rayleigh fading result

$$Z = (1 + rE_b/N_0)^{-1} \qquad \text{Gaussian Noise, Rayleigh Fading} \tag{5}$$

where $E_b$ is now the average received bit energy. For a fixed rate r and a constant $R_0$, the dB loss due to fading is shown in Figure 10 to increase from about 1dB to 3dB as the signal-to-noise ratio increases.

## 6.2 IMPORTANCE OF THE DELAY CONSTRAINT

In many communication systems it is unacceptable to introduce delay which is many times longer than the fade epoch. In speech communications for example, delay must be less than a few tenths of a second whereas HF radio fading channels with speech applications have fade intervals on the order of a second. This condition precludes the use of interleaving to produce the memoryless channel which is required to achieve the detection efficiency of Figure 10. Without interleaving with an average probability of bit error as the performance criterion, error correction coding does not even look attractive. Say we had a code which achieved capabity on the Gaussian noise channel. For this code the probability of bit error is zero when $sE_b/N_0$ is greater than 0.69 (-1.6 dB) and 1/2 when it is less than this value, The average bit error rate is

$$\bar{p} = \frac{1}{2} \int_0^{0.69\, N_0/E_b} p(s)ds = \frac{1}{2}\, \frac{1}{1 + E_b/0.69\, N_0}, \quad \text{Capacity Code} \qquad (6)$$

For uncoded coherent phase shift keying (CPSK) operating under the same Rayleigh fading conditions, one has for large $E_b/N_o$,

$$\bar{p} = \frac{1}{2} \int_0^{\infty} \operatorname{erfc}(\sqrt{sE_b/N_0})\, p(s)ds \doteq \frac{N_0}{4E_b}, \quad \text{Uncoded CPSK} \qquad (7)$$

We obtain the rather dismal result that for large signal-to-noise ratio this ideal capacity code is 1.4 dB worse than an uncoded CPSK system. The average probability of error criterion results in a large penalty when a deep fade occurs. For this reason communication systems have historically used diversity, i.e., the trival redundant code, rather than coding when the delay constraint was present. The loss in detection efficiency due to fading under this delay constraint and this performance measure is then very large. Thus, it is important to evaluate the utility of average error rate as a performance measure.

In many applications information transfer is satisfactory if the bit error rate is less than some threshold and it is unsatisfactory if the bit error rate is any amount larger than this threshold. For these applications, average probability of error is an inappropriate performance measure. The concept of outage probability, i.e., the fraction of time that the bit error rate is greater than a threshold, is becoming widely accepted as a more meaningful performance measure for communications. Under this concept the power gain due to coding in a fading channel is exactly the power gain at the threshold value on a nonfading channel. Even though the coded system may be poorer than an uncoded system under deep fade conditions, large coding gains can be realized at bit error rate thresholds where practical communication takes place. Thus error correction coding under appropriate performance measures can provide significant power gain even when the delay constraint precludes interleaving to span the fade epochs.

## 6.3 PRACTICAL CODING RESULTS

Coding gains for practical error correction coding techniques on HF radio channels must be examined relative to a delay constraint which allows or precludes interleaving over the fading epochs. These fading epochs are on the order of 1 second.

### 6.3.1 Delay Constraint Present

Speech and certain digital data communication systems fall into this class. Under this constraint, error correction coding and modest interleaving will protect against impulsive noise. However no significant coding gain against fading is realizable under an average probability of error criterion. Under an outage probability criterion the coding gain is exactly the coding gain realized on a nonfading channel at the Bit Error Rate (BER) threshold where the outage is determined. An off the shelf 1/2 rate convolutional decoder [19] with constraint length 7 gives a coding gain of 4.6 dB at a threshold BER of $10^{-4}$. This coding gain drops to 3.8 dB at the BER threshold of $10^{-3}$. Concatenated codes [20] which use a convolutional code for the inner code and a Reed-Solomon code for the outer code can do significantly better. Since this delay constraint condition reduces to an evaluation of coding gain on the <u>non</u>-fading channel, only a limited discussion is warranted here.

### 6.3.2 No Delay Constraint

We have shown that theoretical performance on the fading channel can come within 1 to 3dB of the non-fading channel if we interleave the channel state information as well as the soft decision outputs from the modem. It is of interest to examine practical coding schemes to determine realistic coding gains with this approach. Hagenauer [21] has computed the performance of a rate 1/2, constraint length 7, convolutional code for different quantitization levels of channel state information and decision values. The following nomenclature is used in the comparison.

Unquantized Decision = Y SOFT (YS)
1 bit quantized Decision = Y HARD (YH)
Unquantized Channel State = A SOFT (AS)
1 bit quantized Channel State = A HARD (AH)
No Channel State Information = A NO (AN)

The result of his calculations are reproduced in Figure 11.

This code has a 4.6 dB gain at a BER of $10^{-4}$ on the non-fading channel. It requires an $E_b/N_0$ of about 7.2 dB under fading conditions when both decisions and channel state are unquantized. This represents a coding gain over the non-fading channel of about 1.2 dB. The loss due to the fading is then only 3.4 dB. This small loss can be better appreciated by comparing the coded performance with dual diversity in Figure 11. It is also of interest to note that one bit quantization on the decisions and channel state does as well as unquantized decisions alone.

## 7. SUMMARY

For many HF communications applications, the channel may be considered to be slowly fading. We have shown here that when this slow fading is exploited to learn the channel, combined modulation and coding techniques can remove much of the fading effect even in high data applications where intersymbol interference is a serious problem. Experimental results have been presented showing successful communication using equalizer modems in a high data rate application. Fading performance of practical decoding techniques has been shown to be within a few dB of coded systems on non-fading channels.

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FIGURES

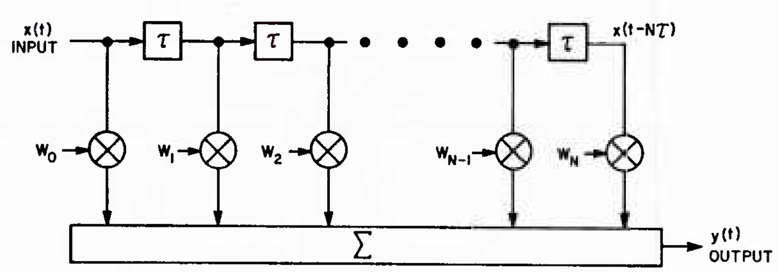


FIGURE 1 TAPPED-DELAY LINE (TDL) FILTER

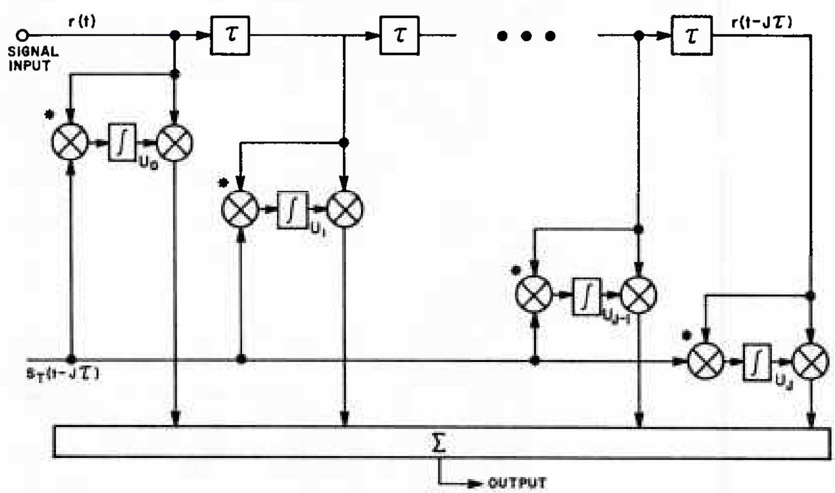


FIGURE 2 RAKE FILTER, COHERENT SYSTEM

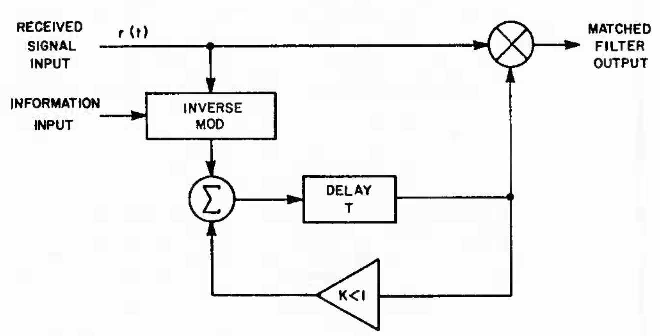


FIGURE 3 CORRELATION FILTER FOR PAM SYSTEM, COHERENT SYSTEM

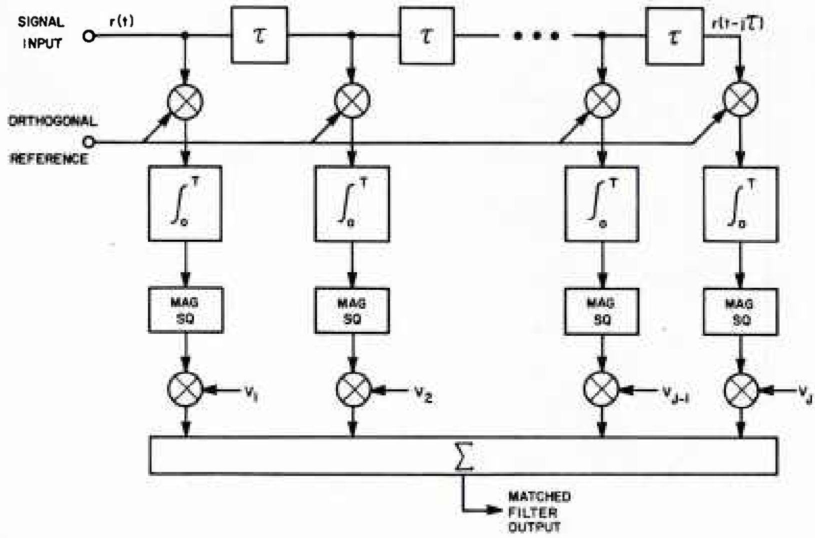


FIGURE 4 ORTHOGONAL SIGNALING MATCHED FILTER, INCOHERENT DETECTION

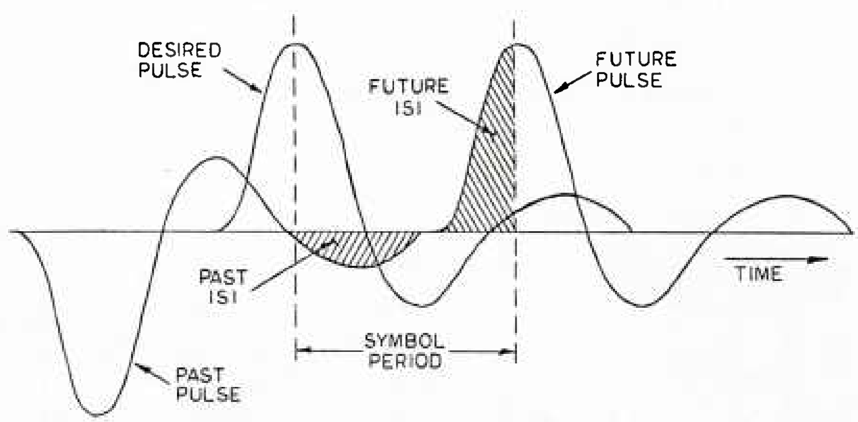


FIGURE 5 RECEIVED PULSE SEQUENCY AFTER CHANNEL FILTERING

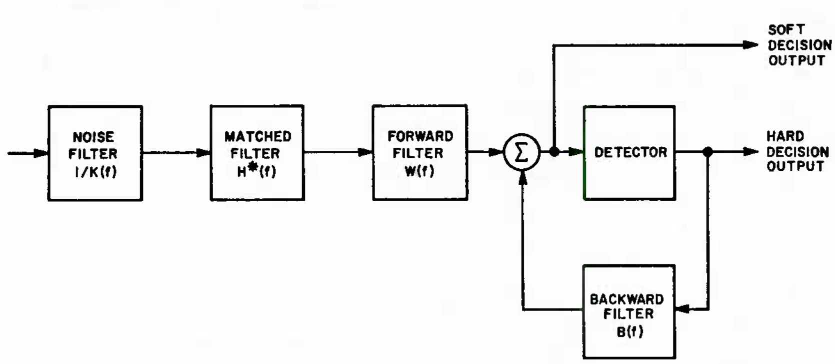


FIGURE 6 DECISION-FEEDBACK EQUALIZER RECEIVER

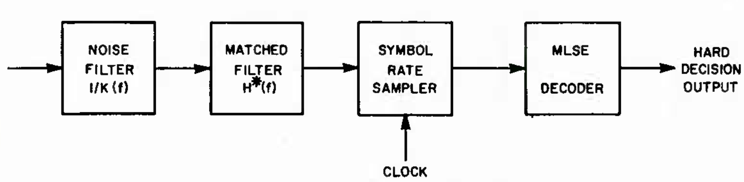


FIGURE 7 MLSE RECEIVER

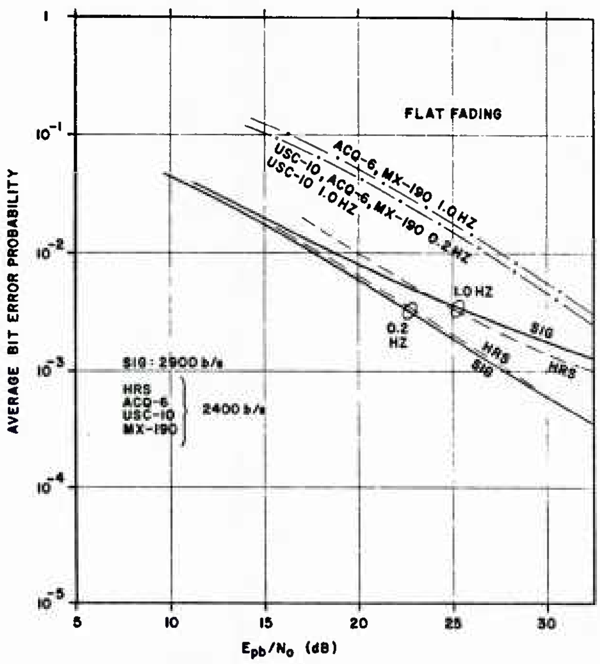


FIGURE 8 FLAT FADING

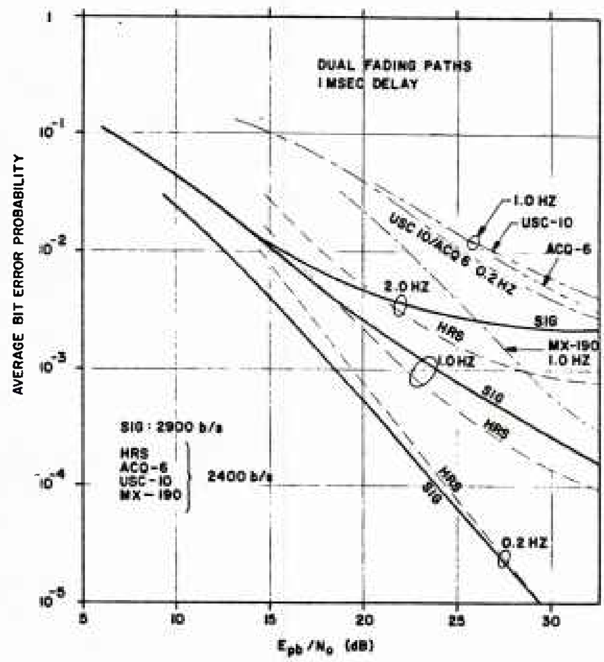


FIGURE 9 WATTERSON TEST CHANNEL

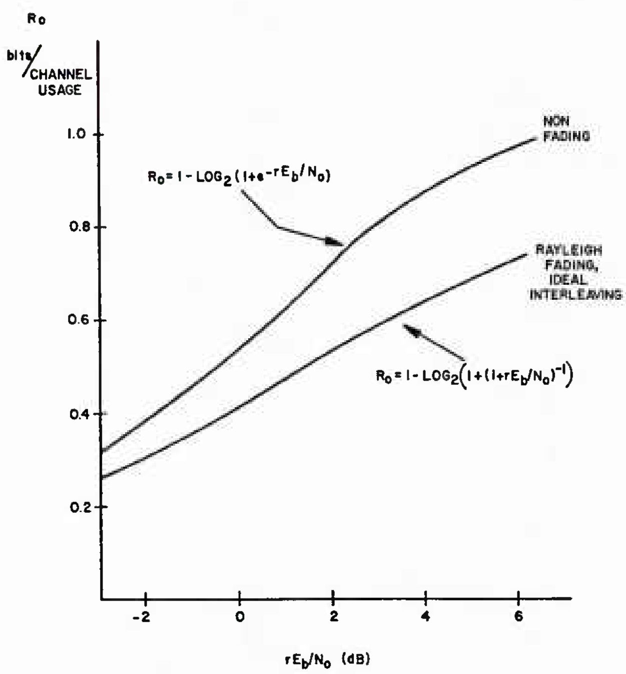


FIGURE 10 LOSS IN DETECTION EFFICIENCY DUE TO FADING

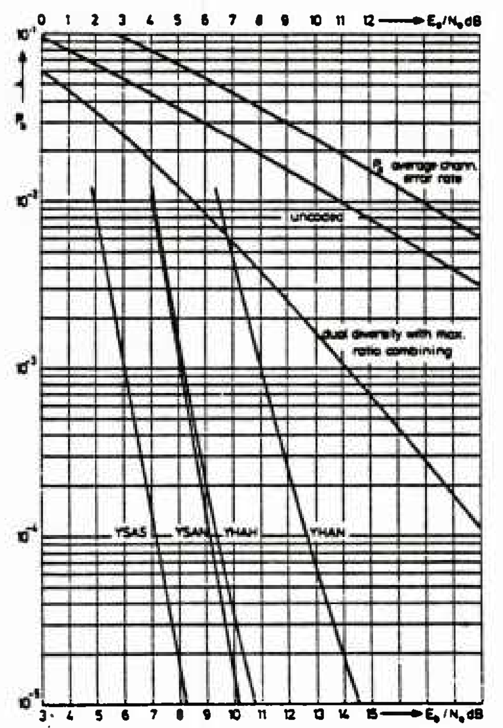


FIGURE 11 BIT ERROR PROBABILITY $P_b$ (UPPER BOUND) AS A FUNCTION OF AVERAGE $E_b/N_0$.

CONV. CODE, RATE = 1/2,
CONSTRAINT LENGTH 7,
SLOWLY FADING RAYLEIGH-CHANNEL FULLY INTERLEAVED, COHERENT BINARY PSK

YHAH (y hard, a hard (BINARY QUANTIZATION)

FROM HAGENAUER [21]

# EQUIPMENT: ANTENNA SYSTEMS

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ABSTRACT

Some antenna fundamentals as well as definitions of the principal terms used in antenna engineering are described. Methods are presented for determining the desired antenna radiation patterns for an HF communications circuit or service area. Sources for obtaining or computing radiation pattern information are outlined. Comparisons are presented between the measured and computed radiation patterns. The effect of the properties of the ground on the antenna gain and pattern are illustrated for several types of antennas. Numerous examples are given of the radiation patterns for typical antennas used on short, intermediate and long distance circuits for both mobile and fixed services operations. The application of adaptive antenna arrays and active antennas in modern HF communication systems are briefly reviewed.

## 1. INTRODUCTION

The performance of many HF communication systems is seriously degraded due to the use of unsuitable antennas for both transmitting and receiving. The antennas should be designed to efficiently radiate and receive energy in the desired directions. A method is described to determine the desired elevation and azimuthal angles of radiation. Sources are presented for obtaining antenna radiation pattern information on typically used HF antennas. The application of prediction techniques is described to determine the best or the most cost effective antenna from several being considered for an HF service. Examples are presented of the reliability of communications for several types of antennas. The advantages and limitations of adaptive antenna arrays and active antennas in modern HF communication systems are briefly described.

## 2. RADIATION ANGLES

Until recently, the desired elevation angles of radiation for a fixed service were calculated manually based on the use of a E region reflecting height of 110 kms and a constant F region height between 250 - 200 kms. Various types of nomograms or charts were used to aid the design engineer in these calculations. At certain times and seasons, the use of a constant F region height yields significant errors in determining the elevation angles of radiation. In addition information on the variance on the elevation angles is required in specifying the beamwidth of the desired antenna lobe pattern. More accurate and additional information can be obtained using the advanced prediction methods described in a previous lecture. The elevation angle is computed for all times, seasons and periods of solar activity over which the HF service is being modelled. The results of these computations are presented in a statistical form as shown in Fig. 1. These computations are based on the selection of the mode of propagation with the least transmission loss. From these results antennas can be specified and types selected that radiate or receive efficiently at the desired elevation angles. For mobile services and other types of services providing coverage over a large geographical area, computations of the elevation angles are done at representative locations and weighted by taking into account the density and location of the communication terminals.

## 3. TYPES OF ANTENNAS

An engineer designing an HF communication system requires information on the radiation pattern of various types of antennas in order to select one that meets the particular needs. Several reports provide graphical information (Dept. of the Army, 1950; Thomas and DuCharme 1974) on the radiation pattern as a function of elevation angle at the centre of the main lobe. In these reports, vertical radiation patterns are presented for antennas typically used for HF communications such as,

a) horizontal half-wave dipole
b) rhombics, type A, B, C, D, and E
c) half rhombic
d) inverted L
e) sloping Vee
f) horizontal log periodic
g) vertical antenna

The antenna gain in a specified direction is given by (Barghausen et al, 1969)

$$G = 10 \log_{10} \left[ 4\pi \frac{\text{Power radiated per unit solid angle}}{\text{Net Power accepted by the antenna}} \right]$$

The net power accepted by the antenna includes ohmic losses, losses in the ground below the antenna and the radiation resistance corresponding to the power radiated from the antenna. The antenna gain does not account for impedance mismatch losses.

The three dimensional pattern for all the above antennas can be calculated using a computer program developed in the U.S.A. (Barghausen et al 1969). Specific current distributions are assumed in the derivation of the equations used to compute the antenna performance. For each antenna, the physical dimensions must be specified as well as the conductivity and dielectric constant of the ground. For example, a rhombic requires specifications for a feed height, leg length, and semi-apex angle while a vertical antenna requires specifications for length of antenna, length of radials, radius of radials, and number of radials.

More sophisticated modelling programs, such as the Numerical Electromagnetic Code (NEC), compute the current distribution on the antenna instead of assumming a specific distribution (Burke and Poggio). With these programs the performance of wire antennas can be computed with great precision but most require the use of a large main frame computer except simplified versions of MININEC which are restrictive in their applications. For example, some versions assume the antenna is erected over perfectly conducting ground.

The type of ground generally has a limited effect on the main lobes of the pattern for horizontally polarized antennas. However for short vertically polarized antennas, the type of ground can significantly alter the efficiency of the antenna and its power gain. The effect of ground on the antenna pattern is shown in Figs. 2 and 3, where for

| | | |
|---|---|---|
| good earth | $\sigma = 10^{-2}$ mho/m | $\epsilon = 4$ |
| poor earth | $\sigma = 10^{-3}$ | $\epsilon = 4$ |
| sea water | $\sigma = 4$ | $\epsilon = 80$ |

A recent survey in Canada indicates the following types of antenna are being used for HF point-to-point communications on short, intermediate and long distance circuits.

| Short Distance 1000 kms | Intermediate Distance 1000 to 4000 kms | Long Distance 4000 kms |
|---|---|---|
| half wave dipole | rhombic | rhombic |
| inverted L | half wave dipole | log periodic |
| vertical | vertical | beverage |
| delta | inverted L | |

Over 80% of the ground based stations used broadband antennas such as the Delta, Rhombic, Beverage, and Log Periodic to enable operation at any frequency in the HF band without a tuning device.

The Log Periodic antenna is used extensively through out the world on high priority communication systems because of its high gain over a broad range of frequencies. Log Periodic antennas are commercially available that use dipole arrays, loop arrays and also spiral configurations. Three versions of Log Periodic dipole arrays are shown in Fig. 4. These types of antennas designed for various communication distances have the following typical gains

| Power Gain (dBi) | Distance (kms) |
|---|---|
| 10 | < 500 |
| 12 | 500-2500 |
| 15 | > 2500 |

A Log Periodic loop array exhibiting an omni-directional radiation pattern is shown in Fig. 5. The antenna is horizontally polarized making its vertical pattern less sensitive to the properties of the ground beneath the antenna. Operation over a wide range of frequencies is achieved by stacking the loops vertically and feeding them in a log periodic manner. The lowest operation frequency is determined by the size and height of the antenna array. The gain is independent of frequency but the vertical pattern is mainly dependent on the height of the antenna.

Travelling wave antennas are used on HF systems since they operate over a broad range of frequencies without an antenna tuning unit. The terminated dipole, sloping Vee, Rhombic, Delta and Beverage are all types of travelling wave antennas. These antennas are terminated in a resistance equal to their characteristic impedance so there is no reflected wave from the end of the antenna. Such antennas must be long in order to radiate efficiently. In order to overcome the low efficiency of the Beverage antenna, several Beverages have been placed in a linear array resulting in an efficient and highly directive antenna (See Fig. 6.).

## 4. ANTENNA SELECTION

Prediction techniques are also used to select the best or most cost effective antenna from several antennas being considered for an HF service. The reliability of communications is computed for each antenna at representative times over which the HF service is planned. For each antenna the average reliability for the representative times is calculated and the best or most cost effective antenna is the one with the maximum average reliability. An example of the average reliabilities for three types of antennas are shown as follows for an HF circuit from Churchill to Inuvik.

| Type of Antenna | Average Reliability |
|---|---|
| Rhombic | 83.5% |
| Horizontal Log Periodic | 78.4 |
| Sloping Vee | 86.1 |

The reliabilities were computed every hour of the day for January, April, July and

October and $R_{12}$=10,40,70 and 100 in order to simulate the HF communications conditions expected from 1981 to 1985. The Sloping Vee antenna exhibits a higher reliability of communications than the other two antennas. In some cases the design engineer may wish information on the statistics of the variability of the reliability for various types of antennas. Such information can also be extracted from prediction program calculations. An example of such information is shown in Fig. 7. The lower values of reliability at certain periods for Antenna #1 compared to Antenna #2 may be of more importance than the average reliability figures.

## 5. MEASURED RADIATION PATTERNS

Measurements of the radiation patterns for HF antennas are reported using an aircraft to tow a small aerodynamically stabilized Xeledop (Barnes 1965; Belrose 1983). The Xeledop consists of a battery operated transmitter and short dipole antenna which is towed in a stabilized position enabling the measurement of patterns for horizontally and vertically polarized antennas. The transmitter frequency is crystal controlled and can be operated in the HF and VHF bands. For certain measurements, a ground based radar is used to measure and control the range between the aircraft and the antenna. With such a measurement facility, the small scale variations of the azimuthal pattern can be measured for both vertically and horizontally polarized antennas. In addition for vertically polarized antenna the small scale variations in the elevation pattern can be measured at specific bearings such as along the boresite of an antenna. Several examples of the measured radiation patterns are shown in Figs. 8,9, and 10 (Petrie 1978; Petrie 1979). In most cases the main lobe of the pattern agrees with the computer calculated patterns to within 2 dB when the electrical properties of the ground are uniform. At elevation angles less than 5 degrees the calculated radiation patterns are at times in error due to variations in the terrain in the vicinity of the antenna. However in the majority of cases, the computer programs provide a good estimate of the main lobe of the antenna but the magnitude and location of minor lobes are not predictable with any degree of accuracy.

## 6. ADAPTIVE ANTENNA ARRAYS

An adaptive antenna array is designed to enhance the reception of desired signals in the presence of jamming and interference. Such systems have been built and tested in a variety of communications applications including mobile, seaborne and airborne platforms (Hanson, 1977; Effinger et al., 1977; Reigler and Compton, 1973; Compton, 1978; Horowitz et al., 1979). The utility of adaptive arrays for communications lies mainly in the ability to reject the undesired signals by virtue of their spatial properties. For a comprehensive basic understanding of adaptive antennas an excellent book has been published recently (Monzingo and Miller, 1980).

Many adaptive arrays operate on the principle of maximizing the signal-to-noise ratio or minimizing the power in the undesired signal at the output of the array by subtracting a reference signal which matches the desired signal from the array output signal. Various techniques are developed to do these complex computations quickly and accurately (Reed et al., 1974; Horowitz, 1980; Widrow et al., 1976). Some of these techniques can be implemented using digital, analogue or a combination of analogue and digital data processing methods. In order to characterize the desired signal so that a reference signal can be generated, a predetermined pilot signal may take many forms: an additive pilot signal, a multiplicative (spread spectrum) code, a time multiplexed pilot signal, and a predetermined set of times when the communication signal is not on. In the presence of a jamming signal it is desirable to code the pilot using a predetermined secure code.

The degree of interference protection afforded by an adaptive array depends on various factors but typical values of protection are about 30 dB. An array of N elements can provide protection against N-1 interfering signals. Adaptive arrays which use the LMS algorithm (Widrow et al., 1967) are reported to adapt to multiple sources of interference in many seconds for a 3kHz signal bandwidth. Arrays processing data using the matrix inversion techniques make efficient use of the signal information and adapt in less than a second for a 3 kHz signal bandwidth. The time varying HF signal environment places certain limitations on the performance of adaptive arrays and requires certain array configurations in order that adequate performance is maintained (Jenkins, 1982).

## 7. ACTIVE ANTENNAS

An active antenna consists basically of a rod or dipole and an RF amplifying device. If the noise level generated at the amplifier output terminal is less than the noise pickup by the antenna, an active antenna system is capable of supplying the same signal-to-noise ratio as a passive antenna. The atmospheric noise when the rod is coupled efficiently to an amplifier. The advantages of an active antenna are

- The short physical dimensions of the rod
- The wide frequency band of operation (ie. 2-30 MHz)
- Fixed output impedance of 50 or 75 ohms

The main disadvantage of the active antenna is the intermodulation distortion products which are often generated by strong interfering signals. The use of VMOS technology in the design of the RF amplifier will improve the dynamic range over which the system can operate with out intermodulation products. However, these distortion products can seriously degrade the performance of an active antenna in an environment with high RF noise levels.

the design of the RF amplifier will improve the dynamic range over which the system can operate with out intermodulation products. However, these distortion products can seriously degrade the performance of an active antenna in an environment with high RF noise levels.

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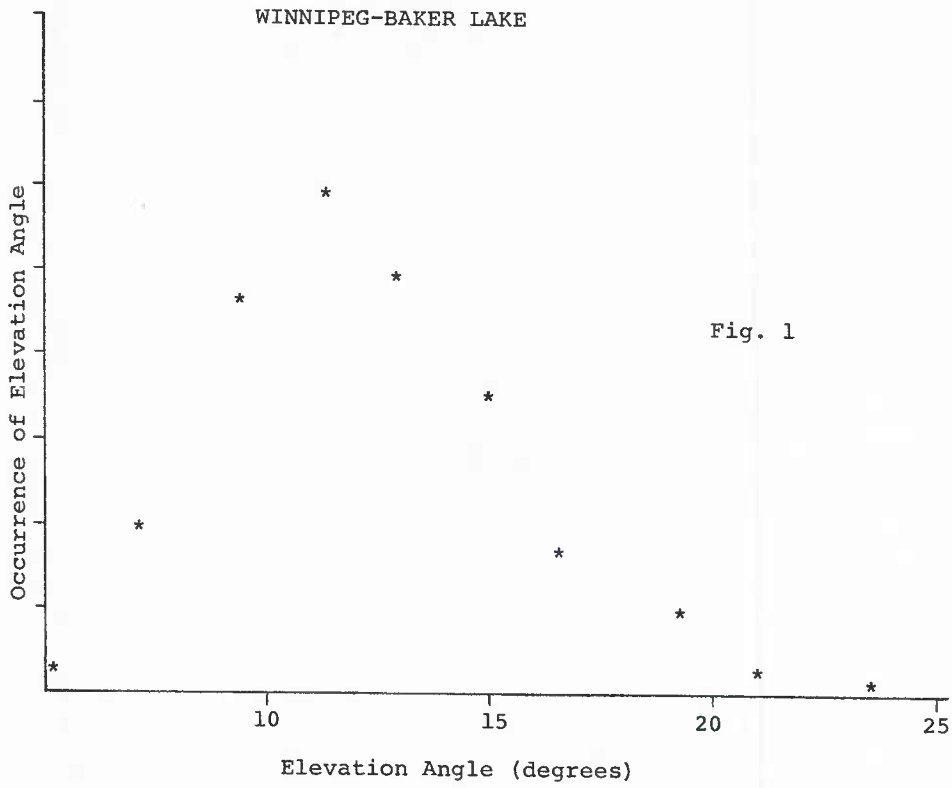
WINNIPEG-BAKER LAKE
Occurrence of Elevation Angle
10
15
20
25
Elevation Angle (degrees)

Fig. 1

# DEVELOPMENTS IN HF EQUIPMENT AND SYSTEMS: MOBILE AND PORTABLE TERMINALS

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01730

## INTRODUCTION

Before the advent of satellite platforms, sophisticated high frequency (HF) propagation and system research promised improved capability during disturbed ionospheric propagation conditions. However, satellite relays captured the imaginations and pocketbooks of the communications community in the mid-1960s. Consequently, extant HF systems aged while satellite systems were implemented. During peacetime, satellite systems transmit quality low data rate communications and navigation aids to mobile users, but there is now renewed interest in the low cost and survivability attributes of HF radio. At this time, when old HF prime systems need replacement for logistical reasons, the need for low cost communications that can survive jamming, nuclear effects, and space warfare is not satisfied. The HF renaissance is the response to this challenge.

Logistical replacement procurements that provide new capabilities are redressing the attrition of vacuum-tube radio equipment over the last decade. Procuring organizations typically compile specifications comprising state-of-the-art and new capabilities offered by competing vendors. Integrated circuits, which include microprocessors, synthesizer elements, and other evolving components, have led to new circuit architectures. The first of the following three sections describes: Receivers; Transceivers and Antenna Couplers; Antenna Kits; and Audio Channel Peripherals.

Emerging new systems automate network operations while adapting to propagation conditions. The second section describes:

ICS-3 Simultaneous frequency-agile transmission and reception on board ships

SELCAL, Scanning, and Channel Evaluation Automatic addressing and path selection

Packet Radio Autonomous messages find their way through nodes

SNOTEL Meteor trails, sporadic E, terrestrial diffractions, aircraft, etc. allow low probability propagation to service 500 stations

DIGISONDE An international digital ionosonde exhibits the level of complexity and many of the functions of an adaptive HF system

These systems are representative of what may be accomplished to satisfy specific needs. The equipment developed for each system is unique. More general capabilities may result from the use of new logistical replacement radios and computers. However, neither new replacement radios nor new systems furnish the potential capability offered by state-of-the-art components.

The third section, which describes the frontier of the state-of-the-art components, focuses on: Receiver Dynamic Range and Noise Figure; Frequency Synthesizers; Power Amplifiers; Portable Antenna Concepts; Processors; and Real-Time Clocks.

HF equipment is considered within the context of application within the network. A network control facility commonly comprises individual components of superior quality because space is not constrained and cost is minimized through decreasing operations personnel. Compromises among performance, size, and cost are usually necessary for the large numbers of mobile or transportable stations. The performance of the system is generally set by the ability of portable and mobile stations to communicate back to net control stations.

The system designer looks for cost-effective means to obtain performance goals. Improving equipment specifications may be less effective than a network alteration such as allowing net members to talk to alternative net control stations, thereby obtaining diversity gain. Or, since receiver/antenna systems are more agile than power emission systems, frequent channel occupancy measurements might afford interference avoidance for the relatively infrequent return link. Each of these techniques can reduce radiated power requirements by an order of magnitude. Many capable radio components are now commercially available for improved system design approaches.

## LOGISTICAL REPLACEMENTS

Evolving equipment designed to meet continuing needs has yielded significant improvement of receivers, transceivers, and input/output devices. Simplified operation through electronic control is a dominant inpetus. The use of highly trained operators in manual systems is held to be economically insupportable. Operators have other duties; equipment that relieves onerous tasks such as radio tuning and radio monitoring is welcome. The technical challenge has been to obtain electronic control without degrading the performance available from the best manual equipment.

### Receivers

HF receiver design has evolved whenever new component technology permitted and has accelerated with changes in modulation usage[1]. Tuned radio frequency (TRF) amplifiers enhanced the selectivity and sensitivity of early vacuum-tube detectors. Single tube regenerative amplifiers/detectors also increased selectivity and provided an audible heterodyne for continuous wave (CW) signals but were quickly superceded by the operating ease of the superheterodyne, in which the signal is translated to a more selective fixed-tuned low intermediate frequency (IF) amplifier. The HRO receiver from World War II combined TRF amplification and a single-frequency conversion superheterodyne.

Single sideband (SSB) voice modulation usage in the 1950s increased HF spectrum capacity, thereby implicitly requiring greater selectivity along with tuning stability. Multiple frequency conversion architectures with higher first IF amplifier stages moved images to filterable frequencies while lowering the translating frequency provided by the variable local oscillator (LO) and retaining selectivity in the lower IF stages. Our best predecessor, vacuum-tube receivers, use mechanically ganged, narrowband TRF amplifiers, mixers, and LOs in multiple-frequency conversion schemes.

The advent of semiconductors increased functional density. Ganged tuning was replaced by an untuned mixer, which translates the signal to an IF filter above the HF tuning range. Digital control of synthesizer LOs eliminated mechanical vernier gear tuning, simplified panel design, and extended control to remote and automatic terminals. The architecture illustrated in figure 1 has obtained universal acceptance. High performance receivers are small and light compared to receivers now being retired. However, replacement receivers using semiconductors initially experienced decreased performance[2].

Competitors concentrate on improving dynamic range and on digital control and displays. The dominant performance improvement has involved decreasing unwanted products from the front-end frequency converter. One objective is a high third-order intercept point. This is the input decibels referred to 1 milliwatt (dBm) level where the $(2f_1 - f_2)$ intermodulation products from two equal power test tones would equal the input level if extrapolated as illustrated in figure 2. Test signal frequencies are arbitrary but usually greater than 20 kHz from the tuned channel. Greater than 20-dBm third-order intercept is now obtainable in receivers with 14-dB noise figures.

The intercept concept simplifies the specification. However, more important third-order products may be formed by three interfering signals. Testing with an increasing level of white noise for an input causes the intermodulation products to rise dramatically after they reach the receiver noise level. This measurement technique [3] uses a filter to eliminate the in-band noise.

LO synthesizer phase noise[4] is now the dominant deterrent to discriminating against a strong adjacent signal. Figure 3 shows how a strong signal acts as a local oscillator and the phase noise from the LO acts as a signal; hence the term "reciprocal mixing." LO phase-lock-loop synthesizers are an economic compromise involving the dichotomy between tuning speed and phase noise. Synthesizers provide multiple channel tuning accuracy under digital control; moreover, they tune more quickly than can be accomplished with a frequency counter in control.

The industry now perceives the economic and performance advantages to digital implementation of selective filters, demodulators, and automatic gain control (AGC) [5,6,7]. Digital filtering can replace banks of narrowband IF filters such as the venerable magnetostrictively driven mechanical resonators. Digital filtering is programmably selectable and exhibits precise amplitude and phase characteristics. These parameters are as stable as the logic clock drive and, therefore, are suitable for phase-matched circuitry performance, such as for SSB rejection. Receivers experience wide signal amplitude variation during tuning, listening to fading signals, and communicating with near and far stations. Automatic gain control for various modulations and conditions is perceived to be a function better handled by digital programming than by analog circuits. Commercial digital signal processing (DSP) chips and boards inspire the new designs. Some of the next competitive offerings will include these digital enhancements. However, "hearability" performance improvements may not be so salable.

Few HF communications systems specify better hearability. However generalized receiver intermodulation criteria[8,9] have been quantified by data on spectral occupancy[10,11]. If cosite performance is not a problem, then a "goodness" criterion is the ability to hear the largest percentage of low level signals that could be heard by a perfect receiver. Comparison is an important test but a pragmatic performance criterion based upon propagation media statistics can show how many more channels could be heard if a specification were improved. Receiver quality is even an issue in making spectral density estimates: do our present receivers or spectrum analyzers really hear the noise floor? It is now timely to polish the tools that have been developed for judging receiver performance. However, the ground rules should satisfy requirements for receiving other modes of propagation, such as groundwave under quiet conditions and meteor burst, and for very high frequency (VHF) extension.

Receivers are more frequency agile than transmitters. Surveillance receivers at control centers can automatically monitor channel occupancy and potential interference to search out clear channels for network operations. Analog AGC voltage level measurements are too inaccurate and require calibration to be used to calculate required power emission levels. Unfortunately, new receivers do not yet measure signal levels accurately, but the inclusion of analog-to-digital (A/D) converters and DSP converters will lead to a useful measurement capability.

The operational convenience of the new receivers is outstanding. Not only are all functions controllable by a microprocessor at a remote location, but checkout is under software control (built-in-test-equipment or BITE). Many receivers use pushbutton and/or rate-of-scan tuning to aid manual frequency selection. The convenience of digital frequency selection enhanced by the ability to store frequencies provides an important reduction in operational labor.

Table 1 lists typical receiver performance parameters. Note that digital modem compatibility now requires a specification on phase linearity or differential time delay as a function of frequency within SSB channels.

**Transceivers and Antenna Couplers**

New transceivers are following the lead of digitally controlled receivers[5]. Transceivers continue to be designed primarily for the SSB modulation voice bandwidth channel, which specifies the interface to digital data peripherals. Fast automatic antenna tuning is available. Manifestations of these developments are discussed below.

Remote controlled transceivers solve several problems. For example, aircraft cockpit space is minimal, as indicated by the relative size of the remoted control box of the HF transceiver components shown in figure 4. Note the large size of the antenna coupler designed to mount in constrained spaces such as aircraft control surfaces. Manpacks need to be controlled by the operator without removing the pack. Tactical ground vehicles are vulnerable to attack so that remoting control is provided. The utility of fiber optics cables and connectors for remoting has been established on an operational multichannel radio set[12]. The convenience of computer terminal control through a standard telephone interface is generally acknowledged and is available in amateur radio transceivers.

The capability to remote the tuning controls can also provide frequency scanning or hopping. Several nomenclatured manpack radios have been modified for demonstrations. However, the implementation into a system implies systems network concepts, which are generally lacking except, apparently, for several European "slow-frequency-hopping" applications or for groundwave propagation systems where VHF conditions apply.

Some transceiver architectures allow separation of a small manpack transceiver from the power amplifier. Others combine the amplifier and the antenna tuner. The need for a VHF extension for interoperability has resulted in transceivers that extend into the VHF region; but, commonly, assemblies of HF and VHF equipment are mounted together. New equipment extending into the VHF region will facilitate the use of diverse propagation modes such as groundwave, unique diffraction opportunities, and meteor trails. These modes are important alternatives during disturbed conditions. The evolving transceiver antenna coupling and power amplifier configurations reflect existing space problems and the new solutions afforded by replacement equipment.

Transportable and mobile antennas have advanced little in the last decade beyond the automatic tuning of narrowband antennas. Antenna coupler tuning speeds of a fraction of a second are obtainable in units that switch vacuum relays to prelearned positions. Rotary mechanisms taking seconds to tune will be phased out because the reliability of these mechanical components (exacerbated by high vibration at poorly accessible aircraft antenna locations) causes radio maintenance problems.

Since human speech communications do not depend on phase or time delay characteristics, SSB radio and telephone service practices have emphasized sharp amplitude response filters which pass 300-Hz to 3000-Hz frequencies and strongly attenuate frequencies outside this region. Phase characteristics important for high rate data transmissions were not considered in older voice radio receivers. The transmitter and receiver channel filters cause especially large phase shifts near the band edges. This deviation from linear phase vs. frequency causes ringing and, consequently, intersymbol interference during serial data reception and leads to degraded bit timing on parallel tone data.

Figure 5 illustrates filter characteristics of two models of a military transceiver. The characteristics are not similar and show considerable ripple. A transmitter/receiver link will have two of these frequency dispersive apertures in tandem. A specific production run is likely to have a unique bandpass characteristic because the technology to obtain the narrowest bandpass has typically been unique; e.g., a magnetostrictively driven mechanical filter or crystal lattice filter at an intermediate frequency such as 455 kHz.

Data signal phase jitter can result from local oscillator instabilities caused by low-Q tuning circuits, phase-lock-loops, noisy circuitry, microphonics, etc. Older radios employing vacuum tubes for free running oscillators are not particularly noisy but also were not specified for this parameter. Microphonics caused by flight vibration can be serious. Transistor oscillator circuits have less microphonics but relatively more flicker noise and often use synthesizer designs which accentuate phase-lock loop-oscillator sidebands. Some commercial transceivers cause degraded data performance when carrying phase-coherent modulation. The newest militarized transceiver procurements do specify these data transmission parameters.

**Antenna Kits**

HF antenna vendors commonly provide kits for large wire antenna assemblages and log periodic horizontal beams. Other intermediate-size antennas are constructed by users guided by manuals or experience. Very large antenna construction is usually contracted.

Antenna kits relieve the buyer of design work and are crated for economical shipping to and from sites. The buyer prepares the site, assembles the antenna, and provides rigging. If, for example, the antenna is an omnidirectional vertical, the buyer may place ground radials and run coaxial cable. A rotary beam requires the rotor equipment and tower in addition to the beam kit. Kit design facilitates the time consuming erection process. However, communication within minutes after arriving on site often depends on simple dipoles or whips. Installing antennas unavoidably requires more experience than exists in most operational organizations; therefore, setup manuals and traveling teams are sometimes employed. Fixed-wire horizontally and vertically polarized log periodic antennas are not usually designed to be transportable.

Field sites may be superior to those of large fixed facilities, which comprise acres for both receiver and transmitter antennas often located a few tens of kilometers apart so that simultaneous opeations do not cause interference. Fixed sites are constrained to large tracts of rural land purchased with due concern for social, environmental, and political considerations. In contrast, a temporary site may have exceptional height above a ground reflection zone. Or a site on salt water may afford an extensive ground plane. And the antenna need not be a part of the station nor conform to standards. For example, Marconi's omnidirectional inverted-cone antenna for trans-Atlantic communications in 1902 is still used extensively because it provides less than 2:1 voltage standing wave ratio (VSWR) over 3–30 MHz. Airborne absolute gain tests (illustrated by figure 6) revealed several decibels lower gain than expected from one popular version, even though serviced with 120 new 250-ft radial ground wires. Buried radials are blamed for the low gain but mowing the field is a requirement. Furthermore, azimuthally dependent horizontally polarized response, which relates to the hexagonal support poles, was dominant at midband frequencies. The vertical pattern illustrated in figure 7 substantiates computations by Bevensee[13]; an antenna gain table for propagation prediction or emission control must represent sharp gradients. A short fat vertical over salt water would be superior.

A portable inverted cone antenna kit with a single support mast is available for high HF frequencies. Fat biconical antennas[14] are more appropriate mechanical structures and are available in portable kits from several sources. A more compact antenna exhibiting 4:1 frequency range, the wire cage in figure 8, includes an internal matching stub. These omnidirectional antennas over good ground are designed to provide low angle radiation to long distances without the tall towers required by horizontal antennas.

Horizontal antenna kits providing high angle radiation can be set up by four people in less than four hours. The design shown in figure 9 features a single mast and radiating elements that are also mast guys. A 4:1 frequency range is aided by resistive loading in the form of stainless-steel wire.

The above generic antenna kits are the response of industry to repeated orders. Successful experimental antennas may stimulate new kit offerings such as the vertical folded monopole and the inverted L mentioned later. Established product lines such as the log periodic beam compete with features such as lightweight wire structures, strength, and adjustable tilt for best elevation angle gain. In general, industry does not do much exploratory research but uses limited development funds towards their known market.

**Audio Channel Peripherals**

Two notable input/output devices now being developed or acquired are the small data terminal and the secure voice terminal. These peripherals interface with radio transceivers at the audio channel and therefore include a voice channel modem that puts performance requirements on the transceiver.

The audio channel modem interface of SSB transceivers, receivers, and exciters is a de facto standard, which came to us via telephone practice. Like most standard interfaces, it has its costs, but careful attention to phase dispersion does ensure transparency through the radio equipment. Data-only radios can be simplified by using a digital interface.

It would be satisfying to report on a consensus on modem design[15]. Several manufacturers have produced unique HF diversity modems for low data rates (< 600 bps); interoperability is out of the question. Although some of the same manufacturers also produce HF simulators for modem testing based on common model concepts, results of competitive testing have not been widely reported and consequently are not available for system design. On-the-air tests are unique and attain credibility proportional to the number of days of testing.

Amateurs struggle with inadequate frequency shift keying (FSK) telephone channel modems. HF users of FSK modems obtain reliable transmission through automatic request for retransmission (ARQ) techniques that are now internationally recognized under Consultative Committee, International Radio (CCIR) Recommendation 476-3 and called AMTOR. This ARQ system avoids the need for good modem performance for low data rates.

There are indications of strong competition[16,17] for 2400-bps modems offering enhanced performance. Serial modems are vying against parallel modems even within the same companies. The perceived advantages of serial modems included the 9-dB advantage of a constant envelope signal; the signal need not be attenuated to avoid amplifier clipping on peaks. Another power advantage may accrue to the use of efficient RF amplification: SSB requires a linear amplifier that commonly is only 15 percent efficient while constant envelope amplifiers can attain 60 percent (6-dB improvement). If serial modems perform best, a new interface and transceiver architecture should appear. The next few years should witness national and international tests on simulators and on-the-air that will establish the best modems.

### Handheld Data Terminals

In the field, handheld terminals substitute for Teletype equipment and computer terminals. They allow off-line composition and editing before transmission at Teletype or burst rates. Internal simple FSK modems are typical. Demodulation, storage and display complete their capabilities. Some representative terminals are illustrated in figure 10.

Field terminals are unique. The keyboard may be sealed against dust and moisture. An abbreviated keyboard and display aids standing or moving operators. One unit can generate a map on a plasma display grid that is also overlaid with a membrane keyboard. Another converts alphanumerics to Morse code and conversely demodulates Morse code to provide an alphanumeric display. A commercial unit provides serial and parallel interfaces to commercial test equipment and is programmable in BASIC or FORTH. Others are marketed for system testing and inventory or data logging.

With such a large number of inexpensive personal computers with typewriter keyboards available, abbreviated keyboard units are likely to be used only in constrained or unique situations. Few computer options incorporate the efficient Dvorak keyboad. Small keyboards that claim data entry efficiency receive skeptical reviews. Voice-operated data entry also has had lukewarm reception except when absolutely needed. Unique military applications should not expect to find an appropriate commercial or consumer handheld device. One popular unit, which requires a second hand to help select the alphabet with one of three keys, has been applied as an order pad at restaurants and as a meteor burst terminal. Another has been designed for single-hand use.

Generic battery-operated personal computers weighing from 0.5 to 5 kg have a QWERTY keyboard and include an operating system, an editing capability, and telecommunications functions. Most use low power complementary metal-oxido-silicon (CMOS) parts or at least memory. Liquid crystal displays of greater than 40 characters on eight lines greatly decrease power consumption and cost. Plasma and fluorescent panels are available on more expensive models. A few of these units are rugged or can be militarized. Alternatively, office models may be an expedient choice. Generic computers usually provide for loading specific application programs from either cassette, disc, or read-only memory (ROM). They are convenient and inexpensive; this paper was drafted and edited on a word processor/formatter costing under $300.

### Secure Voice

The United States has worked towards secure voice communications systems for 20 years. Earlier programs included 2400-bps vocoder competitions, and the development of wireline and HF modems designed to interface with encryption devices. The quality of the transmitted voice and the cost of adding three major additional components (digital voice device, modem, and security device) to radio terminals did not justify acquiring a general purpose capability until now. The new, linear predictive, analog-to-digital convertor, combined with modem and ciphering capability in one package, is a significant improvement[18].

HF modems for high speed data such as 2400 bps digitized voice have evolved toward multiple-tone signaling carrying differential phase shift keying. Serial data modems provide economical service over wirelines; but, on HF radio networks, multipath delay usually causes intersymbol interference. Consequently, parallel tones having durations much longer than the expected multipath delays are favored, even though parallel channels have been an expensive implementation. Digital data processing now minimizes the parallel channel cost penalty; however, the linearity requirements on RF power amplifiers, combined with the high peak-to-average power ratio of parallel tones, degrade system performance compared to that with constant envelope modulation modems.

The conflict between maximizing power output and minimizing self-induced intermodulation noise is addressed in a new HF modem by hard clipping the ensemble of tones to reduce the peak amplitude and then providing band limiting. About 9.5-dB peak clipping provides the best overall performance. Nevertheless, new exploratory serial modem developments may provide performance improvements over the parallel modem.

The advanced narrowband digital voice terminal (ANDVT) data modem need not attain a 1/1000 bit error rate for good digital transmission of data. Furthermore, some data are more important than other data. The ANDVT modem has been optimized for sending voice data over HF propagation paths that protect critical voice data with error-correction logic.

The synchronization of modems and security devices requires several steps that must be especially robust or reception will not begin. Preambles comprise synchronization signaling sequences which may require 0.5 second for transmission. The synchronization time may be impacted by radio automatic level control (ALC) and AGC time constants and by automatic antenna-tuning alglorithms in some transceivers designed for push-to-talk voice communications.

Angalog-to-digital conversion of voice for HF now strives for natural voice transmission within a 2400-bps data rate. Both modem and A/D voice conversion will continue to receive attention in the years to come because secure voice still significantly degrades the ability to communicate, especially under marginal conditions when the modem exhibits threshold performance. The possibility now exists, however, that new serial modems will permit superior performance in the threshold performance area.

## NEW SYSTEMS

The systems discussed below involve networks that are expected to influence future configurations. The great majority of HF systems operate in point-to-point service or as part of a limited network of terminals serviced by a network control station. System agility has been limited by operators, equipment, and antenna systems. As new automatic radios enter service, they will enhance operation of the existing networks and will be capable of participating in more complex networks through greater frequency flexibility and through the ability to choose alternative channels, propagation paths, and modes that require minimum power levels and offer improvements in privacy.

## ICS-3

Navy HF systems aim towards frequency agility on shipboard platforms where colocation interference is a dominant concern. The United Kingdom ICS-3 program[19] of the Admiralty Surface Weapons Establishment is echoed by an attempt by the U.S. Navy to radically lower all self-generated components of the noise floor[20]. Although this symbiosis is indicative of a probable new HF networking capability, ICS-3 primarily enhances groundwave and skywave transmissions through rapid propagation selection and data handling.

Figure 11 contrasts ICS-3 architecture and its predecessor. Frequency agility is obtained by eliminating the complex of tuned narrowband transmitters and multiplexers. Instead, a broadband power amplifier and antenna handle multiple independent signals. Receivers are fed from an active antenna, which is small and located for improved isolation from the transmitting antenna. Control is enhanced by computers.

The penalty for broadbanding power equipment is broadband generation of intermodulation products and phase noise, which obscure signals to be received. The U.S. Naval Research Laboratory embarked on a system-wide attack on improving the ability to hear weak signals in a cosite environment. Their analysis assessed state-of-the-art equipment in relationship to desired system performance. Contracts and research addressed the desired equipment capability. Despite the heroic equipment improvement, a shortfall is reported along with new investigations on shipboard noise canceling[21] and on postselector/preselector filters[22]. Since these approaches have just been reported, one can speculate that other expediencies will appear. One approach is to maintain narrowband capabilities. Another is to use the shore-station solution of separating transmission and reception facilities; i.e., having a receiving station on another ship within reach by line-of-sight (LOS) relay. But most likely, frequency management and emission control will be applied for a compromise capability that involves scheduled operations.

The ICS-3 related programs spawn new hardware of revolutionary capability. That equipment, in turn, should lead to simultaneous networks of increased agility and encourage the design of systems capable of controlling link emissions so as to minimize self-interference and notice by other spectrum users.

**SELCAL, Scanning, and Channel Evaluation**

SSB voice networks that service aircraft seek to auomate HF communications. Elimination of channel monitoring (an onerous additional aircrew duty) and automatic addressing at the mobile terminal via a digital preamble (SELCAL) are collateral improvements now being tested with a scanning system that picks the best of alternative paths and frequencies as determined by channel evaluation. Channel scanning for occupancy originated in VHF equipment and is now available from several HF manufacturers. Channel evaluation algorithms are often proprietary and are under continued development.

Link quality can be assessed via the use of one or more measures[23] such as: signal level, signal-to-noise ratio, digital error detection, sounder signal parameters (time delay continuity), and memory of recent assessments. The requirement is to recognize a desired signal and to rate it among alternatives that may be in different bands or from alternative paths or modes. Quick measurements require using several frequencies simultaneously to obtain frequency diversity. Signal fading and noise variations preclude accurate single-channel measurements in less than several minutes. Often a memory table is the final arbiter of channel selection. One scheme gives greater weight to recent measurements.

SELCAL HF networks using scanning are being automated. Implementation of SELCAL and scanning techniques will take place through the remote control interfaces of new replacement radios. Demonstrations involving several terminals have shown hardware feasibility. Analysis of large networks is lacking. Clocks are not proposed, so that all possible combinations and frequencies may need scanning. Figure 12 shows a proposed scanning scheme that requires 0.5 second per frequency.

A prognosis for the use of frequency scanning for automatic alternative routing via the best frequency and path will be utilized in limited networks, and time division multiple access or packet systems will appear in large networks.

**Packet Radio**

In contrast to either circuit switches or SELCAL and other time division multiplexing networks, autonomous message packets seek a route through nodes[24]. Each node demodulates and stores packets which are traveling further. Contention eventually limits throughput at a node, but the diffuse network allows alternative route choices.

Packet messages carry an address protocol defined at increasing levels associated with network complexity[25]. The dominant concern is avoiding an overload disaster. Message loops are eliminated by protocols that keep track of transitions through node "tiers." Overload is minimized by using "transfer point routing" and by slowing the pace during stress. Unknown addresses may be found by "flooding" the network for routing aid. Experiments have implemented relatively few nodes and consequently have not exercised multitiered protocols. In fact, test network sizes are more appropriate models for using long-range HF as an access to a local LOS repeater node.

Packet radio is promoted for computer communications. SRI research/military experiments[26] using "L-band" (1 to 2 GHz) LOS radios of diminishing size (see table 2) exchange 1000-bit data packets in microseconds. Other bands were considered but lacked sufficient assignable bandwidth. Table 2 illustrates the steady progress in equipment design.

More recently, packet concepts have attracted radio amateurs who by their organic nature use less data and structure. However, "rag chewing" is not desired by these enthusiasts. The attraction of packet radio appears to be bulletin board operation, which avoids operator presence. An operator signs in to see what messages await or affect him. Radio amateurs use 1200-bps packets on LOS VHF repeaters and on satellites. The 20-meter band provides 300-bps gateway access to distant or foreign amateurs. In Boston, one gateway station aims West and another aims toward Europe. Figure 13 shows the gateway block diagram. One bulletin board is maintained by a Z80 board, which has two RS 232c ports and 64 kilobytes of random access memory that holds local addresses. About 50 amateurs in the U.S. alone have bulletin boards with automatic forwarding and 15 of these have HF gateways.

The protocol defined by CCITT X.25 is implemented in modified form by the terminal node controller (TNC). In the U.S., Amateur Radio Relay League (ARRL) radio amateurs have held four annual conferences[27], which have publicized and formalized this enabling agreement, encouraging participation. The development of the TNC that interfaces with transceiver audio has been accomplished by several amateur clubs, one of which subsequently provided kits and extensive documentation at cost (about $250). Several thousand boards have been shipped. Radio amateurs comprise the largest packet radio community. Now consumer manufacturers are participating. If the popularity momentum continues, packet radio may have as important an effect on the amateur and military communications communities as SSB did in the 1950s.

**SNOTEL Meteor Burst System**

SNOTEL is a meteor-burst communications system run by the U.S. Department of Agriculture Soil Conservation Service, which relays snow cover data from over 500 remote data collection sites in the mountains of the western United States to a central processing facility of the Western Technical Support Services Center in Oregon. It is notable for the size of the network (see figure 14), the transient and diverse propagation modes, the small terminal equipment, and the speed of design and installation. Furthermore, it provides another argument against the arbitrary limiting of HF equipment to 30 MHz. Ionospheric propagation below 30 MHz is not sufficiently adaptive to diurnal variations, 11-year sun-spot cycles or a disturbed ionosphere. Extending new HF equipment capability into the low VHF band is expected to improve the use of the HF/VHF border frequencies by utilizing ionospheric propagation above the predicted maximum usable frequency (MUF), sporadic E, aurora, meteors, and field aligned ionization. Diffraction, ground wave, and even reflection off aircraft are important LOS propagation modes. Meteor trail availability is a seasonal and diurnal variable which complements ionospheric propagation. The significance of meteor-burst systems lies in their temporal adaptivity to various low duty cycle propagation path opportunities for the transmission of short messages.

Although non-meteoric propagation modes support a large fraction of meteor burst communications throughput, the dominant design consideration has been transmission of a burst of data during an ionized trail life of a fraction of a second. These underdense trails come from the many random meteors which are much too small to leave a visual trail. The SNOTEL remote terminal generates 300 watts between 30 and 50 MHz to send about 150 bits (78 bits of data) and receive an acknowledgement in 0.1 second. The low duty cycle allows solar cell power for charging the remote terminal batteries. Figure 15 illustrates a portable test terminal which can be used for communications with the base station. Data interfaces directly with modulation and detection circuits instead of utilizing a SSB voice channel; hence, eliminating SSB circuitry.

There are now vendors and equipment specifically for single-frequency meteor burst networks. As yet frequency agility, adaptive data rates, or other enhancements are not incorporated. Differential binary phase modulation is common. Integral microprocessors control data and error detection. Antennas are typically narrowband Yagi arrays. Equipment seems to be used to further network application research, which is pointing out the importance of antenna configuration to increasing throughput and to minimizing man-made noise.

The network is interrogated by two master stations located centrally in Utah and Idaho. Both master stations have four receivers connected to Yagi antennas covering quadrant sectors. Interrogation protocol is based upon the limited footprint afforded by the trail location and orientation with respect to the terminals. Earlier data indicated that the footprint was roughly oval with 200- by 1000-km axes[28]. In SNOTEL a 50-km footprint was expected[29]. Remote stations that are not in the same area are grouped for simultaneous interrogation so that a chance meteor causes only one terminal to respond. First, however, a preliminary call list takes care of those LOS or diffraction paths that would certainly cause collision interference. Then, the address code enlarges the call field. If collisions disable reception, the call field is narrowed again. Finally, a list of remaining stations is called. Although the system experiences difficulties with auroral conditions, it successfully interrogates 90 percent of the field in minutes even though network and modulation are not sophisticated.

The impact of SNOTEL is the demonstrable economy of a simple data terminal that uses many propagation modes in the relatively unused HF/VHF border frequencies. SNOTEL desmontrates that a few network control stations have a strong probability of communicating with a mobile terminal within a few minutes.

### DIGISONDE Networks

Oblique ionospheric sounders (ionosondes) have aided point-to-point communicators for three decades. A commercial low power chirp sounder which includes a spectrum analyzer provides test system documentation and aids frequency selection on instrumented paths[30]. The chirp sounder has also been applied to network frequency management, but requires operator interpretation and manual control of the communications system. The communications overhead of a large network under manual frequency and path control may be unmanageable.

SRI implemented an automatic digital sounder system called CURTS[31] to advise data communications operations from Hawaii to California on when to change frequency. The patented Canadian CHEC sounder system[32] added a base interference assessment to the sounder signal, allowing the mobile terminal to choose the best frequency. These efforts were aborted in the early 1970s with the expectation of satellite systems. Furthermore, concern for sounder signal pollution and the hope for the economy suggested that sounders should be imbedded into the communications equipment.

In the late 1970s both the National Oceanic and Atmospheric Administration (NOAA) and the University of Lowell built digital ionosondes for the international scientific community, which desired a coherent amplitude and phase measurement capability and the ability to sense wave polarity[33] for the capability to track moving ionospheric clouds. An international network of ionosondes is to be implemented to support communications.

Scientific usage should not obscure the general capability of these later systems as a model for adaptive HF systems. Synchronization of a network of oblique sounders parallels proposed adaptive communications systems. However, ionosondes are less concerned about monitoring functions. Their features include:

- Frequency agility
- Broadband RF power source
- Broadband antennas suitable for polarization measurements
- Fast antenna array switching
- A/D converters for I and Q channel amplitude and phase
- Numerical processing
- System clocks
- Software control
- Remoting

Systems such as the DIGISONDE should encourage the communications community towards a generic capability.

## NEW COMPONENT TECHNOLOGY

New system concepts have promoted component improvements that really represent new capabilities. The ICS-3 analysis recommended that the linearity of receiver mixers and of power amplifiers be materially improved and that the phase noise of fast hopping synthesizers be reduced so that the noise floor caused by collocated equipment would approach atmospheric noise. Some of the reports from development programs are now appearing.

Few portable HF antenna kits are new, but diverse innovative concepts provide more gain. Some ways to obtain a wideband capability are also described.

Computer processors provide the key to enhancing adaptive communications. The processors need mating to analog radio equipment and to real-time clocks. Trends are discussed below.

### Receiver Dynamic Range and Noise Figure

Although the new frequency-agile receivers exhibit good dynamic range, noise figure suffers. Required receiver performance is very much related to HF system requirements. For example, the third-order intercept point increases in importance as the level and density of unwanted signals increase aboard warships where sensitive receivers are operated in close proximity to high-power transmitters. An additional problem specific to receivers collocated with transmitters is that of receiver front-end protection. Shipboard installations often include pads in the receiver antenna leads to minimize the problem. Inefficient receive antennas (such as short active whips) are located for decreased coupling with transmitting antennas. These approaches assume the presence of an excess of signal-to-noise ratio; this assumption is inconsistent with scenarios involving ground wave or meteor burst propagation and excessive ionospheric absorption.

The newest diode mixers allow systems to obtain +45-dBm input intercept points[34] but suffer 12-dB noise figures. Diode mixer loss of 6 dB contributes directly to the noise figure along with various circuit insertion losses from the first IF filter down through the receiver chain.

Two circuit approaches to achieving low noise figures and large dynamic range are to use a low-noise, low-gain amplifier in front of the mixer and to use mixers which exhibit gain. Varactor upconverter mixers have been employed in VHF systems which exhibit 6-dB noise figures and 20-dBm intercept points[35]. Amplifiers are characterized in figure 16 by noise figure, and input third-order intercept point. Apparently, 2-dB higher noise figure is traded for about 10-dB greater signal handling capability. In general, the use of an amplifier provides a range of compromise.

Input filtering can help decrease intermodulation products from interference sufficiently far from the signal frequency. High-power, agile, programmable filters have been under research for ultrahigh frequency (UHF) transmitters, but not for HF receivers[36]. So far nothing has appeared to replace switched 1/2-octave filters.

One system approach to ameliorating nonlinearities is to use a programmable attenuation ahead of the receiver so as to operate always at the point at which receiver noise has just begun to degrade performance; thereby, a lesser intercept point specification is possible. Another approach is to provide both a low noise figure circuit and a high intercept point circuit and to select the appropriate mode depending on propagation conditions. Electronic switches are somewhat more linear than the best amplifiers and mixers and do not introduce nonlinearities of the magnitude caused by AGC control of amplifiers. Finally, low noise is more important at higher frequencies where intermodulation may be less.

**Frequency Synthesizers**

The programmable frequency synthesizer performs the local oscillator function in frequency agile and frequency hopping receiver and exciter subsystems. Tuning speed (settling time) and spectral purity (particularly off-tune residual phase noise) are the two most important parameters. The former sets the limit on channel scanning or hopping rates and the latter limits the selectivity of receivers. This latter problem occurs both from cosite transmitter signal spreading and because of reciprocal mixing whereby synthesizer noise mixes with high-level, off-frequency interference to produce the receiver IF frequently (see figure 3).

Two basic kinds of synthesizers have evolved. One is the indirect synthesizer that utilizes phase-lock loops and hence tends to have a relatively long setting time (good performance is in the order of 1 ms). The other is the direct synthesizer that uses mixing/filtering processes, which settle quickly (in the order of tens of microseconds). The relatively poor noise performance of early direct synthesizers has now been largely overcome and comparable noise-floor performance is available from commercially available units of both types (phase noise floor in a 1-Hz bandwidth in the order of 135 to 140 dB below the desired output). However, the direct synthesizer is significantly more costly than the indirect type.

Note that despite what has been achieved, and despite the utility of synthesizers in modern systems, noise floor performance still needs to be improved[20] (by at least 20 dB) to reduce the effects of receiver reciprocal mixing in a high noise environment and to reduce the noise bandwidth of collocated transmitters, given contemporary block diagram architecture.

Frequency synthesizers for communications purposes are usually built into the communications receiver or exciter but, because of their usefulness as laboratory equipment, several superior commercial devices exist as stand-alone units. The performance noted in the foregoing paragraph generally applies to these stand-alone units. Noise performance of built-in synthesizers is poorer than for laboratory type equipment (see figure 17). Synthesized communications receivers and transceivers currently available use phase-lock-loop systems that exhibit several millisecond tuning times rather than the several microseconds tuning times available through the use of larger and more expensive direct frequency synthesizers. Furthermore, their performance is frequently poorly documented.

New designs underway make use of novel technology. For instance, integrated circuits (ICs) that count down either by n or n+1, clock faster, and permit finer countdown steps. Faster ICs are being developed from GaAs semiconductor materials. Surface acoustic wave (SAW) devices allow VHF/UHF hi-Q oscillators and filtering in small packages.

Synthesizers satisfying frequency agility requirements as well as having acceptably small phase noise are still expensive and oversize despite original equipment manufacturer (OEM) and laboratory equipment competition. To come ahead, new circuit architecture specifically for receiver or transmitter design may be as important as new technology such as surface wave devices.

**Power Amplifiers**

Most power amplifiers in military use employ vacuum tubes. Output stages are tuned to minimize harmonic, intermodulation product (IMP), and other spurious and wideband noise outputs. Tuning is typically accomplished through use of servo systems which are elctromechanical in nature, and are therefore relatively slow (on the order of 10 to 30 seconds to tune across the HF band) and unreliable.

Alternatively, wideband amplifiers have been produced, with early vacuum tube models using distributed transmission line technology. These suffered from relatively high harmonic, IMP, and spurious outputs, as well as from wideband noise. However, since they eliminate the function of output tuning, they permit system tuning times governed only by settling times within the signal sources and by the characteristics of associated antenna and coupling networks.

A compromise approach at the lower power levels is to use wideband technology and switched half-octave power amplifier (PA) output filters. All harmonics lie outside the selected half-octave pass band. This approach permits implementation of a PA amplifier having improved harmonic and spurious characteristics, essentially instant tuning within any of the half-octave bands, and band-to-band tuning as fast as the switching time of the vacuum relays (about 25 ms). Reducing this switching time through the use of diodes has been demonstrated at UHF but not at HF[36]. Additionally, there is concern for spurious products resulting from the nonlinear diode characteristics and the power consumption associated with high diode current biasing.

With the development of RF power transistors, a new linear PA architecture has emerged whereby pairs of number of relatively low-power modules are combined using 90° hybrid couplers to produce a high-power unit, typically several hundred watts. Hybrid couplers facilitate matching semiconductor frequency dependent input impedances while maintaining broadband module interface matching. The system is wideband throughout, permitting essentially instantaneous frequency selection, and may be fitted with output half-octave filtering to reduce spurious emissions, as noted earlier. A major advantage of this architecture is that modules can fail without a resulting catastrophic system failure. If there are four output modules, failure of one module results in a loss of one-half the output power. The important concept of graceful degradation of performance can be supplemented by a capability for on-line module replacement, thereby minimizing PA down time.

Modern PAs do not suffer damage from output mismatch. Circuitry senses the mismatch and turns off the input signal before damage occurs, even for open-circuit or closed-circuit conditions. A wideband amplifier approach uses 90° hybrid circuits to dissipate reflected power in a dummy load.

PA units for use at HF are currently designed for linear amplification of SSB signals. Such operation permits easy implementation of output emission level control through manipulation of low-level input signals, minimum output filtering requirements, and minimizes interactive generation of spurious signals by multiple collocated PA units (for example, in ships). The serious problem of collocated equipment in the ICS-3 system has prompted recent activity to produce better linearity (and lower noise) through the use of feed-forward techniques[37]. Figure 18 shows a block diagram of the techniques. Delay lines allow comparing the desired signal with the amplified signal and inserting a lower power correction signal. At the present time, 100-dB suppression of two-tone, third-order products has been obtained for 100-W peak effective power at 5 MHz[38]. Performance is poorer by 20 dB at 30 MHz. Unfortunately efficiency is low (less than 10%); hence mobile equipment performance would suffer.

The use of linear class-A PA architecture prohibits achievement of the higher efficiencies offered by class B, C, or D nonlinear circuits. Efficient nonlinear amplifiers that can be programmed to operate in a linear mode have been investigated but are not available. Efficiency appears to require tuned circuits and constant envelope modulations. However, 6-dB more RF power from portable-power-limited transmitters is available through efficient emitters. Of course, nonlinear circuits cannot in general reproduce signals that do not have a constant envelope.

Linear 1-kW PA units, available commercially, are capable of frequency-agile operations over the entire HF band and into the low VHF band. These units provide acceptable harmonic supression and low-level intermodulation products, mismatch protection, and utilize the principles of graceful degradation of performance and on-line PA module replacement. They do not yet make use of the new feedforward techniques.

### Portable Antenna Concepts

Diverse concepts drive new portable antenna systems:
- Broadband antennas decrease requirements for fast hopping antenna couplers.
- Directive antennas discriminate against jammers and intercept sites.
- Quick-erecting masts and towers increase gain at low radiation angles.

#### Broadband Portable Antennas

Fast frequency-hopping antenna couplers soon expend the service life of their vacuum relays. Solid-state relays are still unsatisfactory at 100-W power levels. One solution is to hop only within the bandwidth of the antenna coupler system. Another is to ameliorate or eliminate coupler problems altogether with a broadband antenna. Mobile HF operation presently proscribes whips with an antenna coupler. However, investigators and venders are searching for quickly erectable broadband HF/VHF antennas.

German and U.S. amateurs resistively load folded dipole and monopole antennas at the point opposite the feed[39]. A commercial HF vertical version is available. A U.S. patent describes replacing the resistive loading with another fold as shown in figure 19[40]. Operational tests at low VHF indicated results similar to a standard whip.

Chinese and Japanese texts[41,42] describe optimum shapes for wire equivalents of fans and bodies of revolution, furthering work by Brown and Woodward[43] on cones and fans and Mason[14] on biconical antennas. Transformer matching maintains VSWR over larger bands. The U.S. Navy uses fans incorporated with superstructure. Figure 20 shows several configurations including a biconical sleeve monopole version that raises the feedpoint. Fan antennas have greater impedance excursions than bodies of revolution and exhibit frequency-dependent directivity. Short ground screens are frequently part of kits.

Another broadbanding technique places impedance along the antenna. A multiple antiresonant trap antenna (shown in figure 21) was designed for shipboard[44]. Single resonant traps are used by amateurs to lengthen the antenna below resonance and to shorten the antenna above resonance; this is also effective for fat cage antennas[45,46]. Resistive loading is sometimes employed but is inherent in coil losses. Continuous inductance has been tried but not found competitive with the folded whip mentioned above. However, a patented helix of inverted umbrella proportions provides omnidirectional horizontal polarization.

The combination of an antenna coupler with antennas that exhibit low impedance excursions is likely to cover an optimum HF working bandwidth, as illustrated by the ionospheric path signal level data in figure 22. Resistive loading is gaining acceptance for broadbanding antennas constrained in size.

#### Directive Portable Antennas

Yugoslavian and German investigators[46,47] have used the method of moments to optimize the shape of a 3/4 wavelength monopole for maximum azimuthal gain. These antennas act like ordinary verticals at 1/3 the optimized frequency. Forward-leaning fan antennas are applied to broadband TV reception. Figure 23 shows results for optimum shapes.

Directing a sector ground screen provides gain for vertical antennas. Ground has considerable influence unless the screen is raised several feet. When on the ground a wavelength is shortened by slow wave propagation. Studies indicate that a modest screen may outperform larger screens for low angle radiation. Figure 24 shows computed directivity[48].

#### Fast-Erecting Masts and Towers

Horizontal dipoles and log periodic beams may sometimes be raised to a height which obtains ground reflection at a desired elevation angle (see figure 25). Wire versions of beams and fat dipoles are appropriate for portable masts and towers. A number of venders supply various quick-erection systems.

One articulated tripod stands in 100-mph winds on roofs or hills. It is manually jacked up and down 7 meters in minutes. A vehicular system uses a pneumatic mast with hydraulic outriggers for carrying several-hundred-pound antenna systems in 50-mph winds. Vehicular-mounted towers are more common. A generic hydraulic man-platform lift can be modified for carrying antennas in mild weather. Often the antenna, tower, and vehicle are an integrated unit. Figure 26 illustrates several concepts.

#### Antenna Summary

Recent antenna investigations and designs are international and diverse and appear to be growing. Efficiency concepts are giving ground to broadband techniques since antenna couplers incur RF and direct current (dc) losses too and are an expensive logistical problem. Good field sites dramatically improve gain at low elevation angles, but operational measurement support is lacking. Not enough tools are available to support rapid deployment in the field. Too much operator judgment and capability are still required.

Emission control for privacy and for moderating cosite and other user interference requires antenna calibration. Airborne testing is time consuming. Ground conductivity measurements aid calculating patterns. Quick testing of vertical pattern gain would be a desirable operational capability.

### Processors

Digital processors are an increasing part of modern HF systems. Initially processors replaced integrated circuit logic boards used for message handling and error detection and correction. Then processors were applied to the remote and automatic control of the frequency and mode of receiver and transceiver terminals. The manual process of tuning antenna couplers was automated. Processors enhance control of modern systems by orchestrating message protocol, timing, frequency hopping, channel evaluation, propagation prediction, and adaptivity to noise, interference, and path loss. Now digital processing is applied to narrow band filtering and demodulation functions and to controlling receiver gain (AGC).

Minicomputers were the initial processors. Then instrument controllers developed by the measurements industry simplified interfaces and spread the use of the interpretive language BASIC for easy programming. The proliferation of personal computers radically decreased costs while providing internal space for printed circuit boards that customize display and input/output capability. A large number of inexpensive portable personal computers now compete for the keyboard and display terminal market. Equipment manufacturers agreed on the CCITT V.28 (RS-232C) and the IEC 625 (IEEE 488) serial and parallel interface standards, and the "Centronic" printer interface, and cooperate on internal bus structures for common microprocessor systems. The success of IBM in opening the operating system and internal structure of their personal computer to second-party hardware suppliers appears to have encouraged new bus structures that are supported by a plurality of international board suppliers.

These last several years have generated an amazing market. Designers can choose among a large number of microprocessors and family chips for use within equipment. They can control equipment with a low cost generic personal computer. Alternatively, they can build up a processor capability from a cage of printed circuit board modules, such as the Eurocards hosted by the VME bus or boards designed to the Multibus or STD bus. If portability or battery power is required, CMOS parts and boards are sold for popular microprocessor families.

Although BASIC is the popular interpretive language, tiny-BASIC, and FORTH are used for small control applications because they speed processing. Interpretive languages can also call out assembly language subroutines when occasional speed is required. Bigger software programs such as in the DIGISONDE are developed and altered in a higher language such as FORTRAN, PASCAL, or "C." Then the current program version is compiled into machine language for running in real time. Disc memory storage facilitates quick data transfer from development system to real time system.

Fast "number crunching" operations employ byte-parallel multiplier circuits and "bit slice" processors and speed the process by "pipeline" overlapping of program fetch and execution cycles. These digital processing systems are commercially available on boards.

Although digital circuit designers have many device options, they lack sophisticated solutions to unique electromagnetic compatibility (EMC) problems. Military processor packaging illustrates the increased care required for security. The commercial processor community is aware of its responsibility to comply with national emission standards and is looking to new technology such as conducting plastics for electromagnetically shielded enclosures, printed circuit board bypassing of conducted noise, and ribbon cable shielding. However, system engineers must still guide EMC design.

Processor industry competition is likely to sort out and consolidate markets during the next few years. Vapor phase soldering of surface-mounted components on multilayer printed circuit (PC) boards promises to quarter the required area, effecting economies associated with fewer boards. Higher density supports higher clock rates by minimizing lead inductance and cabling.

Digital processing is likely to be an increasingly satisfying arena for developing new HF/VHF systems. Evolving software development tools are important in this new technology.

**Real-Time Clocks**

Knowing the expected time of arrival of communications and navigation signals facilitates synchronization of demodulator circuits and thereby increases message throughput. Electronic clocks are increasingly seen in new communications systems as well as in navigation aid equipment. Calendar clocks and timers are often available in, or as a PC card accessory for, instrument controllers, but their accuracies and resolutions are unacceptable for most communications or navigation uses. Synthesizers, receivers, and transceivers make provision for connecting external precision clocks for frequency tuning accuracy. These several needs (i.e., frequency control, operations clocks, and signal synchronization,) can be aided by using a standard clock in each terminal.

Electronic clocks provide three classes of capability. TV stations employ high-accuracy, primary standards (cesium) to avoid disturbing the TV display when switching between unsynchronized program sources and when nonlinearities in frequency multiplexed circuits might cause moving distractions. Mobile terminals, constrained by size, cost, power, and warm-up time, frequently use temperature-controlled crystal oscillators. An intermediate capability is available in small, rubidium gas standards.

The cesium primary standard is considered to be drift free and is suitable for large control stations where it provides the reference for the network. The standard is generally rack mounted and kept on continuously in a benign environment. A rubidium gas oscillator provides an order of magnitude less accuracy in a 12-cm cube package at about 1/3 the cost of a cesium standard. An oven-controlled crystal oscillator in a 7-cm cube provides 1/100 the accuracy of a rubidium oscillator at 1/10 the cost. The latter two oscillators are secondary standards; i.e., they require periodic adjustment as indicated by the rising curves on the right in figure 27.

HF propagation typically exhibits propagation delay dispersion of less than 2 ms, which corresponds to a path length variation of 600 km. An illustration of typical clock usage is the oblique sounder where the stepped frequency transmitter emissions and receiver tuning should remain in synchronization without operator intervention. Accuracy of 1ms can be maintained for approximately 50 days, 5 days, or 1 hour for the cesium, rubidium, or crystal oven clocks, respectively, if the latter two clocks can be calibrated at least once a year. Another example is a 2400-bps synchronized data system, which imposes a bit timing accuracy of about 40 ms and can be maintained for 2 days, 5 hours, or 3 minutes (for cesium, rubidium, or crystal oven clocks, respectively) during the absence of a signal.

The circuits which interface between processor and clock are an important consideration since clock timing may be degraded by the number of computer cycles required for functional control. Therefore, a communications clock will probably require a direct-memory-access (DMA) interrupt interface with the internal microprocessor bus.

Rubidium gas oscillators and portable clocks of hand luggage size are available from a number of vendors, adequate testimony to the developing market.

**CONCLUSION**

Radio equipment is electronically controllable by small computers. New communications systems that depend on electronic agility are in experimental stages. Radio frequency circuits with amazing performance specifications are being developed for these computer-controlled systems. A new architecture is evolving that provides increased agility and adaptivity in HF/VHF systems. Software, firmware, and processor development provides increasing support to system control functions and to measurement and modulation and demodulation functions.

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**Table 1. Typical Receiver Specifications.**

| | |
|---|---|
| Noise Figure | 15 dB |
| Linearity | +25 dBm third-order intercept for 2 equal tones at 30 kHz and 60 kHz from channel |
| Reciprocal Mixing | 90-dB suppression of tone 30 kHz from 3 kHz channel |
| Tuning Speed | Less than 25 ms |
| Extraneous Responses | 80-dB suppression of images, IF frequencies, etc. |
| Spurs | Not more than ten spurs 3dB above receiver noise<br>None greater than 10 dB above receiver noise floor |
| Bandpass Time Delay | Less than 0.5 ms variation between 600 Hz and 2700 HZ |
| Phase Jitter | Less than 3° averaged over 10 ms in 3kHz channel |
| Controls | All remoted via RS-232C, IEEE 488, or BCD parallel |
| Built-in Test | Function level |

**Table 2. Packet Radio Evolution.**

| | Experimental Packet Radio | Improved Packet Radio | Value Eng Packet Radio | Low Cost Packet Radio |
|---|---|---|---|---|
| Development | 1974-1977 | 1976-1977 | 1980-1982 | 1981-1983 |
| Number | 28 | 27 | 18 | 2 |
| Net Timing | Asynchronous | Asynchronous | Asynchronous | Asynchronous |
| PN Code | Fixed | Fixed | Fixed | DES Driven |
| Processor | IMP-16 | TI-9900 | TI-9900 | MC68000-Like |
| Memory (bytes) | 7 K | 16 K | 22 K | 32-64 K |
| Size (ft3) | 1.3 | 1.7 | 1.0 | 0.15 |
| Weight (lb.) | 40 | 50 | 30 | 12 |
| Prime Power (W) | 40 | 100 | 100 | 25 |

RF Band: 1710-1850 MHz; XMIT Power: 10W: Bandwidth: 20 MHz.

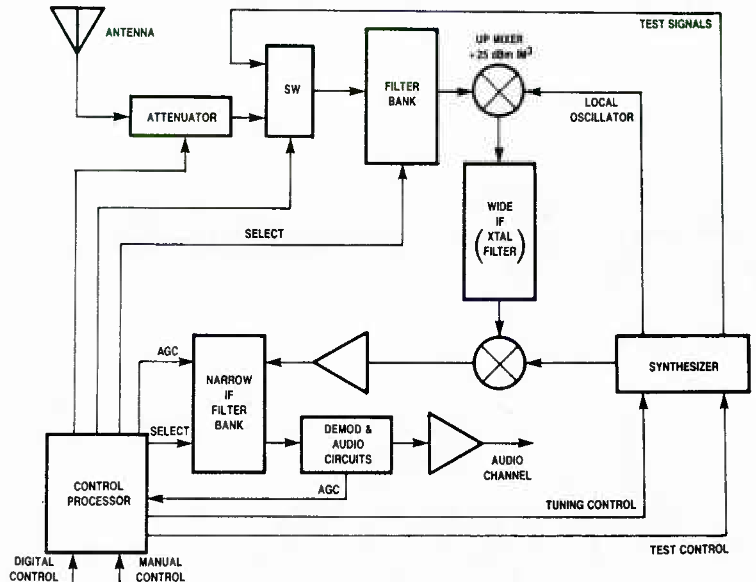



Figure 1. BLOCK DIAGRAM OF A CONTEMPORARY RECEIVER

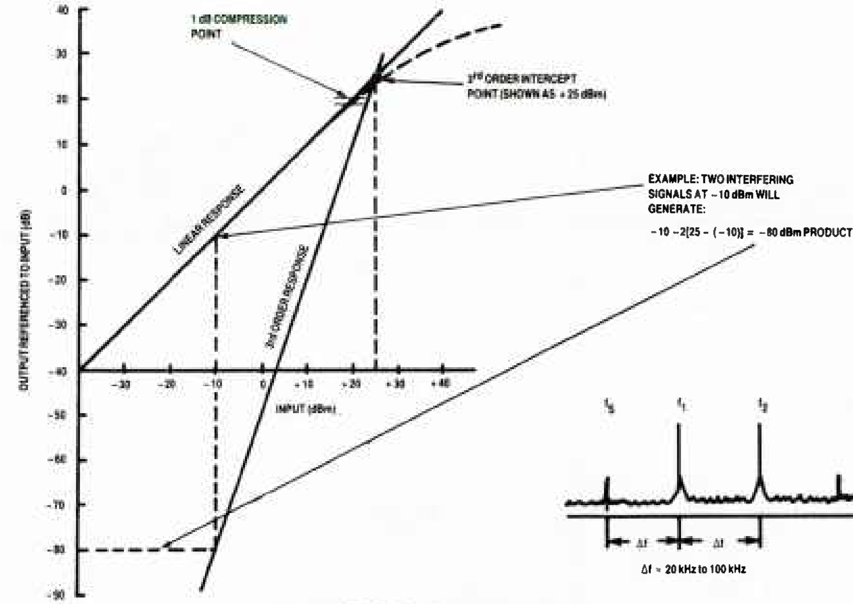



Figure 2. THIRD-ORDER INTERMODULATION INTERFERENCE

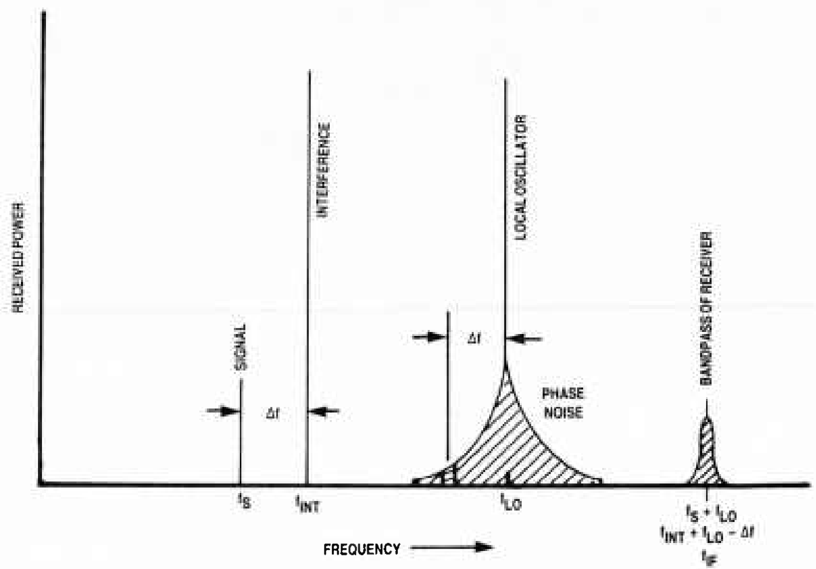



Figure 3. RECIPROCAL MIXING OF INTERFERENCE AND LOCAL OSCILLATOR PHASE NOISE ENTERS THE RECEIVER IF BANDWIDTH

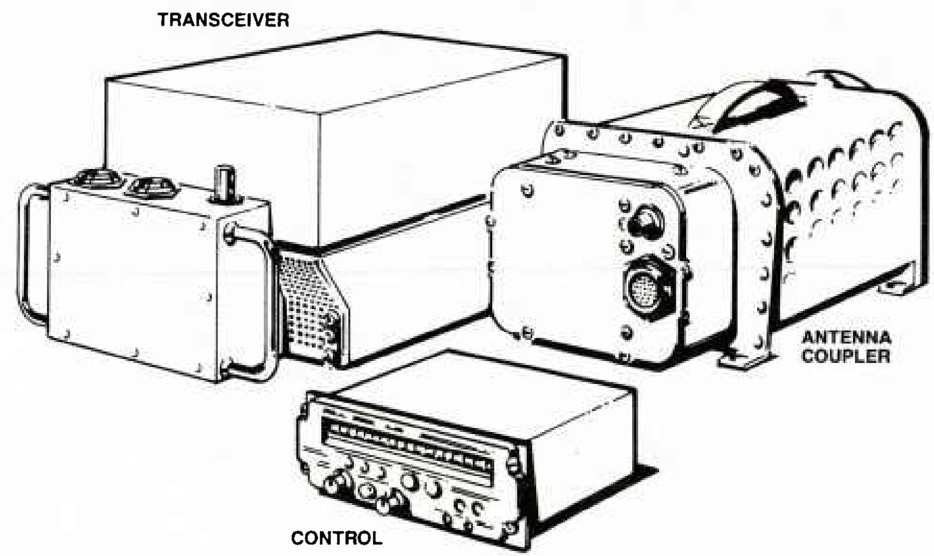



Figure 4. AIRBORNE TRANSCEIVER COMPONENTS

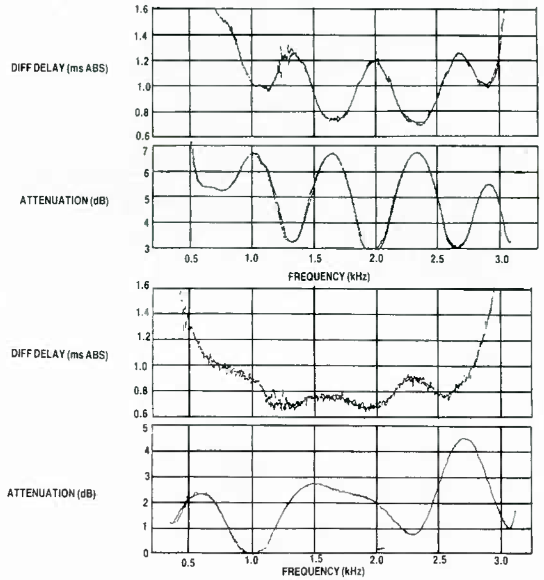



Figure 5. IF FILTER RESPONSES OF ONE TRANSCEIVER MODEL

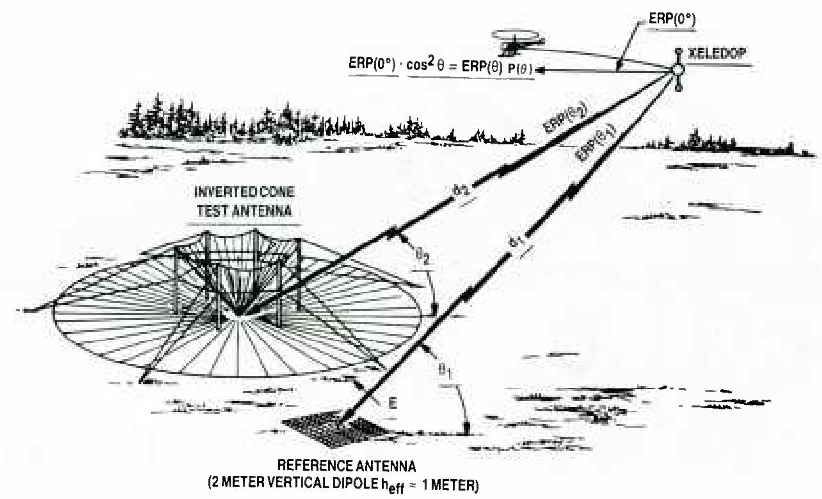



Figure 6. ANTENNA TESTING CONFIGURATION

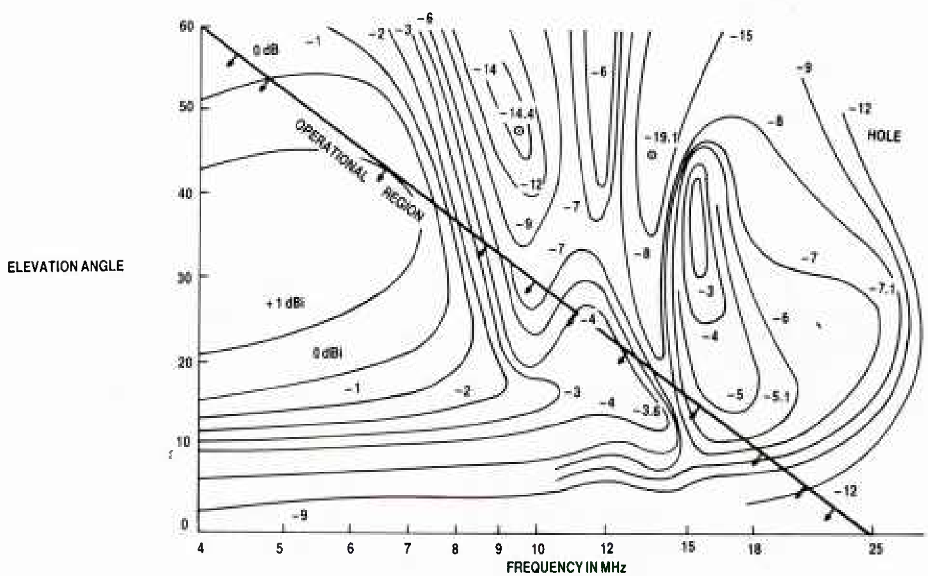



Figure 7. INVERTED CONE ANTENNA VERTICAL GAIN

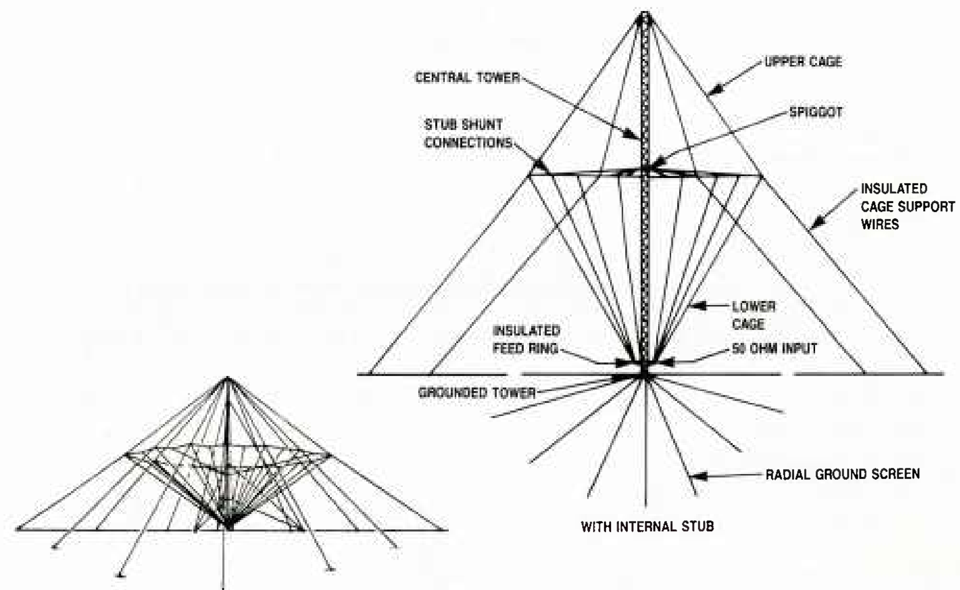



Figure 8. BICONICAL VERTICAL ANTENNAS

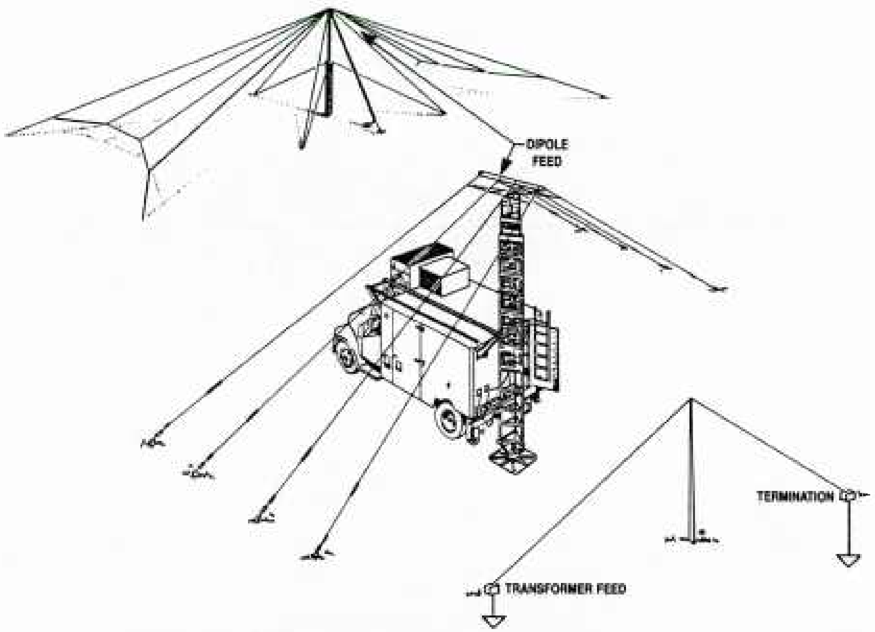



Figure 9. BROADBAND HORIZONTAL AND VERTICAL SINGLE POLE ANTENNAS

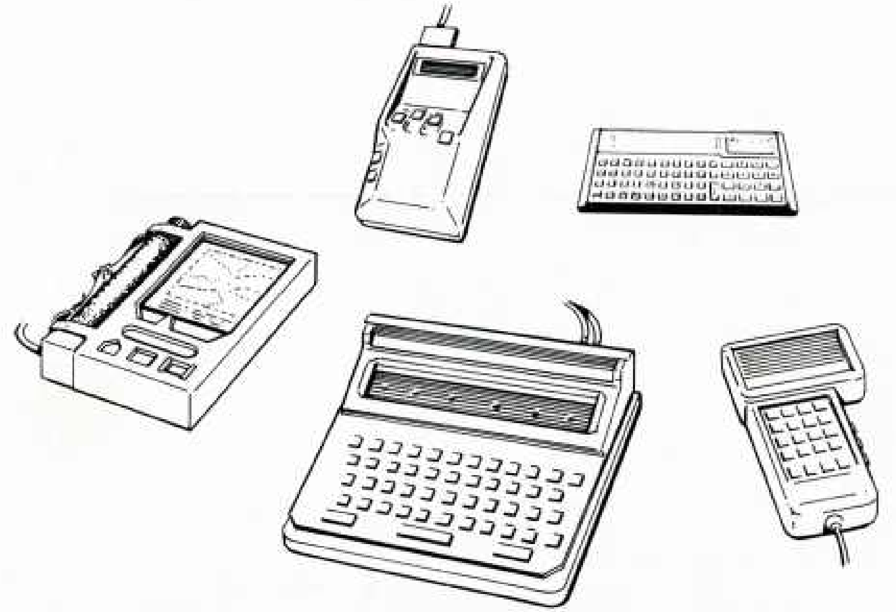
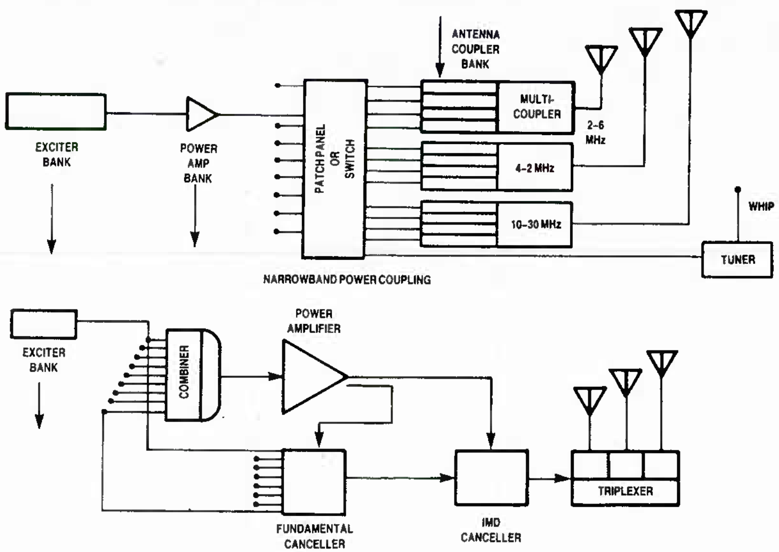


Figure 10. SMALL INPUT/OUTPUT DEVICES




WIDEBAND EXCITER COUPLING

Figure 11. TRANSMITTER ARCHITECTURES FOR SHIPS

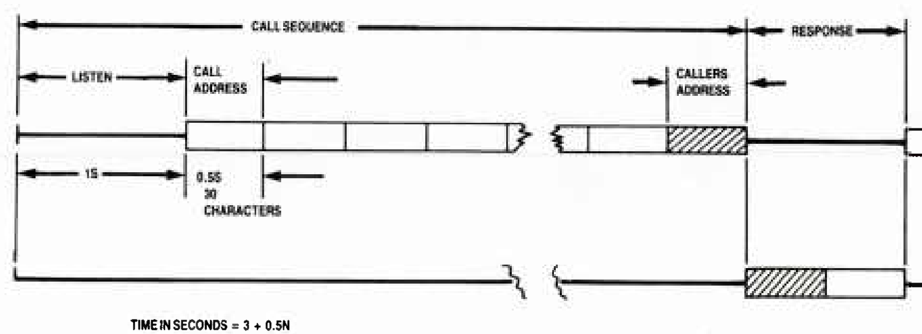



Figure 12. CALLING SEQUENCE FOR N CHANNEL SCANNING

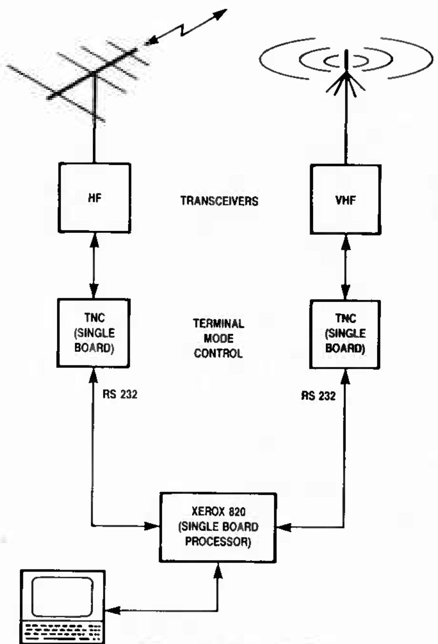



Figure 13. AMATEUR PACKET RADIO GATEWAY

Figure 15. PORTABLE METEOR BURST TERMINALS

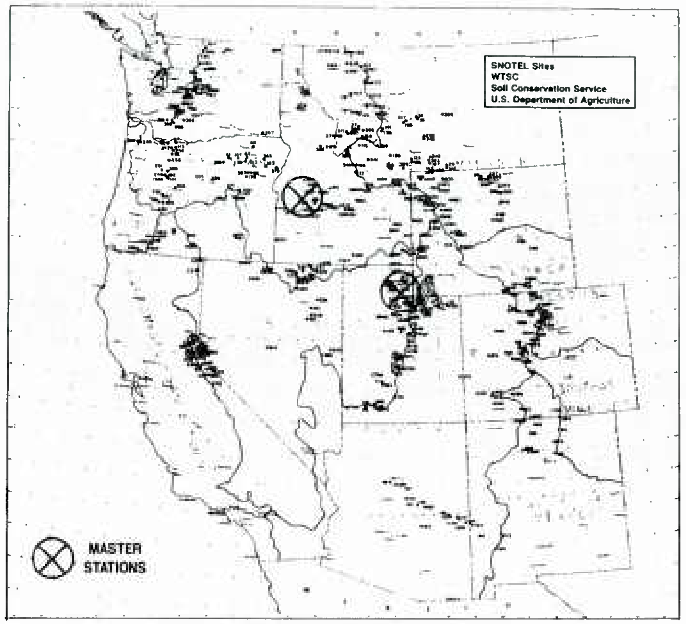



Figure 14. WESTERN UNITED STATES SNOTEL SITES

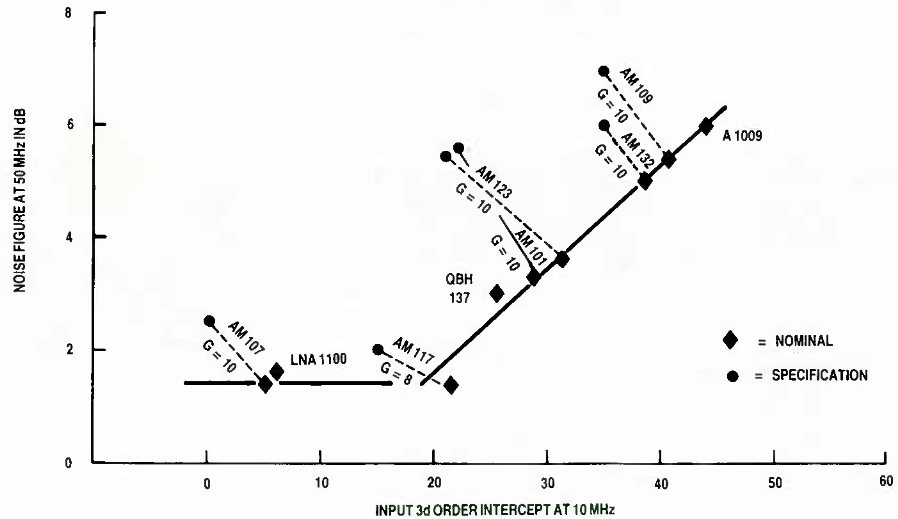



Figure 16. SELECTED NOISE FIGURE VS. INPUT THIRD-ORDER INTERCEPT PERFORMANCE OF RF AMPLIFIER MODULES

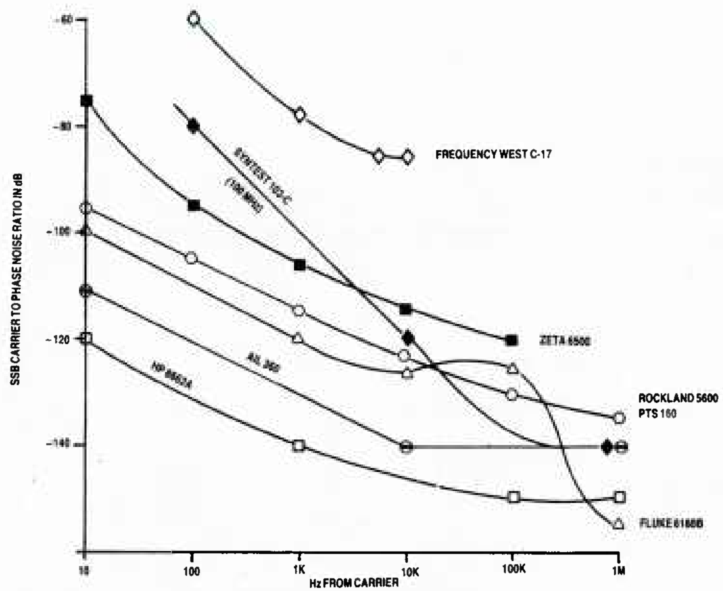



Figure 17. PHASE NOISE OF REPRESENTATIVE SYNTHESIZERS

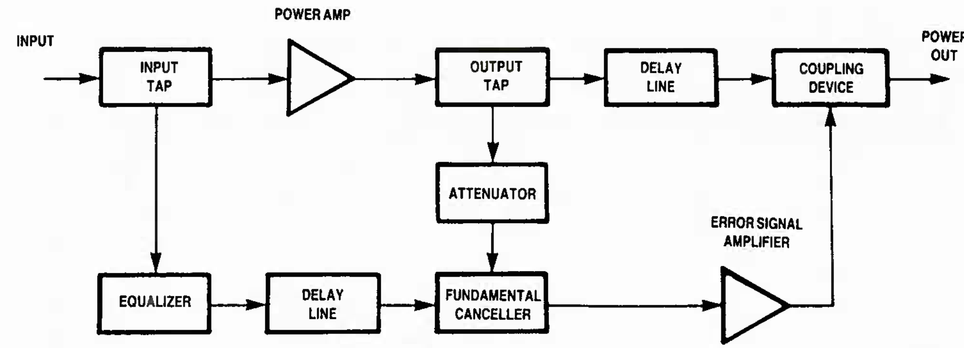



Figure 18. FEED FORWARD AMPLIFIER LINEARIZATION

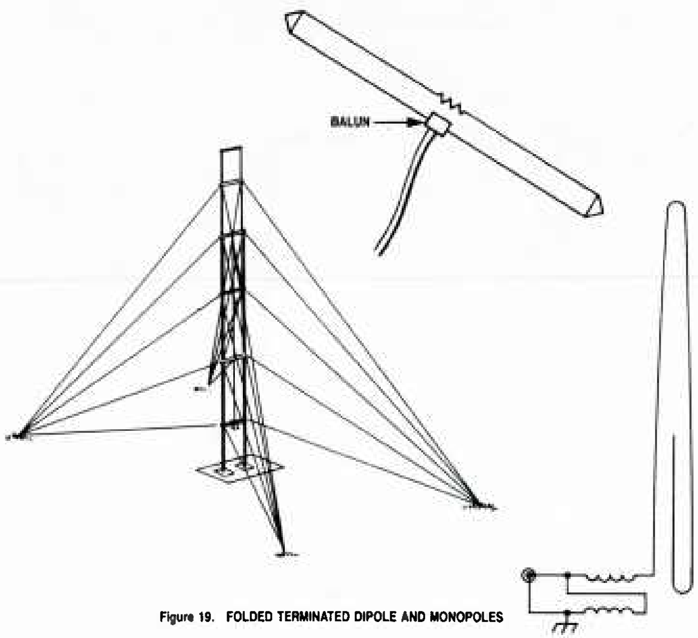



Figure 19. FOLDED TERMINATED DIPOLE AND MONOPOLES

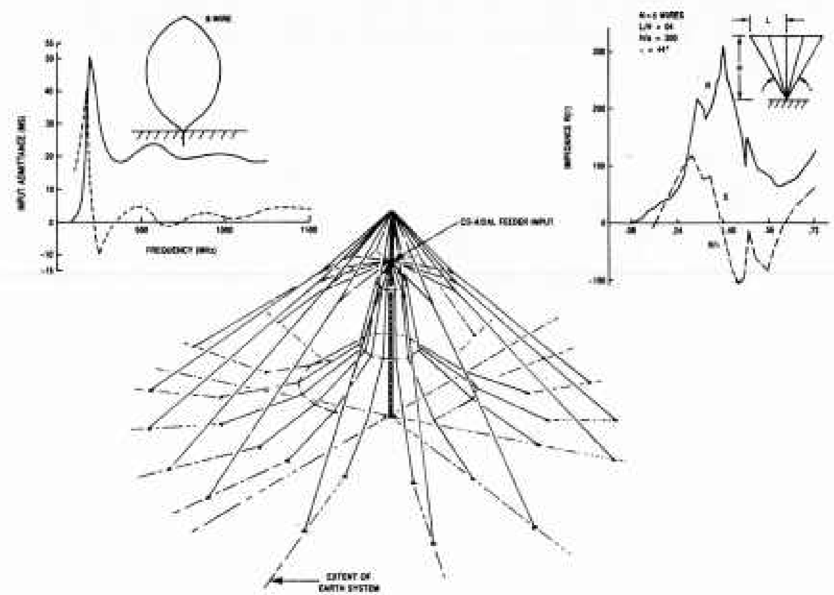



Figure 20. VERTICAL BROADBAND WIRE ANTENNAS

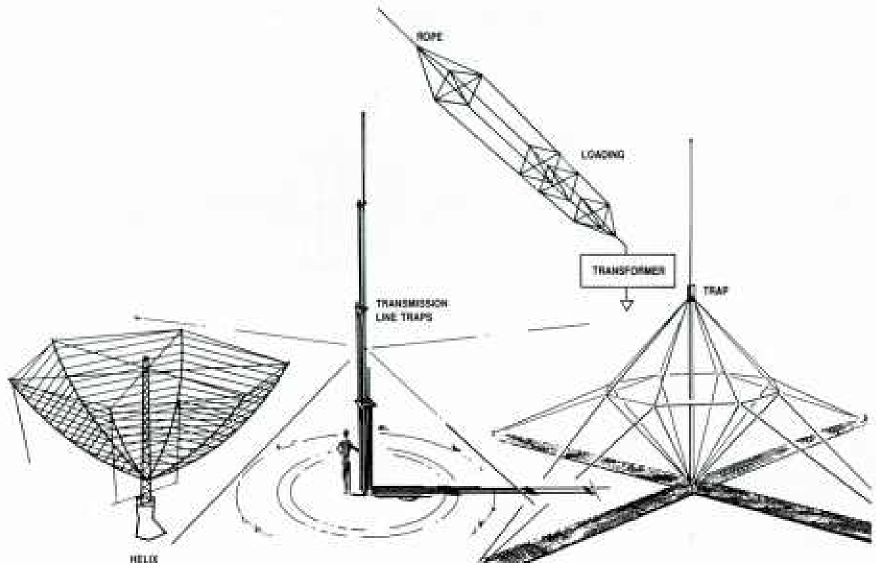



Figure 21. COMPACT WIDEBAND LOADED MONOPOLE ANTENNAS

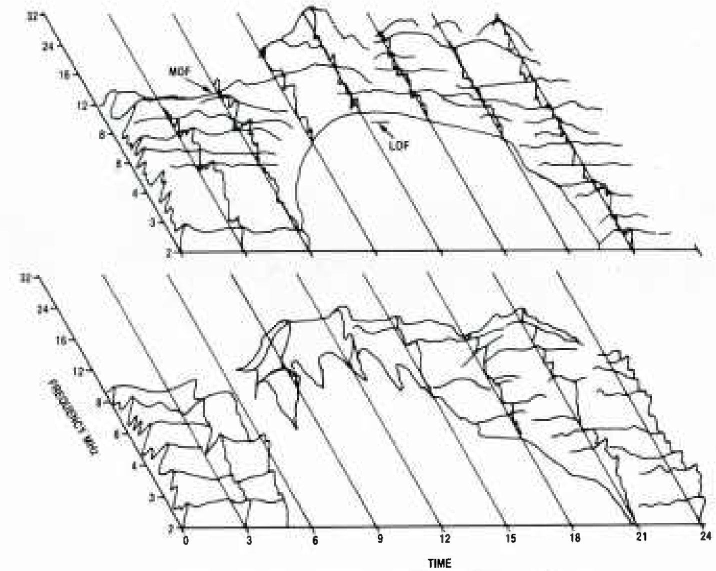



Figure 22. EXAMPLES OF SIGNAL STRENGTH AS A FUNCTION OF FREQUENCY DURING A DAY

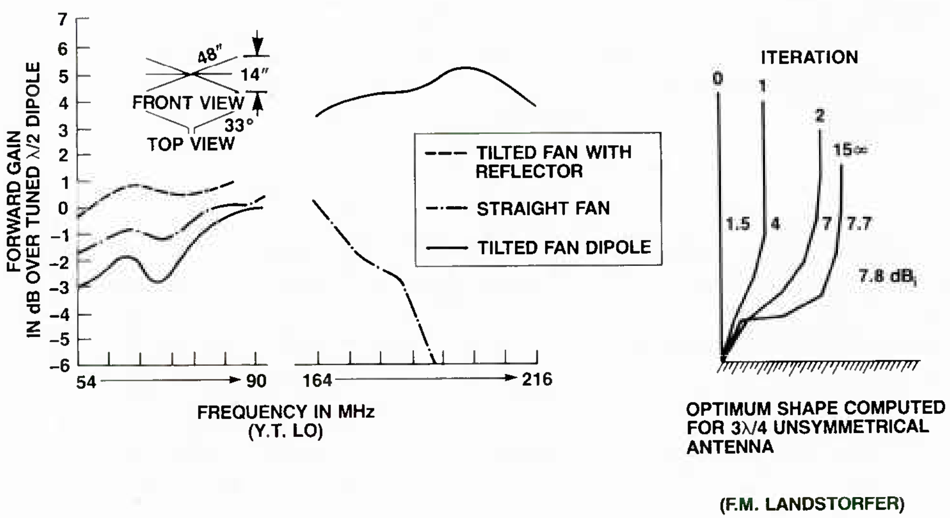



Figure 23. FORWARD GAIN OF TILTED ANTENNAS

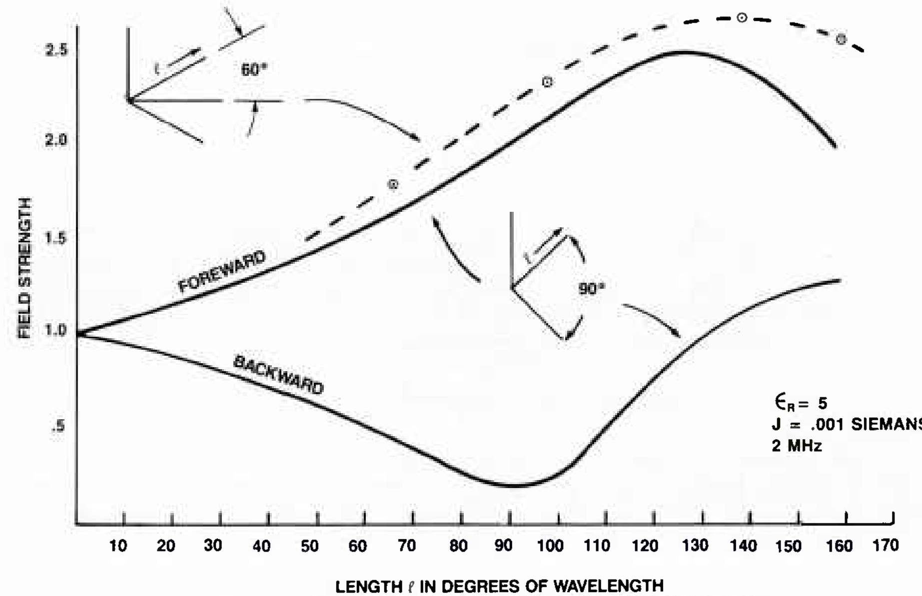



Figure 24. CALCULATED ENHANCEMENT AT 3° ELEVATION ANGLE VS. GROUND SCREEN LENGTH
(C. L. CHEN, W. L. WEEKS)

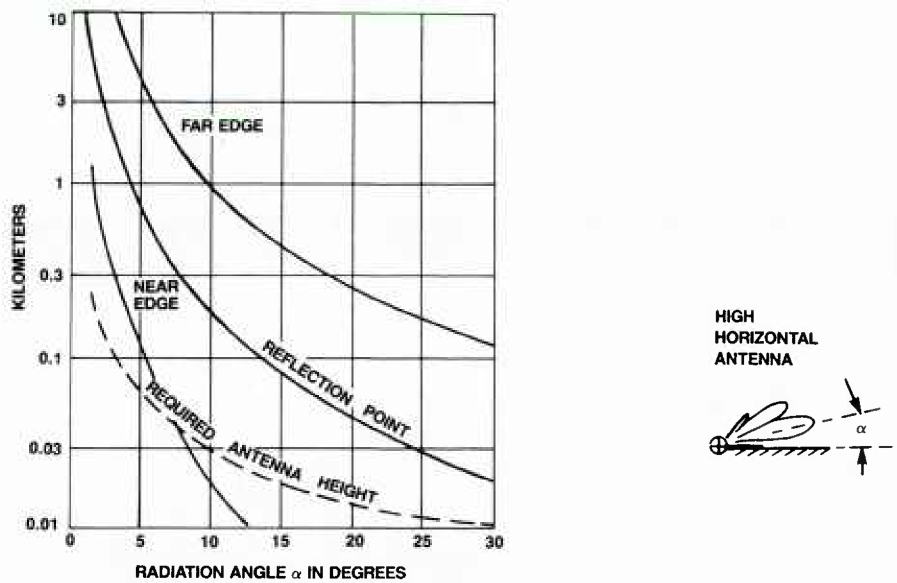



Figure 25. DISTANCE FROM ANTENNA TO FRESNEL REFLECTION ZONE AS A FUNCTION OF RADIATION ANGLE

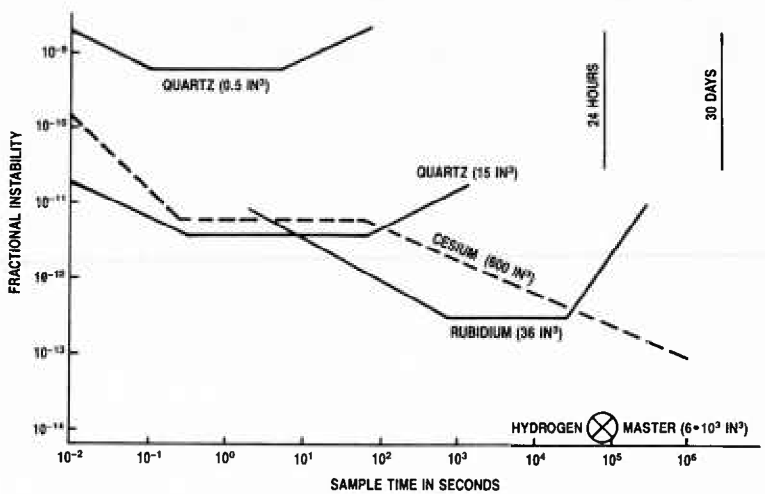



Figure 27. FRACTIONAL STABILITY OF SEVERAL CLOCK TYPES

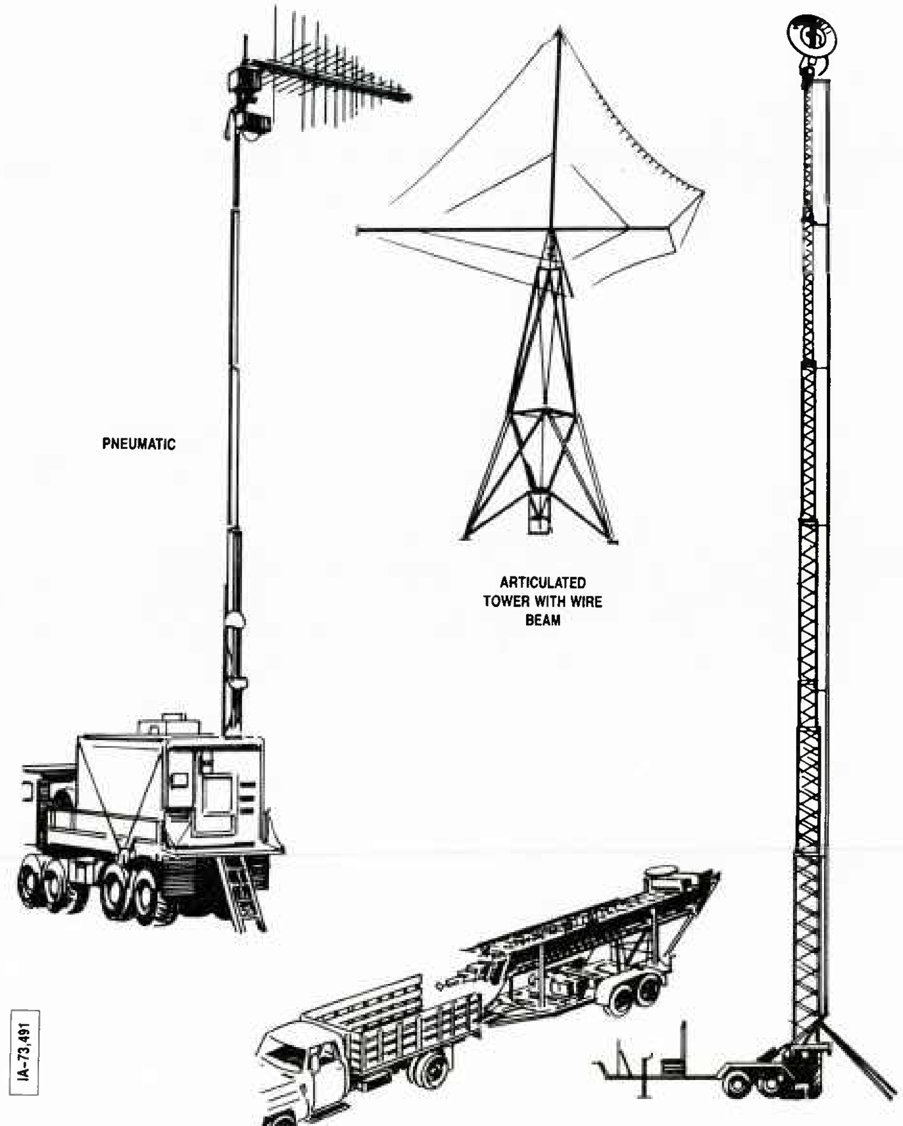



Figure 26. PORTABLE MASTS AND TOWERS

# PROPAGATION II. PROBLEMS IN HF PROPAGATION

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ABSTRACT

The ionosphere is not a perfect reflector for HF-waves, and the lecture will review some of the resulting propagation problems. Some of these are encountered during undisturbed ionospheric conditions, such as multipath reflections, but most problems are associated with geophysical disturbances. Solar flares and associated magnetic storms cause absorption, low MUF (Maximum Useable Frequency), scatter due to irregularities etc. The ionosphere is particularly variable in high latitudes where auroral phenomena influence the reflecting properties of the ionospheric layers. The lecture discusses the presently available short term forecasting techniques, and it also deals with possible ways of minimizing the effects of ionospheric disturbances, such as path and time diversity, the use of early warnings, and back-up systems.

## 1. INTRODUCTION AND OUTLINE

The ionosphere is by no means a perfect mirror for HF-radio waves, and the purpose of the present lecture is to discuss some of the most important propagation related problems in HF communication. Some of these problems are caused by the inherent properties of the undisturbed normal ionosphere and its variations in time and space. Other problems are related to disturbances in the ionosphere causing irregular and often sudden changes of the propagation medium from its normal state. The lecture will deal with these problems separately. The geophysical disturbances are most frequent and severe in high latitudes, and a special section will discuss high latitude propagation.

In the companion lecture we have discussed the long term predictions of monthly mean ionospheric parameters. There is a need for short term forecasts for periods of days, hours, or even minutes, to allow the operator to adapt to rapidly changing conditions. Some techniques for short term forecasts will be discussed. Finally, discussion will center on the possibilities of avoiding or minimizing the propagation problems.

## 2. PROPAGATION PROBLEMS DURING NORMAL, UNDISTURBED IONOSPHERIC CONDITIONS

In this section we shall discuss some propagation problems encountered in the normal, undisturbed ionosphere. The distinction between undisturbed and disturbed conditions is bound to be somewhat arbitrary. For the purposes of our discussion, however, the ionosphere is undisturbed, or normal, when no distinct geophysical events, such as solar flares, or magnetic storms can be identified. Under such conditions, the critical frequencies, the layer heights and the ionospheric absorption show no large deviations from the monthly mean values, as predicted by the ionospheric models discussed in the companion lecture.

### 2.1 Multipath propagation

Figure 1 illustrates a typical situation in which several propagation modes exist simultaneously, and the received signal is a vector sum of electromagnetic waves arriving at the receiver from different angles, with different time delays, amplitudes and polarizations. The result may be severe distortion of the signal, for example deep and rapid fading. Such fading may occur because the different signal paths change with time in different ways. In order to detect the resultant signal above the background noise, a larger transmitter power may be needed than in the absence of multipath propagation. Frequency allocation in the HF band is based on a 3 kHz channel spacing. Within a 3 kHz bandwidth the fading may not be correlated at different frequencies. This selective fading can produce distortion if the carrier wave fades to a level below the side bands. Selective fading in a 3 kHz bandwidth channel occurs when the time delay between different paths exceeds 500 μs, that is when the distance in path length for the modes exceeds 150 km.

Different modulation techniques are affected in different ways by multipath fading. Analogue voice transmissions are relatively insensitive to selective fading, whereas digital data transmission is strongly affected by fading, even for time delay differences as short as 50 μs (path length difference of 15 km).

Fading due to interference between different rays frequently occurs near the skip distance, that is when the transmission frequency is near the MUF of the circuit. As discussed in the companion lecture, high and low angle rays tend to converge near the skip distance. Fading may also be caused by the interference of ordinary and extraordinary magnetoionic components of the wave, since these components follow different paths through the ionosphere. Figure 2 shows an oblique incidence ionogram in which the delay time over the path is recorded versus frequency. The situation is complex with many modes present. Figure 3 illustrates a similar situation as observed at one frequency (8 MHz).

### 2.2 Effects of ionospheric tilts

The models described in the companion lecture assumed that the ionospheric layers are parallel to the earth's surface. Tilts, that is horizontal gradients in the electron density, are, however, very important for radio wave propagation. Large scale horizontal gradients, such as the day-night transition region near the terminator, may cause deviations of the ray path from the great circle path between transmitter and receiver, with corresponding changes in signal characteristics. Similar problems are also encountered for propagation along the auroral zone. Such propagation is difficult to model, because the effects depend strongly on path geometry and because the ionosphere changes rapidly with time. Tilts with medium scales (a few hundred kilometers) must also be considered as a normal feature of the ionosphere. Such horizontal gradients are associated with travelling ionospheric disturbances (TIDs). These disturbances have wavelike structures with periods of the order of 10-30 minutes, and cause focusing and defocusing of the radio waves. Figure 4 shows an example of a strong TID influencing the virtual height of reflection in the F-layer (Georges, 1967). Röttger (1978) has studied the occurence rate of TIDs in the equatorial region. Figure 5 shows the results, which indicate the presence of such phenomena 80% of the observation time during certain periods. Rice (1976) has measured the direction-of-arrival-errors at HF for a path through the auroral zone. Figure 6 shows an example of the distribution of measured bearings. The medium scale tilts will in general cause fading and distortion of the signals.

### 2.3 The effects of irregularities

In our previous discussion we have implictly assumed that the ionosphere is smooth over the area (the first Fresnel zone) which reflects the wave. In practice the electron density may have small scale structures both in time and space, superimposed on the general background. These irregularities act as scatterers of the incident radio wave, and they may drift across the beam of the transmitter. The effect of radio wave reflection from an irregular ionosphere is that (as illustrated in Figure 7) the received wave appears to the observer to come from an extended area, rather than from a point. The ionosphere will act as a diffracting screen, projecting a moving diffraction pattern on to the earth's surface. The observed signal amplitude will fluctuate rapidly, and the fading will have certain statistical characteristics which depend upon the properties of the screen. If the ionospheric screen is random so that none of the individually reflected wavelets dominate, the amplitude will have a Rayleigh, asymmetric, probability distribution function, as shown in Figure 8. If, however, the received signal has a strong steady component from a mirror-like reflection, and in addition weak random components from irregularities, a more symmetric "Rice" distribution will be observed as indicated in the figure. Both types of distributions are observed in practice, and a knowledge of the statistical properties of the signal amplitude and phase is important for the determination of required signal-to-noise ratio.

Ionospheric irregularities occur in all layers. In the F-layer ionosonde recordings may show "spread F", that is diffuse traces where the signal appears to be returned from a wide range of heights. Spread F occurs most often at frequencies near the critical frequency, in the night time ionosphere. It is more common in high latitudes and near the equator than in middle latitudes.

Irregularities in the E-region is often associated with sporadic E, which will be discussed in the next subsection. Below the E-region HF signals may be scattered from irregular structures. Some of these are caused by meteors, and this type of scatter will be discussed in Section 6.

### 2.4 Sporadic E

In addition to the regular ionospheric layers there are several transient or irregular layers, of which the sporaidc E-layer is the most important. Although this lecture deals with <u>problems</u> in HF-propagation, it should be pointed out that efficient use of sporadic E reflections may improve HF transmissions.

Sporadic E ($E_s$) occurs at heights between 95 and 120 km as a layer which may have much higher critical frequency than the regular E-layer. Sometimes the ionogram indicates that the $E_s$-layer is thick and opaque with a well defined maximum, at other times the layer may be thin, patchy and partly transparent, so that higher layers are observed through the $E_s$. The $E_s$ has different characteristics in different latitudinal zones, and there may be several physical mechanisms governing the behaviour of these layers. It is believed that windshears in the neutral air, acting on the E-region plasma can create sporadic E-layers (Chimonas & Axford 1966).

The sporadic E-layer is a problem in HF-communications because of its irregular and (as yet) unpredictable behaviour.

The observed statistical occurence rate of sporadic E is included in current prediction schemes for HF-communications such as IONCAP, (see companion lecture). Figure 9 shows the statistics of $E_s$ with critical frequency $fE_s > 5$ MHz. As will be seen, $E_s$ is a night time phenomenon in the auroral zone, and a daytime phenomenon in the equatorial zone. At middle latitudes there is a strong seasonal variation with maxima near the equinoxes.

One of the most important effects of $E_s$ on HF propagation is the screening effect. As illustrated in Figure 10, the sudden occurence of an $E_s$ layer can prevent the signal from reaching the F-layer, and thus limit the range of the transmission. The patchy and irregular structure of an $E_s$-layer may introduce rapid fading, as discussed in the previous section.

### 2.5 Non-linear effects in the ionosphere

The ionosphere is a non-linear plasma, that is, a radio wave travelling through the medium changes the properties of the plasma, so that the wave influences its own propagation as well as the propagation of other waves travelling through the same region. The most important non-linear process, from the point of view of propagation, is most easily understood by considering the theory for the refractive index of a radio wave in a plasma. The complex refractive index has real and imaginary parts

$$n = \mu(N_e,\nu)-i\chi(N_e,\nu) \qquad (1)$$

which both depend upon the electron density $N_e$ and on the collision frequency $\nu$ of an electron with neutral molecules and ions. In the companion lecture we discussed how the electromagnetic energy carried by the wave is lost to thermal energy in the plasma through the collision process. The collision frequency $\nu$ is proportional to thermal energy, that is to the temperature of the gas, and thus absorption of a radio wave leads to larger collision frequency, which in turn leads to greater absorption. This self-modulation may distort a radio signal, and may be important for signals from very powerful transmitters. Cross modulation of signals in the ionosphere was first reported by Tellegen (1933) who had observed that a transmission from Beromünster in Switzerland (650 kHz), received in the Netherlands, was modulated by the signal from the powerful radio station in Luxembourg (252 kHz). The effect is often called the Luxembourg effect, and has been shown to occur mainly in the lower ionosphere (E- and D-region) when the collision frequency $\nu$ is large. The degree of self- or cross-modulation depends strongly upon transmitter power and upon frequency, and may be of the order of 1-10% for waves near 2 MHz and transmitter powers of 10-100 kW. (See for example Davies 1969, and articles in AGARD CP 38 1974). Self-modulation may render an increase in transmitter power selfdefeating because it introduces an effective transmission loss (Megill 1965).

Non-linear effects in the ionosphere have been studied studied extensively by many workers, and the reader is referred to AGARD CP 138 and references therein for further information.

## 3 PROPAGATION PROBLEMS ASSOCIATED WITH GEOPHYSICAL DISTURBANCES

The reflecting and absorbing properties of the ionosphere often show deviations, from the regular diurnal and seasonal changes, that can be associated with geophysical disturbances. These disturbances fall into two classes i) those directly associated with solar flares which eject energetic radiation in the form of ultraviolet radiation, X-rays and particle radiation towards the earth, and ii) those associated with changes in the structure and circulation of the earth's neutral atmosphere. Disturbances in the first class are best understood, and their causes and effects have been studied since the discovery of the ionosphere. The possible importance of the second class of disturbance has been realized only in fairly recent years, but the "weather" systems in the upper atmosphere and their relation to lower atmosphere weather and climate are still poorly mapped. The physical mechanisms are not well understood.

Geophysical disturbances and their effects upon the propagation medium have been reviewed by Thrane (1979) and by Thrane et al (1979), and in this lecture only a very brief review of the HF-propagation problems encountered during such disturbances will be given. There are many ways of classifying the disturbances, none of them very satisfactory. Table 1 lists some disturbances which have important propagation effects.

Entry a) in the table represents the initial effects of a solar flare caused by electromagnetic radiation. Entries b) to e) represent delayed flare effects, due to energetic particles entering the upper atmosphere. These effects are all part of the very complex sequence of phenomena called a magnetic storm. Entries f) and g) are disturbances which may be caused by particle precipitation, but together with entry h) they may also belong to class ii) discussed above, that is they may be triggered by changes in the neutral atmosphere.

In this section we shall deal with some propagation disturbances in middle and low latitudes, and discuss high latitude problems separately in the next section.

### 3.1 Sudden Ionospheric Disturbances (SID)

A sudden burst of X-rays and UV radiation from a solar flare will cause increased electron density in the lower ionosphere, and "black-out" of HF-circuits may occur over the entire sunlit hemisphere. If the black-out is not complete, the signal may suffer sudden frequency shifts and phase changes, due to a sudden lowering of the reflection point in the ionosphere. Although the disturbance is normally short-lived (~ 1 hr) it may cause serious disruption of traffic, and because of its sudden and unexpected onset, may cause the operator to search for technical faults in his system.

### 3.2 Magnetic storms

During such disturbances currents flowing in the ionosphere cause changes in the earth's magnetic field. In our context the most important mid-latitude effect of a magnetic storm is the decrease, of the maximum electron density $N_{em}F2$ in the F-layer. Such decreases lead, of course, to decrease of the MUFs. Storms are also often associated with absorption, due to influx of precipitating particles at middle and high latitudes. The absorption enhancements increase the LUFs, and the result is a narrowing of the frequency range available over an affected circuit. Figure 11 shows typical changes in $N_{em}F2$ in different latitudinal zones (Matsushita 1959). Strong and erratic time variations in MUF can cause major communication problems. Both MUF variations and storm associated absorption (see entries d) and c) in Table 1) are strongest and most likely in latitudes above $40^\circ$-$50^\circ$.

### 3.3 Winter anomaly in ionospheric absorption

At middle latitudes ($35^o$-$60^o$), ionospheric radio wave absorption in winter does not follow the simple solar zenith angle dependence to be expected from the averaged summer observations. The general background of winter absorption is enhanced relative to summer values at the same solar zenith angles, and in addition days or groups of days occur in winter when HF-absorption is greatly enhanced above this background. The consequences for communication can be serious, since absorption values of 60 dB in excess of normal may occur on MF and HF-circuits. Figure 12 shows typical values of absorption during a winter period (Schwentek, 1971). The horizontal scales of the disturbed regions are of the order of 1000 km, and the magnitude and frequency of absorption events increase with increasing latitude. It now seems clear that the large and variable absorption in winter may be associated with particle precipitation (Sato 1980, Manson 1981 and Sato 1981) and that there also exists a "meteorological type" winter anomaly. This type is associated with neutral atmosphere circulation and planetary waves, which through creation of turbulent transport of minor constituents (nitric oxide, NO), influences the ionization balance in the lower ionosphere (Offermann 1982). The physical models suggested to explain the "meteorological type" anomaly are still crude, but may point the way towards an understanding of the propagation effects of other dynamical phenomena, such as the stratospheric warming. Increases of D-region electron densitites of a factor of up to 10 have been observed during stratospheric warmings. (Belrose 1967, Rowe et al 1969).

## 4. PROPAGATION PROBLEMS IN HIGH LATITUDES

Disturbances in high latitudes merit special attention because the simplicity and mobility of HF-communication equipment make this type of communication particularly useful in remote areas, and for mobile units. This section will discuss the most important high latitude disturbances and their effect upon communication systems.

### 4.1 Polar cap absorption (PCA) events

After certain types of major flares the polar regions are illuminated by high energy protons and alpha particles which penetrate into the lower ionosphere and cause wide-spread and long-lasting disruptions of HF-communication circuits. As seen from Table 1 such disturbances do not occur often, but they may last for periods of up to a week to ten days, and may cover the entire polar caps down to latitudes of about $60^o$. The absorption may be severe, up to 10-20 dB at 30 MHz has been observed. This means that HF skywave communication systems in the polar regions may be rendered completely useless for long periods during such events. This fact has to be faced both by military and sivilian users, and back-up systems should be available where necessary. Possible back-up systems will be discussed in Section 6.

### 4.2 Auroral absorption

Auroral phenomena are often associated with radio black-outs. While the visual aurora itself is caused by soft electrons (energies 1-10 keV) the enhanced absorption is caused by electrons with energies in excess of 10 keV penetrating into the D-region. Figure 13 shows a map of the statistical occurence rate of auroral absorption measured by riometers at 30 MHz. Note that the absorption is strongest in the auroral zone and has a variation in magnetic time with a maximum in the early morning hours. Table 2 indicates the ratio between the oblique path absorption for a signal propagating in a 1F mode over a path of 450 km, and the riometer absorption at 30 MHz.

| Frequency MHz | Ratio = $\frac{\text{Oblique path absorption (dB)}}{\text{Riometer absorption (dB at 30 MHz)}}$ |
|---|---|
| 2.5 | 128 |
| 3.5 | 80 |
| 8 | 22 |
| 15 | 7 |

Table 2 Approximate relation between oblique incidence and riometer absorption

We note that auroral absorption may have severe consequences for radio circuits crossing the auroral zone. Strong auroral absorption is, however, often limited geographically to patches of a few hundred kilometers in extent, and the duration is typically 1/2 to a few hours.

### 4.3 The high latitude E- and F-region

The morphology of the high latitude E- and F-region has been reviewed by Hunsucker (1979). Both layers are characterized by great variability in time and space. Figure 14 shows an example of E-region electron density variations during an auroral event. During very brief periods the E-layer critical frequency is up to 15 MHz, which means that for a MUF factor of 5 (see companion lecture) the layer could support MUFs of more than 70 MHz. That the auroral E-layer sometimes can support VHF-propagation is supported by Figure 15, which shows maximum observed frequencies over paths from College Alaska to Greenland and Norway.

The F-region also shows horizontal gradients, of particular interest is the F-region trough, which is a night time minimum in the variation of the F-layer critical frequency with latitude. The trough marks a transition between the mid-latitude and high latitude ionosphere. Figure 16 (Besprozvannaya et al 1979) shows the variation of the foF2 (normalized) with latitude for all months during 1964. Note the "wall" of ionization occurring poleward of the trough. This sharp gradient could cause reflections of radio waves, and result in deviations from propagation along the great circle between circuit terminals.

Propagation in the auroral regions may introduce rapid fading. Figure 17 shows examples of fading observed on an auroral and a mid-latitude path.

Note that the ionospheric models used for long term prediction purposes do not properly allow for the variability in time and space of the high latitude ionosphere. The data base is certainly inadequate for detailed modelling, and much more work is needed before useful models can be developed.

4.4 Some results of HF-transmission tests at high latitudes

It may be of interest to demonstrate some of the characteristics of HF-propagation in the disturbed high latitude region, by reporting on the results of some transmission tests made over two circuits in Norway (Thrane 1979).

The purpose of the tests was to investigate the importance of frequency flexibility and space diversity in and near the auroral zone. Figure 18 shows the path geometry. The long path (1250 km) normally has its reflection point well south of the auroral zone, whereas the short path (459 km) lies inside the disturbed region. A simple digital test signal was transmitted on four frequencies (2.5, 3.5, 8.1 and 15 MHz) over both circuits and the error rate of the received signals was recorded for selected times of day and season. The data was also divided into periods with different degrees of auroral zone disturbance, as measured by a riometer in the auroral zone (see figure 18). The results from measurements on the four frequencies and over the two paths were combined to simulate different systems. Thus computations were made for five cases to find:

a) The reliability when only one path and one frequency are available (two cases, long and short path).

b) The reliability when four frequencies and one path are available, always using the best frequency (two cases, long and short path).

c) The reliability when two paths and four frequencies are available, using the best frequency and best path at any time. This situation simulates a relay system in which a message may be transmitted from B to R via A. (We have assumed a 100% reliable channel from B to A).

| IONOSPHERIC CONDITIONS<br><br>SYSTEM | QUIET<br><br>I(0-0.1 dB) | MODERATE<br><br>II(0.2-2 dB) | DISTURBED<br><br>III(> 2 dB) |
|---|---|---|---|
| SINGLE FREQUENCY SHORT PATH (3.5 MHz) | 73% | 51% | 15% |
| FOUR FREQUENCIES SHORT PATH (3.5 MHz) | 78% | 72% | 37% |
| SINGLE FREQUENCY LONG PATH (8.1 MHz) | 71% | 71% | 40% |
| FOUR FREQUENCIES LONG PATH | 88% | 85% | 54% |
| FOUR FREQUENCIES TWO PATHS | 90% | 89% | 62% |

Table 3 Reliability for different systems and for different degrees of ionospheric disturbance

Table 3 summarizes the results. We note that, particularly during disturbed conditions, a substantial improvement in circuit reliability may be achieved by means of frequency flexibility and relaying, (that is space diversity).

Monthly means of the measurements have also been compared with monthly mean reliabilities predicted by three different prediction models, Applab III (Bradley 1975), IONCAP (Lloyd et al 1981) and Bludeck (CCIR 1978) (see also companion lecture). Figure 19 shows the results for the two paths during summer noon conditions. There are considerable differences between measurement and prediction, as well as between the different predicion methods. There is a great need for more and more accurate measurements to provide a basis for improvements of the prediction methods.

## 5. SHORT TERM FORECASTING TECHNIQUES

Predictions for periods equal to or less than the solar rotation period, 27 days, are called forecasts. Disturbances in progress are described by warnings. A number of centers throughout the world issue warnings and short term forecasts of solar and ionospheric parameters (Davies 1978). Twelve of these are grouped into the International Ursigram and World Days Service (IUWDS) for the exchange of data and cooperation in solar geophysical observations. In the USA the most important forecasting centers are US Air Force Global Weather Central im Omaha and NOAA Space Environment Forecast Center in Boulder. The USSR also issues short time forecasts (Avdyushin et al 1979). The forecasts normally give qualitative statements on the degree of disturbance expected, for example "moderate HF absorption" or "general improvement of MF propagation conditions". Forecasts for the degree of VLF phase disturbance are often issued for PCA's; predictions for SID's have only recently been attempted experimentally (Swanson & Levine, private communication). Improvements in recent years in ionospheric forecasting are due mainly to more efficient data acquisition and assessment. Examples are the real time propagation assessment systems "Prophet" and the real time navigation monitor developed by the US Naval Ocean System Center (Rothmuller, 1978), (Swanson & Levine, private communication).

One interesting possibility for short term forecasting is to update a simple standard frequency prediction program at intervals by means of some effective index of solar activity (such as the 10.7 cm solar radio noise flux) which can be monitored and distributed to the user. Uffelman and Harnisch (Naval Research Laboratory, private communication) have found that an update about every three hours during a magnetic storm was sufficient to keep the error in the predicted MUF less than 1 MHz.

One of the important questions concerning warnings and short term forecasts is the timely distribution of information to the user in a form that he can readily use. Even priority telex messages may take 24 hrs to reach the user. A working group on D-region prediction (Thrane et al 1979) has recommended the development of telemetry for dissemination of disturbance information to users. Even one bit to indicate the presence or abs ence of a disturbance would be useful. Two methods were proposed as worth considering: a unique, non-interfering modulation could be added to world wide Omega signals, or to HF-signals from WWV or elsewhere to indicate the presence of an event. The former has the advantage that the Omega signal (10 kHz) is continuously available on a global basis, even during total HF-blackout.

## 6. METHODS FOR MINIMIZING HF-PROPAGATION PROBLEMS

From the above discussions it should be clear that propagation problems at HF cannot be eliminated, but may be alleviated by proper system design and by the development of suitable forecast and warning systems. We have pointed out the usefulness of frequency flexibility and path diversity for avoiding problems during disturbances. One of the difficult problems facing an HF-operator is interference from other users of the HF band. Real time channel evaluation is a powerful tool both for avoiding such interferenceand for adapting to rapidly changing ionospheric conditions.

It should be stressed that the ionospheric channel is not always available. During strong natural disturbances such as PCA's or after nuclear explosions in the upper atmosphere, complete HF-blackout may occur for long periods and over wide areas. Wherever high reliability is required, back-up systems to HF-communication are necessary. The difficulty is to design back-up systems that have the simplicity, mobility and low costs of HF-systems. Transmission of VHF-signals via meteor trails is an interesting possibility in this connection. Ionized trails from meteors occur in the height range 80-120 km and radio signals scattered from the trails may be observed over distances of 200 ~ 2000 km. The meteor trails provide large bandwidth, short duration channels, which can be exploited using modern modulation techniques. Simple yagi antennas suffice for this type of circuit.

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Table 1

Ionospheric disturbances

| Disturbance | Propagation effects | Time and duration | Approx occurrence frequency | | Possible cause |
|---|---|---|---|---|---|
| | | | Solar max | Solar min | Possible cause |
| a) Sudden Ionospheric Disturbance (SID) | In sunlit hemisphere, strong absorption, anomalous VLF-reflection, F-region effects | All effects start approx simultaneously Duration ~1/2 hr | 2/week | 2/year | Enhanced solar x-ray and EUV flux from solar flare |
| b) Polar Cap Absorption (PCA) | Intense radiowave absorption in magnetic polar regions. Anomalous VLF-reflection | Starts a few hours after flare. Duration one to several days | 1/month | 0 | Solar protons 1-100 MeV |
| c) Magnetic Storm | F-region effects; increase of foF2 during first day, then depressed foF2, with corresponding changes in MUF | May last for days with strong daily variations | 26/year | 22/year | Interaction of solar low energy plasma with earth's magnetic field, causing energetic electron precipitation |
| d) Auroral Absorption (AA) | Enhanced absorption along auroral oval in areas hundred to thousand kilometers in extent. Sporadic E may give enhanced MUF | Complicated phenomena lasting from hours to days | Usually present (see figure 13) | | Precipitation of electrons with energies a few tens of keV |
| e) Relativistic Electron Precipitation (REP) | Enhanced absorption, VLF-anomalies at sub-auroral latitudes | Duration 1-2 hours | Variation not known usually present to some degree | | Precipitation of electrons with energies of a few hundred keV |
| f) Travelling Ionospheric Disturbances (TID) | Changes of foF2 with corresponding changes of MUF sometimes periodic | Typically a few hours | | | Atmospheric waves |
| g) Winter Anomaly (WA) | Enhanced absorption at midlatitudes | One to several days | 20/year | 20/year | Probably many causes, such as changes in concentration of minor species, temperature changes, particle precipitation |
| h) Stratospheric Warming | Changes in absorption, VLF-anomalies | Days or weeks, in late winter | 2/year | 2/year | Changes in global circulation pattern |

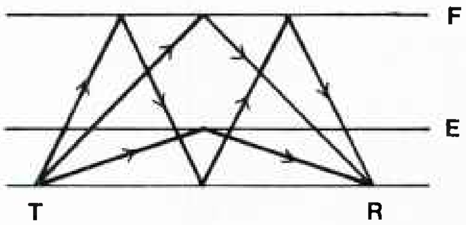



Figure 1 Schematic representation of possible multipath propagation

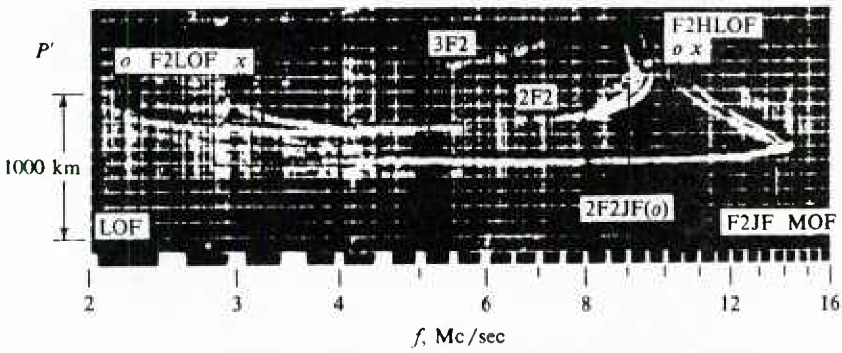
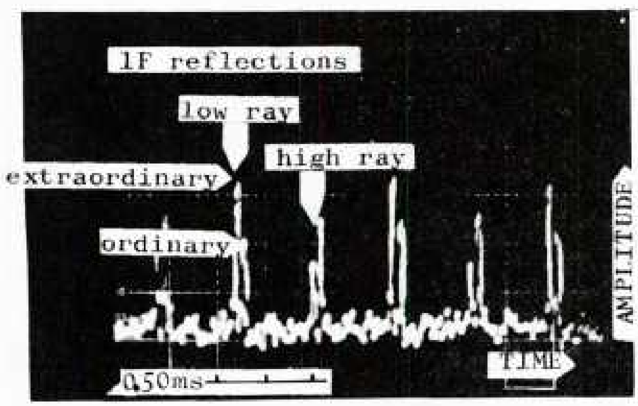



Figure 2

Example of oblique incidence ionogram with multipath propagation (Möller 1964).




f = 8.1 MHz, MUF = 8.2 MHz

Figure 3 Multipath echoes from spread spectrum transmissions at 8.1 MHz (Skaug 1981).

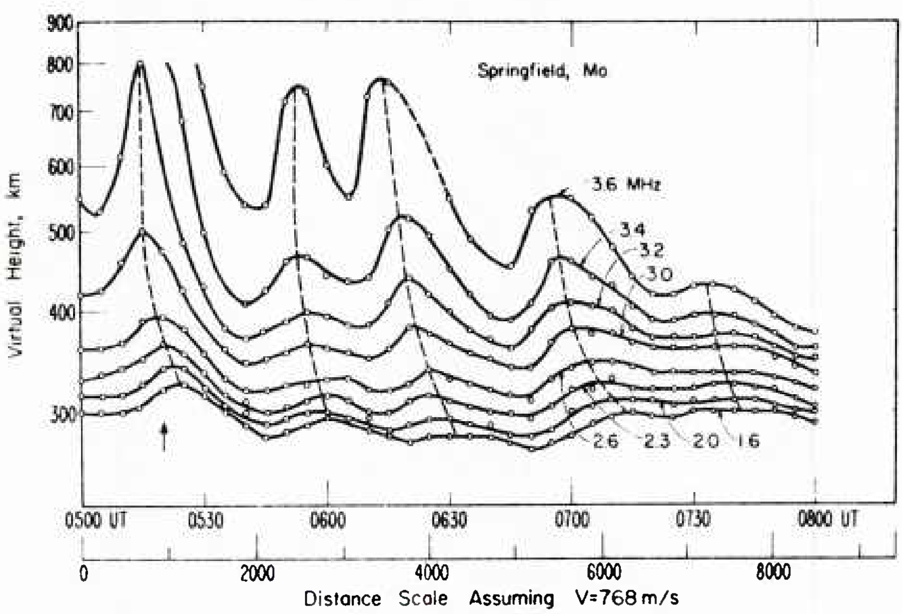



Figure 4 Virtual height variations observed in the F-layer during a large scale travelling ionospheric disturbance (Georges 1967).

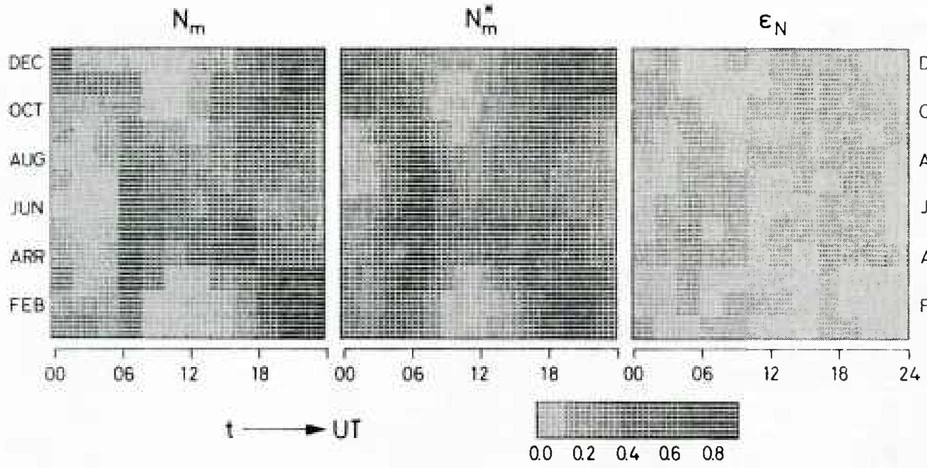



Figure 5 Mean occurrence pattern $N_m$ of TIDs near equator (mean taken over 28 months). The gray scale gives the occurrence frequency between 0 and 1. $N_m$* is the filtered occurrence pattern, taking the Fourier series up to the third harmonic where $\Sigma_N$ is the residual value (Röttger 1978).

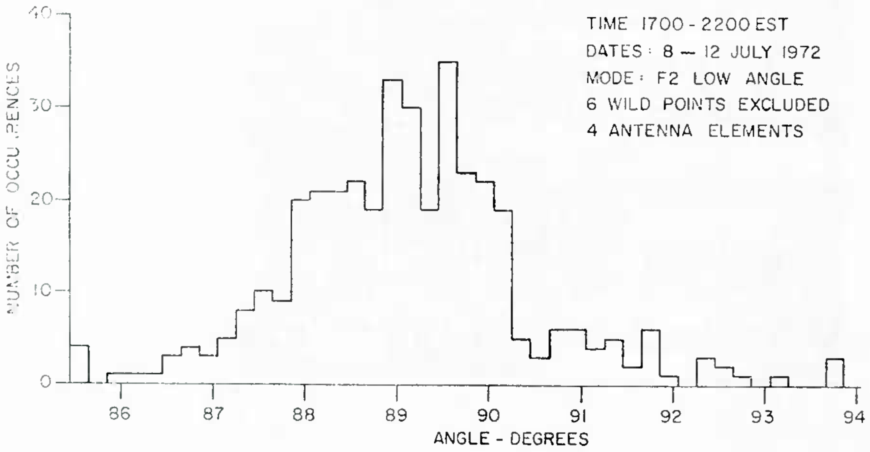



Figure 6 Measured distribution of direction of arrival angle. The angle is here measured relative to the antenna axis. The great circle path to the transmitter has an angle of 89.43° Rice 1976).

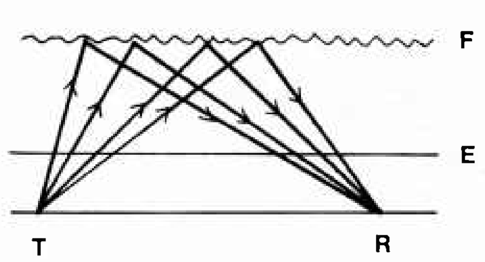



Figure 7 Illustrating reflection from an irregular ionosphere

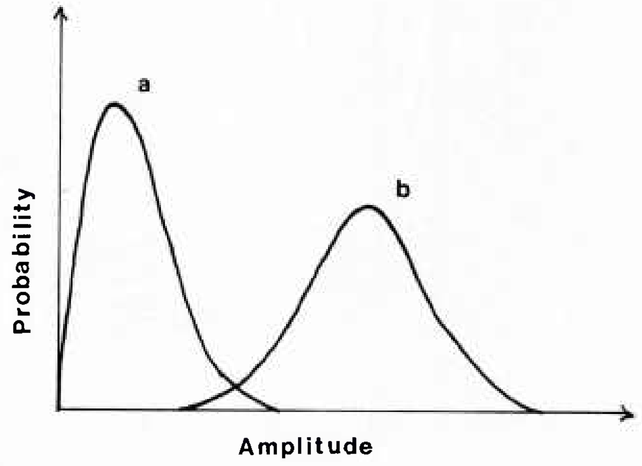



Figure 8 Schematic probability distribution of signal amplitude a) Rayleigh distribution from a random screen b) Rice distribution in the presence of a steady (specular) component (Davies 1969).

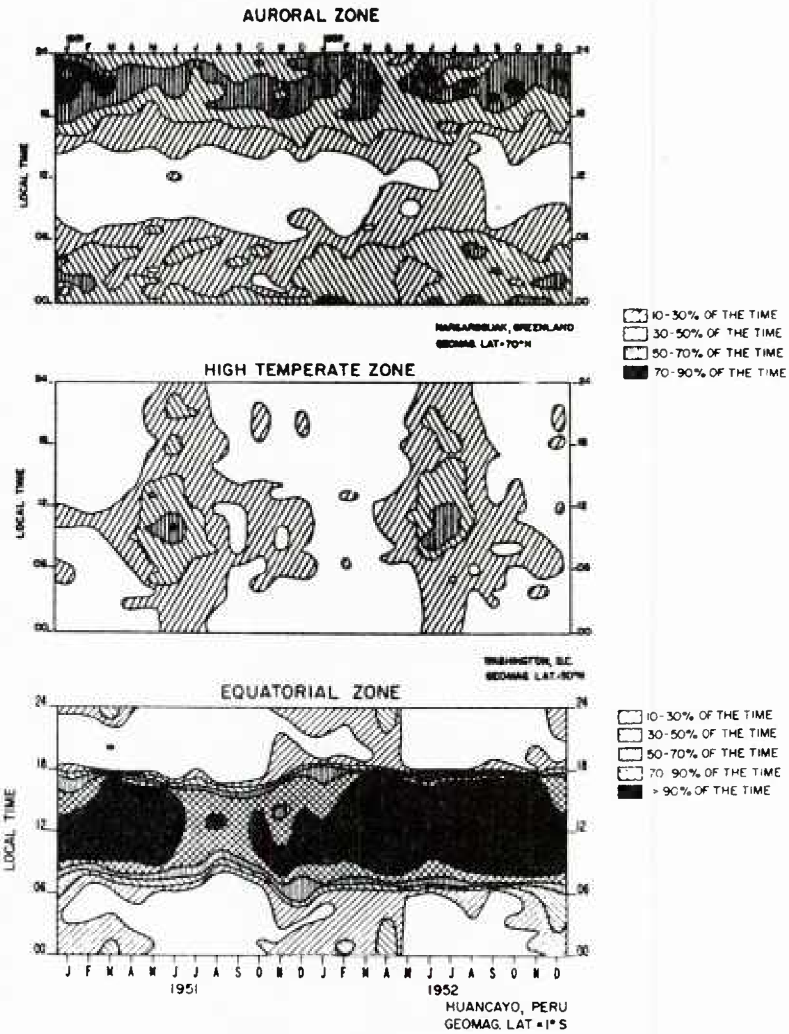



Figure 9 Fraction of time $fE_s$ > 5 MHz (Smith 1957).

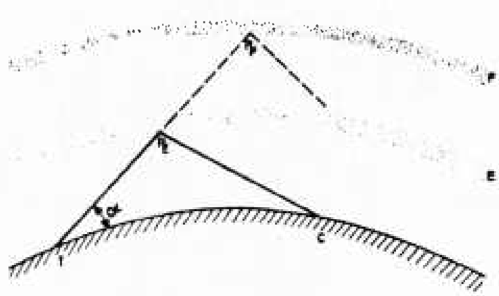



Figure 10 Illustrating the screening effect of a sporadic E-layer.

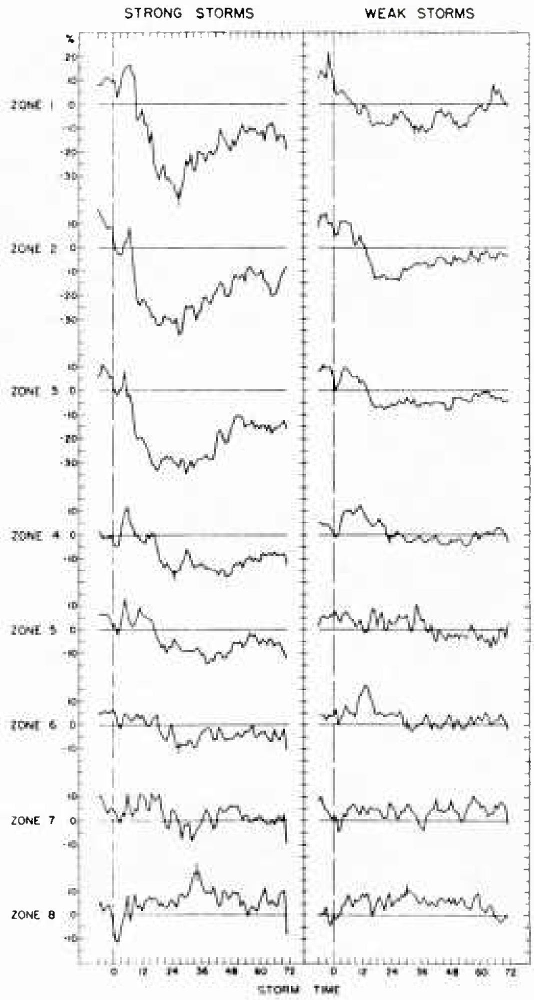



Figure 11 Variations of maximum electron density in the F-region $N_{em}F2$ during magnetic storms. The ordinate is the approximate percentage deviation from the quiet day behaviour versus storm time. The zone number is in the following shown in parenthesis between the applicable geomagnetic latitudes 60°(1) 55°(2) 50°(3) 45°(4) 40°(5) 30°(6) 20°(7) 10°(8) -10° (Matsushita 1959).

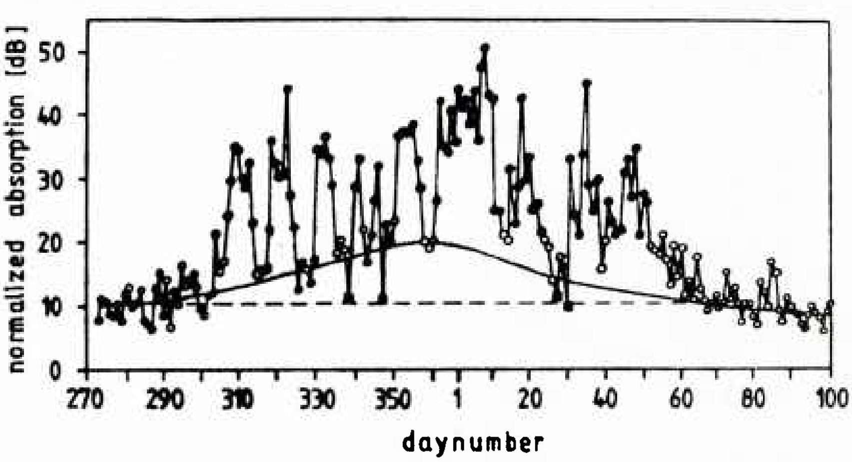



Figure 12 Typical values of HF absorption measured at constant solar zenith angle during winter. Full squares show "summerlike" days (Schwentek 1971).

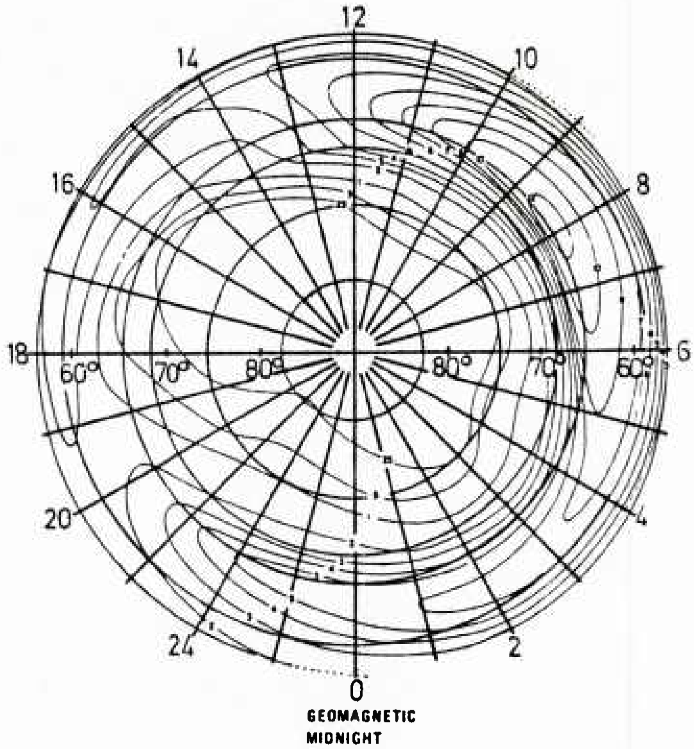



Figure 13 Percentage of the time that auroral radio wave absorption of 1.0 dB or more occurred at 30 MHz. The data are plotted as a function of geomagnetic latitude and mean geomagnetic time (Harz et al 1963).

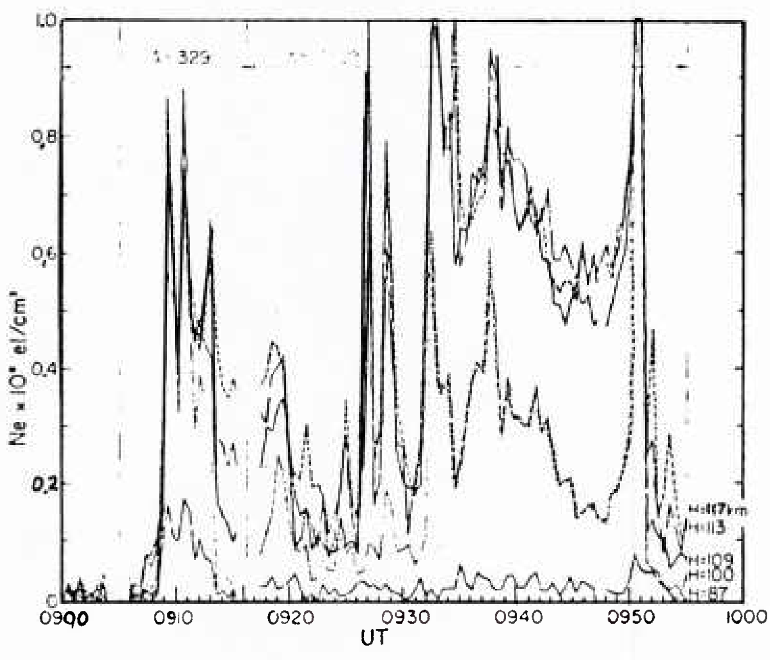



Figure 14 E-region electron density time variations measured during an auroral event of 2 April 1973 with the Chatanika radar (Hunsucker 1975).

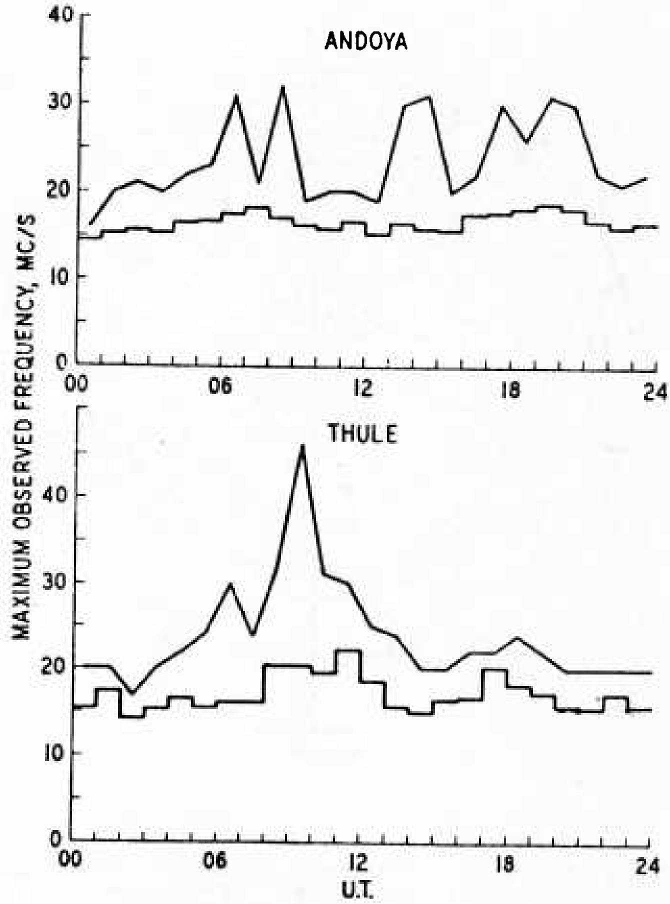



Figure 15 Sporadic E maximum observed freauencies (MOF) for paths from College, Alaska to Thule and Andøya for the period from Nov 27, 1963 to Febr 12, 1964. The histograms give the hourly averages and the upper curves denote the highest MOF for the hour (Hunsucker and Bates 1969).

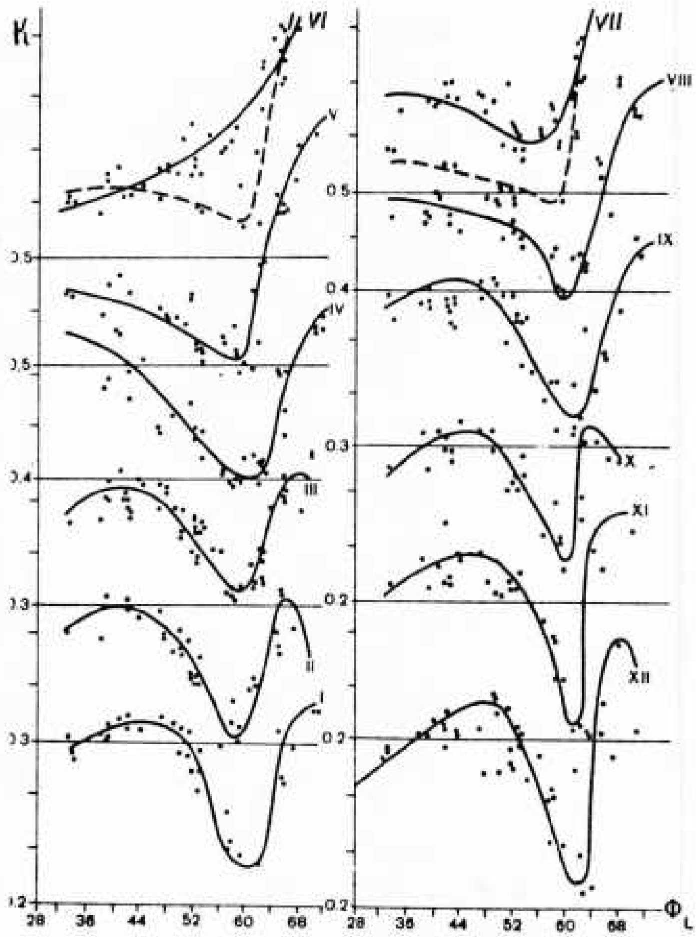



Figure 16 Latitude dependence of the normalized-to-noon values of the critical frequency for the F2-layer at 02 Local time for the months of 1964 (Besprozvannaya et al 1979).

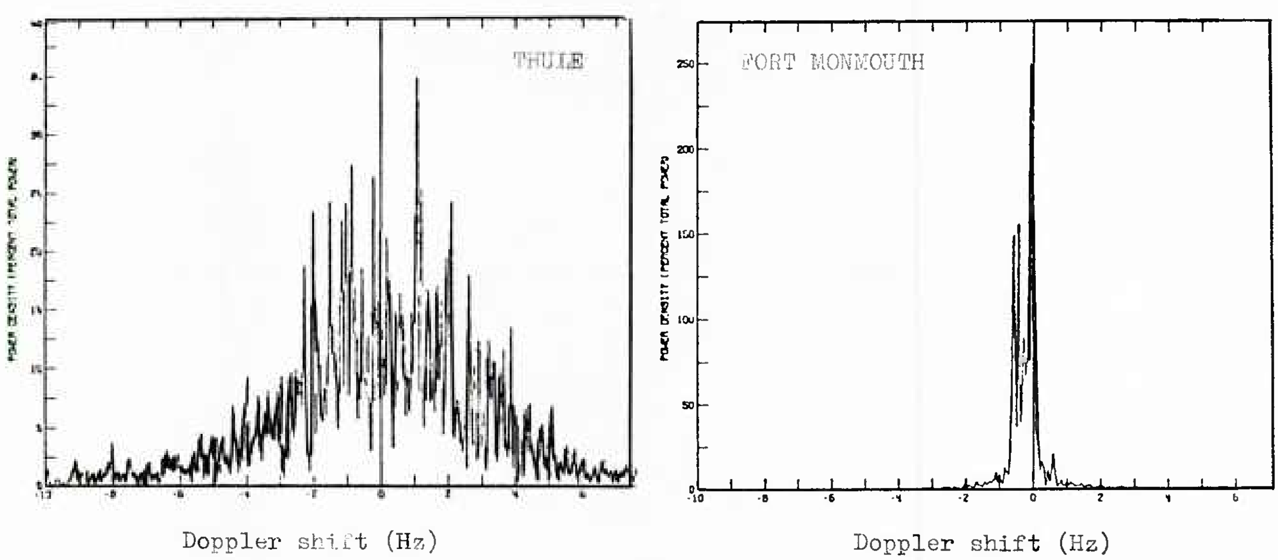
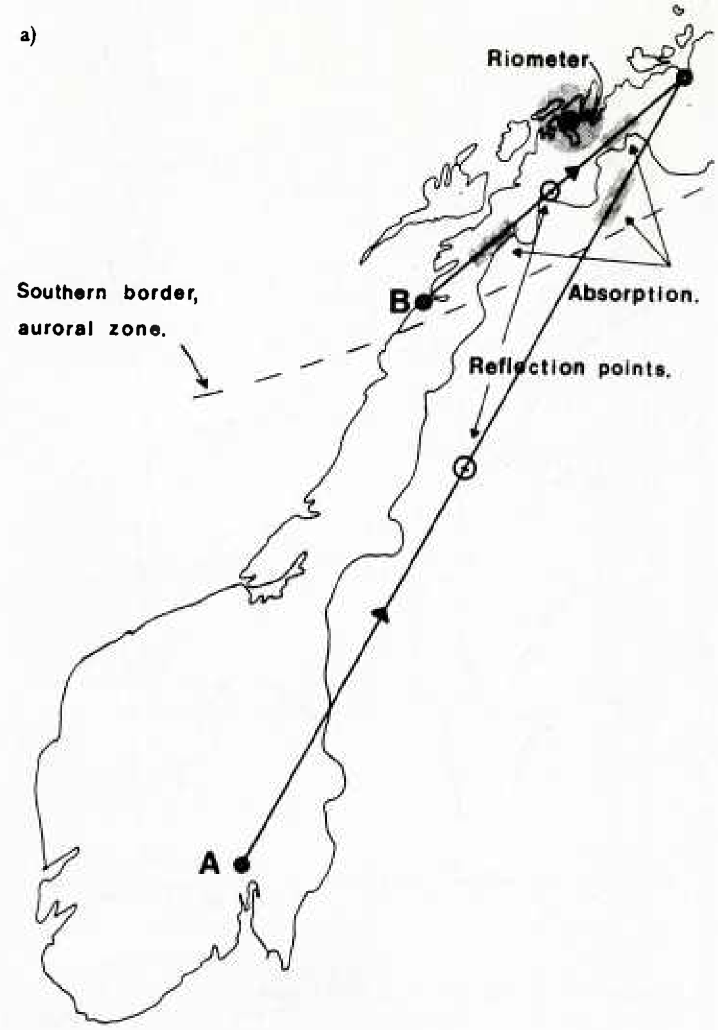



Figure 17 Typical power spectra of transmissions from Thule (high latitude) and from Fort Monmouth (mid-latitude) to Palo Alto (Hunsucker and Bates 1969).



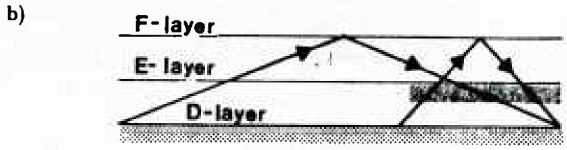




Figure 18 a) The two transmission paths used in the tests.

b) Schematic representation of the wave paths through the disturbed auroral D-region (Thrane 1979).

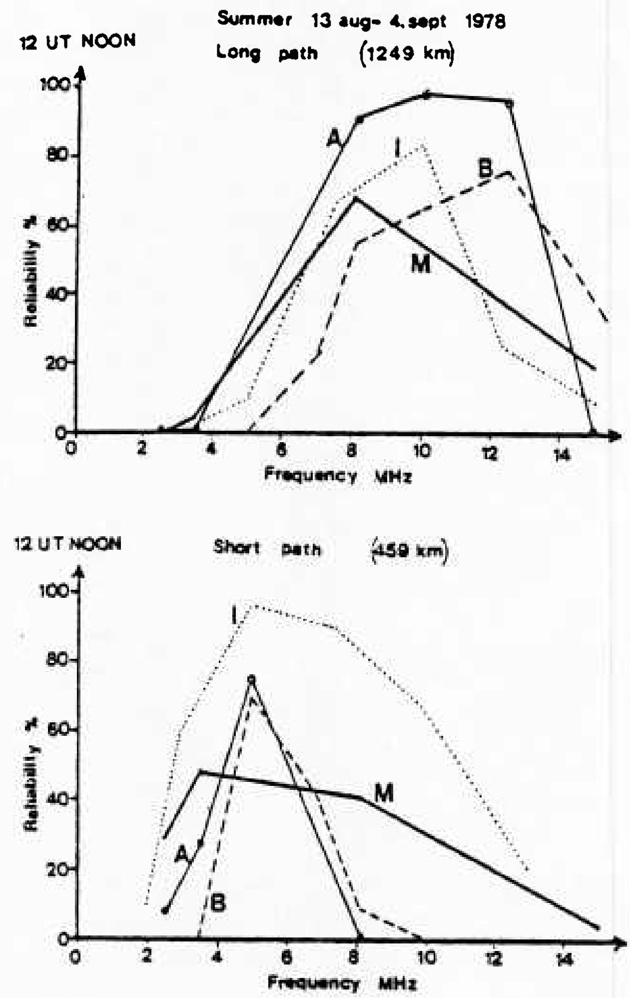



Figure 19 Measured and predicted reliabilities for the two paths for summer noon

Code: A Applab III
B Bluedeck
I IONCAP
M Measurements

(Thrane and Bradley 1981).

# MEASUREMENTS OF HF PROPAGATION PARAMETERS FOR REAL-TIME CHANNEL EVALUATION (RTCE) SYSTEMS

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## Summary

HF propagation paths are time-spread and frequency-spread channels, and are characterized by severe variability in the time domain of all their properties, inclusive of path losses. In addition, even when path conditions would be affordable, the link may be severely interfered with by other transmitters. Improvements over present-day link's performance in terms of circuit reliability, data rate and error rate can be achieved through the use of adaptive schemes capable of coping with the variability of the channel. Recent advances in microprocessor technology, in frequency-agile HF equipment and in the understanding of the propagation medium, make it feasible and practical to use adaptive schemes at HF. The simplest approach is to make the link adaptive to a selected channel parameter such as the Signal-to-Noise ratio. At the other extreme, the adaptive links can be designed to respond to all relevant channel functions, such as Signal-to-Noise ratio, multipath spread, Doppler spread, etc. In all cases, a necessary prerequisite for link's adaptivity is Real-Time Channel Evaluation (RTCE). The RTCE data gathering must match in scope and complexity the adaptivity scheme that is meant to serve. For instance, if the link's data rate is adjusted only to the Signal-to-Noise ratio, the RTCE must be kept very simple and must be reduced to the sole measurement of the signal intensity and to the level of the noise, inclusive of interference. Measurement of pertinent link's parameters by the RTCE must be performed at several spot frequencies in the band of interest, in order to identify automatically the most suitable carriers for the transmission of the information. RTCE is at present in the R&D phase, and the related activities consist almost solely of the measurement of signal amplitude and noise levels, inclusive of interference, at several frequencies in specific HF paths of interest, with only a few instances of inclusion of multipath spread measurements. At the successful completion of the R&D activity, it is expected that in a form or another, RTCE and the related adaptive schemes, will enter the practice of modernized HF communications.

## A. INTRODUCTION, RATIONALE FOR RTCE MEASUREMENTS, AND PRINCIPLES OF SOUNDING/PROBING

### 1. General

HF propagation paths are time-spread and frequency-spread channels, and are characterized by severe variability in the time domain of all their properties, inclusive of path losses. In addition, even when path conditions would be acceptable, the link might be severely interfered with by other transmitters.

Improvements over present-day link's performance in terms of circuit reliability, data rate and error rate can be achieved through the use of adaptive schemes capable of coping with the channel variability.

Real-time, oblique ionospheric sounding between two terminals of the link, noise and interference measurements at the receiving end, and channel probing simultaneously performed between the two link's terminals are the data gathering operations that provide the inputs on which to base the adaptive control of the link's performance parameters.

The master station of the link, where the sounding transmitter is located, could perform a sounding scan similar to the one made by an ionospheric oblique sounder of the normal practice, and could also generate the waveform for channel probing, while still operating as a complete terminal for two-way digital communications. During the pauses of the emissions, measurements of noise and interference could be performed at both the master and the slave station of the link, for use by decision-making microprocessors and control units. At each terminal, the transmitting and the receiving facility could have

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separate units for sounding/probing and for communicating, or these functions could be performed by the same equipment in different modes of operation. In the latter case, the equipment at the two terminals could be identical and the assignment of the master and of the slave roles would be dictated solely by operational requirements.

By processing the data obtained by sounding and probing, it would be possible to select automatically the group of frequencies to be used for communciating. At each sounding cycle, information about the frequency selection and about the waveform to be employed is exchanged between terminals and is used locally to achieve adaptivity. During the next sounding scan, the group of frequencies that were selected for communicating could be excluded from the sounding frequency plan. Instead, information on the changing status of the group of communicating frequencies could be obtained from measurements performed on the coded waveform that is part of the communications bit stream [Gupta and Grossi, 1980, (1); 1981, (2)].

The properties of the path that must be monitored in real time are:

1. Noise and interference spectral density
2. (Signal + Noise) level
3. Multipath spread
4. Doppler spread

It would be useful to know accurately the time variability, and related statistics, of these properties. Unfortunately, this information is available only for particular cases, and a reliable experimental investigation on the properties above is thus far an unfulfilled requirement. In general, we can say that these ionospheric channels exhibit time fadings that are important in determining the design of the signal and, in addition, show long-term variations due to large-scale fluctuations of the medium. Such slow effects have a time constant significantly greater than 5 to 10 minutes, an interval of time that appears appropriate as the basic sounding/probing periodicity. Adaptive approaches to the communications problem are required to circumvent this long-term variability in propagation conditions.

In this lecture, we will make the usual distinction between path sounding and channel probing, with the former devoted to the measurement of path losses and of noise and interference levels and with the latter devoted to the measurement of such parameters as multipath spread and Doppler spread. The following criteria were adopted:

1. The link is assumed reciprocal, except for the noise and the interference levels at each terminal. Therefore, the decision on the frequencies to be used (this decision is based on the results of the sounding operation) is based on the measurement of interference and noise both at the master and at the slave station, and on the one-way measurement of the path losses between the two;

2. Processing of the multipath spread and of the Doppler spread data (provided by the channel probing operation) is performed at the slave station, and the results are transmitted back to the master station, for use in the final selection of the frequencies to be used in communicating;

3. Channel probing is to be undertaken only at the best frequencies put in evidence by the path sounding, in order to shorten the overall operation sounding/probing.

It was assumed by Gupta and Grossi (1 and 2) that a very large number of spot frequencies was available to the adaptive link: 1125 or 3375 carriers, respectively for a mid-latitude and for a transauroral path, in the band 3 to 30 MHz. These authors advocated a tightly integrated sounding/probing/communications scheme, in which a set of frequencies was simultaneously transmitted to achieve waveform diversity. The power density (Watt/Hz) at each frequency was low enough to be received below noise by standard HF receivers, as long as they were located outside a circle with about 100 km radius, centered at the link's transmitting terminal. Only those receivers that were coherently processing the waveform containing the information to be exchanged between the two terminals were able to detect the waveform above noise. The sounding scan proposed by these authors was lasting 100 to 160 seconds, and was repeated every 300 to 480 seconds.

More recently, Aarons and Grossi [1982, (3)], have proposed an approach that reduces substantially the number of spot frequencies at which the sounding is performed. They pointed out that achieving complete adaptivity to the path is impractical: too many spot frequencies are needed. Although, in principle, a frequency-spread waveform is the way to go, there are practical limits to this spread. If adaptivity, then, cannot be extended beyond certain limits, we cannot dispose of the continuing need for ionospheric forecasting on the path, and of the need of warnings about magnetic storms and solar proton activity. A hybrid approach, where the traditional forecasting and warning functions are kept intact, and where the system is made partially adaptive (within practically acceptable limits), represents an advisable approach toward improvement and modernization of HF technology and systems.

Aarons and Grossi advocated that forecasting should continue to coexist with adaptivity, in order to provide a first-cut identification of the frequency windows that are available for use in the channel, as a function of geographic and geomagnetic coordinates, time of the day, sunspot number, etc. Forecasting should actually be extended in scope, to include the prediction of the time-spread and of the frequency spread of the path. Concerning the ionospheric warning function, we should expect that the use of solar-magnetic sensors will also be continued. These sensors make it possible to have warning lead times that range from a few minutes to several hours, sufficient therefore to adjust system operation to the forthcoming conditions of the path. Early knowledge of PCA events, of magnetic stroms and of similar phenomena, will be a prerequisite for the effective performance of an adaptive system. Choice and amount of path diversity, amount of data rate, extent of use of error-correction schemes, etc., are all functions that can be optimized by the simultaneous use of forecast/warning and partial adaptivity.

As in any adaptive approach that involves real time data gathering to adjust the parameters of an electromagnetic system to the propagation medium, the preferable way is to adapt, first of all, the system to the best a-priori model of the medium, in our case, to the model of the HF ionospheric path. Generally, this model is a software subroutine stored in the memory of the microprocessor that is used in the logic units of the system. This model is periodically modified and adjusted to reflect the instructions of the forecasting and warning operational functions. It is then updated and brought in close agreement with the actual path conditions by means of real time observations performed by the communications link itself. One approach of this kind has been already successfully implemented in correcting for ionospheric-induced errors in high-performance radars [Katz et al., 1978, (4)]. Figure 1 [taken from Aarons and Grossi, 1982, (3)] gives the block diagram of principle of this HF communications approach. The combination of mean-model plus real-time updating should be taken in serious consideration, when designing next-generation HF communications links.

In order to help visualizing real-time sounding and probing, we illustrate in the following Sections (2 and 3) the scheme that was conceived by Gupta and Grossi (1 and 2).

2. Path Sounding

Path sounding has the scope of measuring path losses at an adequate number of spot frequencies in the band of interest (3 MHz to 30 MHz) and of measuring at the same time noise and interference levels, at the same frequencies and at both ends of the link.

Table I gives the parameters of the sounding scan proposed by Gupta and Grossi (1). The master station radiates sequentially 1125 to 3750 carriers to cover the 3 to 30 MHz band, in a time interval 100 to 160 seconds long (88 to 47 milliseconds per carrier). Of the two values given above for each sounding parameter, the first applies to a mid-latitude path, the second to a transauroral path. The scan is repeated every 5 to 8 minutes.

Once that a set of frequencies has been chosen for communicating, it is automatically excluded from next sounding cycle. However, information on the channel status for each one of the frequencies excluded from sounding and probing is still updated once every 5 to 8 minutes by measurements performed on the communications waveform. Frequency switching is preceded by a "tone" of notification and takes place even while communications go one, for the case in which the channel deteriorates and another set of frequencies is found more suitable to carry out communications. The block diagram in Figure 2 pertains a two way adaptive link and illustrates the various functions of the two terminals. One-way sounding and probing is achieved in this case from the master station to the slave station. Acknowledgement is from the slave to the master station. Communications are two-way exchanges between the two stations.

Table I - Sounding Scan Parameters

| | Mid-latitude Path | Transauroral Path |
|---|---|---|
| Band covered | 3 MHz to 30 MHz | 3 MHz to 30 MHz |
| Number of spot frequencies | 1125 | 3375 |
| Separation between two adjacent spot frequencies | 24 KHz | 8 KHz* |
| Sounding scan time | 100 seconds | 160 seconds |
| Rate of sounding scan repetition | one every 300 sec | one every 480 sec |
| Dwelling time per spot frequency | 88 milliseconds | 47 milliseconds |
| Nominal bandwidth of sounding receiver | 24 KHz | 8 KHz |
| Width of sounding pulse | 41.5 microseconds | 125 microseconds |
| Pulse repetition frequency | 100 pps | 100 pps |
| Pulse per dwelling time | 8 pulses | 4 pulses |
| Noise and interference measurement's integration time, for each spot frequency | 53 milliseconds | 47 milliseconds |
| Overall noise and interference measurement time | 60 seconds | 160 seconds |

[From Gupta and Grossi, 1980, (1)]

The sequence of steps through which sounding and probing are performed is as follows:

For the midlatitude link

Step 1 - 100 seconds devoted to sounding operation.
Step 2 - 60 seconds devoted to measurement of noise and interference at both terminals of the link.
Step 3 - 20 seconds devoted to computations, taking into account the need of accumulating at a single terminal (the slave station) the information pertaining noise and interference at both terminals. During this step, the microprocessor at the slave station selects the frequencies and designates them to the master station.
Step 4 - 100 seconds devoted to channel probing, to be performed only at the frequencies designated by Step 3.
Step 5 - 20 seconds devoted to computations, acknowledgement and information exchange between the two terminals, in order to perform the final selection of frequencies to be used in communications, by taking into account the data on multipath spread and Doppler spread.
Step 6 - The two terminals are now ready to initiate communications. The frequencies finally adopted for communications are excluded from next sounding/ probing cycle (one every 300 seconds), although they continue to be monitored by measurements on the modulation waveform.

For the transauroral link

Step 1.- 160 seconds
Step 2 - 160 seconds
Step 3 - 20 seconds
Step 4 - 120 seconds
Step 5 - 20 seconds
Step 6 - The two terminals are now ready to initiate communications. The sounding/probing cycle is repeated every 8 minutes (480 seconds).

*This value is chosen because 8 KHz is the bandwidth of the signal waveform selected for the transauroral link. The path coherent bandwidth is only 660 Hz.

Communications are therefore inhibited only in the first 300 (or 480) seconds of link operation. After this initial adaptive adjustments of the link's terminals, any readjustment is performed without requiring a discontinuation of communications.

3. Channel Probing

The importance of time-dispersive and frequency-dispersive effects in HF propagation has been amply treated in the literature. These effects are determining factors in the conceptual design of an adaptive system. Channel probing is aimed at gathering information on these effects, after the path sounding has determined path losses and noise (plus interference) levels at the available spectral lines, and has identified the frequencies promising enough to be worthy of the channel probing effort. All these functions are slowly varying functions, so that one sample every 5 to 8 minutes is adequate. The measurement of multipath spread and Doppler spread can be achieved with a variety of methods, either based on the direct measurement of these two quantities or on indirect measurements such as the ones based on the fact that, at a given frequency, the reciprocal of the Doppler spread gives the fading period of the arriving e.m. wave, or that the reciprocal of the multipath spread, at a given instant of time, gives the frequency interval within which carriers fade coherently. Because the amount of time required to process the information on the dispersive properties of each channel is not trivial, these measurements should be performed only for those frequencies for which path sounding has indicated acceptable path losses and affordable noise and interference levels. Therefore channel probing has to follow, in time, the sounding operation.

Ideally, channel probing should provide a reliable estimate of all the parameters of the path that are indicated in Figure 3, 4, and 5. These are the quantities $B_o$, $B_d$, $B_m$ and the function $S(\xi, \nu)$, called the path's scattering function. In actual practice, it is sufficient to simplify the scattering function to a group of N gaussoids, with $B_o = 0$, $B_d = 0$, and to reduce therefore the function to the one shown in cross-section in Figure 6 and 7. An intuitive picture of the scattering function can be obtained as follows. Suppose that the path is such that a pulse of infinitesimal length is transmitted unaffected and that a spectral line of great purity, also, is not broadened. We can say in this case that the scattering function of the path is

$$S(\xi, \nu) = \delta(\xi)\ \delta(\nu)$$

In other words, the scattering function is the product of two delta functions. This is a highly idealized case. In real propagation, an infinitesimally short pulse, and a spectral line of infinitesimal width, are actually broadened, respectively in the time and in the frequency domain. This causes the scattering function to have a finite width both along the frequency axis and the time axis. See Green [1968, (5)], to visualize the relationship between scattering function and other well known channel functions, such as impulse response, etc.

Going back to Figures 6 and 7, the analytical expression for the scattering function becomes:

$$S(\xi, \nu) = \sum_{i=1}^{N} P_i \ (2\ \pi\ B\ L_i)^{-1} \exp\left\{-\frac{1}{2}\left[\frac{(\xi - \xi_i)^2}{L_i^2} + \left(\frac{\nu}{B}\right)^2\right]\right\}$$

In this formula, the parameter N represents the number of paths in the structure, $\xi_i$ and $L_i$ are the mean delay and the multipath spread, B is the Doppler spread of the path, and $P_i$ represents the relative strength of the $i^{th}$ path. Further simplifications can be achieved by representing the scattering function as a single gaussoid, whose amplitude is a function of the path losses and whose width $L_{tot}$ and $B_{tot}$ (Figure 5) are respectively the total time spread and the total Doppler spread.

The measurement of the properties of a communication channel is particularly important in digital communications, because high-speed data transmission critically depends upon them. Kailath [1959, (6)] pointed out that the problem of the measurement of the system functions in random, time-variant, channels might be unsolvable. This author introduced a parameter called the "spread factor" as the measurability criterion. If the product of the maximum Doppler spread and of the maximum multipath spread is larger than unity, and if no other information is available on the channel, the channel parameters cannot be measured accurately. Fortunately, for ionospheric HF paths, the spread factor is less than unity, so that standard measurement techniques, as the ones described here below, are applicable. Bello and Esposito [1970, (7)] have analyzed these measurement techniques for random, time-variant, dispersive channels, and they list them in three levels of increasing complexity:

1. Measurement of multipath spread and Doppler spread; of Doppler shift and spectral skewness;
2. Measurement of second-order channel functions;
3. Measurement of instantaneous channel functions.

For the parameters in item 1 above, measurement techniques used are based upon differentiation, level crossing and correlation. For item 2 above, the techniques used are correlation techniques, multitone approach, pulse pair method and chirp technique. For the measurement of the parameters in item 3, the methods used are the cross correlation, the multitone approach and the pulse pair technique. Gupta and Grossi (1) give an account of the methods usable in probing the channel and illustrate possible hardware implementations (see Figure 8). They review the measurement of Doppler spectrum parameters, the simultaneous measurement of Doppler spread and multipath spread, the determination of the instantaneous impulse response, as well as the measurement of time-variant channel functions and of channel correlation functions. In this lecture, we concentrate on the methods for measuring the multipath spread and the Doppler spread, that are the channel parameters characterized by the highest priority.

## B. TECHNIQUES FOR REAL-TIME CHANNEL EVALUATION (RTCE)

### 1. Introductory Remarks

Measurement of Signal intensity and of noise levels (inclusive of interference), together with the determination of multipath spread and of Doppler spread are the basic objectives of RTCE functions. These quantities are fundamental prerequisites to achieve link's adaptivity. Of the three parameters, techniques for the measurement of signal and noise are well known and already part of HF communications practice. We will briefly review them in the summary at the end of this Section. Our attention will focus on the more difficult task, and on the less known related approaches, concerning the measurement of multipath spread and Doppler spread. We have already indicated, earlier in this lecture, that instead of measuring the complete scattering function, it is sufficient, and obviously simpler, to measure the delay power spectrum $Q(\xi)$ and the Doppler power density spectrum $P(\nu)$ [Bello, 1963, (8); 1965, (9)].

### 2. Measurement of Multipath Spread and of Doppler Spread

The measurement of gross parameters of the channel's second order functions, such as multipath spread, Doppler spread, coherence time and coherence bandwidth are sufficient information for the preliminary evaluation of a given channel. Let's first of all define these gross parameters [Bello and Esposito, 1970, (7)].

$$D = \text{Doppler Spread} = 2\sqrt{\frac{\int (f-\bar{f})^2\, P(f)\, df}{\int P(f)\, df}}$$

where: $\bar{f} = \dfrac{\int f\, P(f)\, df}{\int P(f)\, df}$ is mean Doppler Shift (Centroid of Power Sprectrum)

$$M = \text{Multipath Spread} = 2\sqrt{\frac{\int (\xi - \bar{\xi})^2\, Q(\xi)\, d\xi}{\int Q\,(\xi)\, d\xi}}$$

where: $\bar{\xi} = \dfrac{\int \xi\, Q(\xi)\, d\xi}{\int Q(\xi)\, d\xi}$ is mean path delay (Centroid of delay power spectrum)

The measurement approaches that are available for the evaluation of the gross parameters of the channel's second order functions are as follows:

Differential approach

For the Doppler Spread, there is a simplified formula, usable for Gaussian fading channels:

$$D = \frac{1}{\pi\,\alpha}\sqrt{\frac{<[\dot{e}(t)]^2>}{<[e(t)]^2>}}$$

where

$$\alpha = \frac{1}{\sqrt{2}}, \text{ for linear envelope detector}$$

$$\alpha = 1, \text{ square-law detector}$$

$$e(t) = k\ (x^2 + y^2)$$

For the Multipath Spread, there is again a simplified formula:

$$M = \frac{1}{\pi\,\alpha}\sqrt{\frac{\{(dE/df)^2\}}{\{E^2\ (f)\}}}$$

where

$$E\ (f) = k\left[X^2(f) + Y^2(f)\right]$$

Level Crossing Approach

For the Doppler Spread, we have:

a) if n is the number of times/sec that the envelope crosses its rms value, then

$$D = \frac{2\ e\ n}{\sqrt{\pi}}\ ;$$

b) if m is the number of times/sec that the in-phase and quadrature components cross the zero line, then

$$D = m\ .$$

For the multipath spread, we have:

a) if $\hat{n}$ is the average number of times/Hz that the envelope of the transfer function

$$T\ (f,\ t) = \int g\ (t - \xi)\ \exp\ [-j\ 2\ \pi\ f(\xi)\ d\xi]$$

(Fourier transform of the channel impulse response) crosses its rms value, then

$$M = \frac{2\ e\ \hat{n}}{\sqrt{\pi}}$$

b) if $\hat{m}$ is the number of times/Hz that the real (or the imaginary) part of the Channel's transfer function crosses the zero line, then

$$M = \hat{m}$$

Correlation Approach

For the Doppler Spread, the computation is based on the intensity of the received carrier, and on the autocorrelation function at one delay.

By adopting a parabolic approximation $\rho(\tau)$ for the correlation function $R(\tau)$ of the in-phase (or the quadrature) component of a received fading carrier, we have:

$$D = \frac{1}{\pi\ T}\sqrt{1 - \rho\ (T)}$$

where T is the chosen delay time. For the multipath spread, we can proceed similarly, and we have:

$$D = \frac{1}{\pi\ T}\sqrt{1 - \hat{\rho}\ (F)}$$

where $\hat{\rho}$ is the normalized frequency autocorrelation function of the in-phase (or the quadrature) component of the channel transfer function, and F is the selected frequency separation.

3. Measurement of the Channel's correlation functions, of the instantaneous impulse response and of the Channel's transfer function

Correlation Functions

The methods that are available are:

pulse pair probing techniques
correlation technique
chirp technique
multitone technique

Impulse Response

a) by pulsing probing: measurement of the continuous delay power spectrum
measurement of the discrete delay power spectrum
b) by cross-correlation: as above

Transfer Function

The method that is available if the multitone measurement of this function.

4. Summary and recapitulation of channel evaluation techniques

a) Measurement of Signal and Noise

This can be best achieved by using an "off-on" keyed signal, with the noise measurement preceding the signal measurement at any given frequency. Choosing for the noise measurement integration time the values indicated in Table I, the estimation of $\sigma_n$ can be performed with adequate accuracy. The signal level also must be measured by averaging a sequence of observations. Table I gives two examples of pulse widths, pulse repetition frequencies, and overall time required for both signal and noise determination. The measurement accuracy for signal level estimation is limited by the channel fluctuations, i.e.:

$$\frac{\mathrm{Var}\ (\hat{A})}{A^2} = 1/N_{eq}$$

where $N_{eq}$ is the number of independent observations available in one sounding scan time's (let's take the case of the midlatitude path in Table I) repetition period, which is 300 seconds. Because the fading rate is $1/B_{tot}$, we have $N_{eq} = 300\ B_{tot}$. If $B_{tot} \simeq 1$ Hz, we have that $N_{eq} = 300$. The noise level estimate has a ratio

$$\frac{\mathrm{Var}\ (\hat{\sigma}\ n)}{\sigma_n^2} = \frac{1}{TW} = 7.86\ 10^{-4}$$

again, by taking the numbers in the column of Table I devoted to mid-latitude path (53 millisec x 24 KHz).

b) Measurement of Doppler spectrum characteristics

The accuracy achievable in these measurements strictly depends on the dwelling time on each frequency at which channel probing is performed. If we want, for example, a Doppler spectrum resolution of 0.33 Hz, we require an observing time, per frequency, of 3 seconds. In the example that we gave immediately after Table I, we chose 100 seconds as the time devoted to channel probing (Step 4). By assuming that Doppler measurements are performed in a non-interference basis with the other measurements, only 33 carriers can be measured in the allotted time interval. We can see therefore that determination of Doppler spread is very demanding in terms of length of observations.

c) Measurement of delay power spectrum

Two types of signals can be considered:

1. Single pulse

$$p(t) = \begin{cases} 0 & , \text{ otherwise} \\ 1 & , 0 \leq t \leq \Delta \end{cases}$$

2. Coded sequence (phase reversal keying)

$$z_o(t) = \sum_{k=1}^{N} a_k \, p(t - k\Delta)$$

where

$$a_k = \pm 1$$

$$N = 2^m - 1$$

$$T = N\,\Delta = \text{total duration}$$

The coded sequence gives the same performance as a single pulse for equal energy; i.e., if $P_1\Delta = P_2T$ (neglecting degradation due to sidelobe structure of sequence autocorrelation function).

Concerning the measurement of $Q(\xi)$ in absence of noise, the basic limitation in accuracy achievable is the number of independent channel's "snapshots" (impulse responses) that can be obtained in one scanning cycle:

$$\frac{\text{standard dev. } [\hat{Q}\ (\xi)]}{\text{mean } \hat{Q}\ (\xi)} = \frac{1}{\sqrt{N_{eq}}}$$

In presence of additive noise, performance depends upon the shape of the true $Q\ (\xi)$ being measured. Let's consider an idealized function consisting of M equal rectangular modes:

$$Q(\xi) = \begin{cases} \dfrac{1}{ML_o} & ; \text{ during mode } (L_o \text{ is the duration of each mode}) \\ 0 & ; \text{ elsewhere} \end{cases}$$

Then, at the mode centers, $Q\ (\xi)$ can be estimated with the formula:

$$\frac{\text{standard dev. } [\hat{Q}\ (\xi)]}{\text{mean } [\hat{Q}(\xi)]} = 1 + \frac{\sigma_n}{E\ Q(\xi)}$$

where:

$\sigma_n$ = noise variance in the receiver bandwidth

E = equivalent pulse energy

A single pulse will not give adequate performance. When a sequence of length $2^{15} - 1$ is used, estimation of $Q(\xi)$ will be limited by channel fluctuations rather than noise. Because of the large number of frequencies being probed, the overall (noise-limited) performance is degraded.

## C. CONCLUSIONS

Adaptive HF approaches can be kept limited in scope and made to respond only to the variations of a selected channel parameter, such as Signal-to-Noise ratio. At the other extreme, they can be designed as all encompassing and capable to respond to all relevant channel functions, such as SNR, multipath spread, Doppler spread, etc. We have seen in this lecture pertinent examples. The RTCE data gathering must parallel in scope the adaptivity scheme that it is meant to serve. For instance, if data rate is adjusted only to SNR, the RTCE must be kept very simple and must be reduced to the sole measurement of the signal intensity and to the level of the noise (inclusive of interference).

In this lecture we have given examples of adaptivity schemes, and of related RTCE approaches, that tended toward the complicated side (see, for instance, the scheme depicted in Figure 2). This was done because learning of a complex solution makes it easier to visualize the simpler ones. There is little question, in fact, that the simpler solutions will enter the HF communications practice first, and that several years will pass before we see in operation a system such as the one depicted in Figure 2. Even the R&D activity presently underway on adaptivity schemes and related RTCE techniques, is almost exclusively limited to the investigation of link's adaptivity to SNR. However, there is little question that the demand for high data rates, and the quest for better HF link's performance (to bring this channel up to the quality of competitive approaches) will provide enough motivation to implement in practice present day designs of adaptive links, that will necessarily include RTCE features. All these improvements will enter the practice gradually, with adaptivity to multipath spread and Doppler spread coming last. When they will be in place, the era of truly modernized HF communications will have come.

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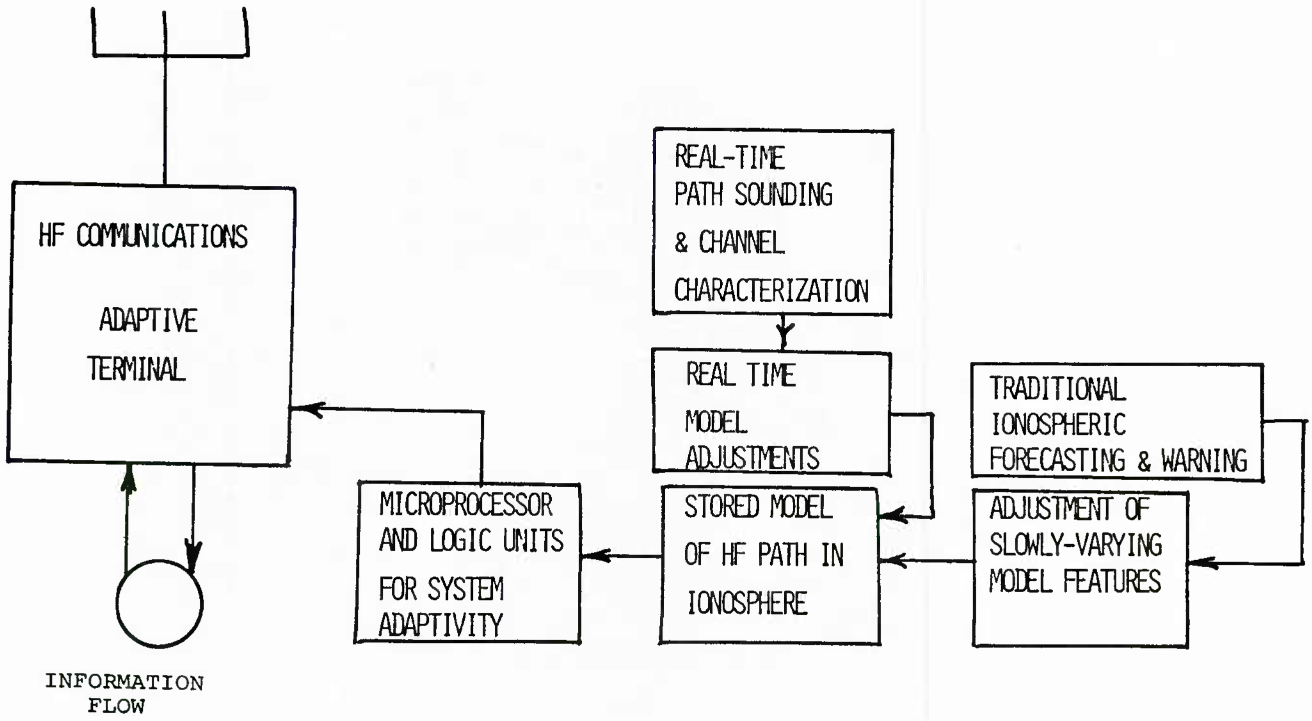


Figure 1 - Block Diagram of an Adaptivity Approach that Uses Traditional HF Ionospheric Forecasting and Warning, in Addition to Real-Time Path Sounding and Channel Probing ( taken from Aarons and Grossi, 1982, [3])

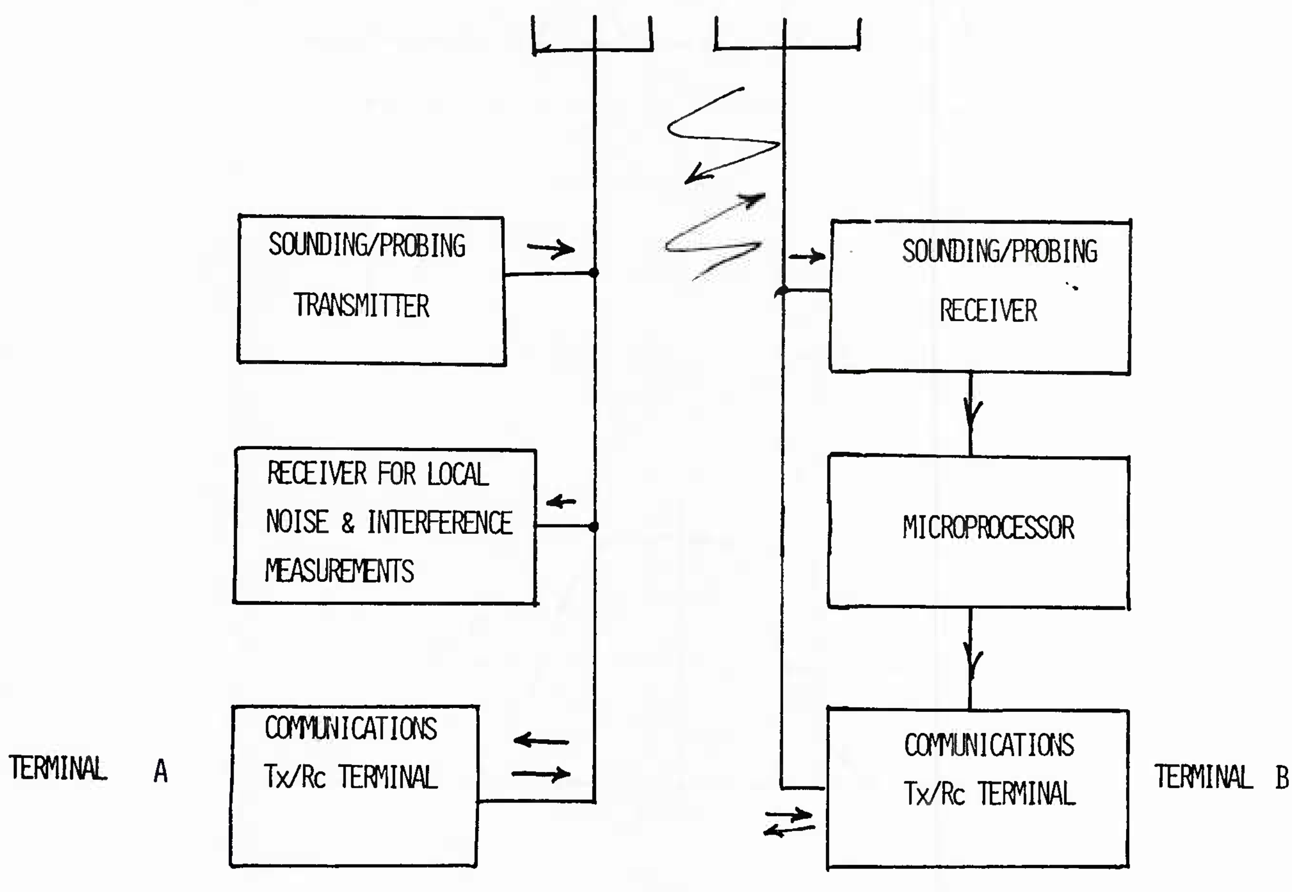


Figure 2 - Simplified Block Diagram of Advanced Adaptive Link with RTCE

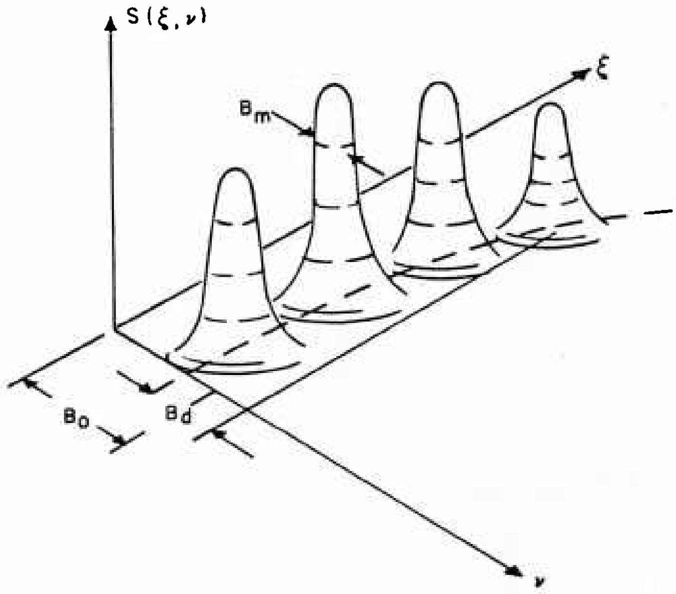


Figure 3 - Multimodal Scattering Function S $(\xi, \nu)$

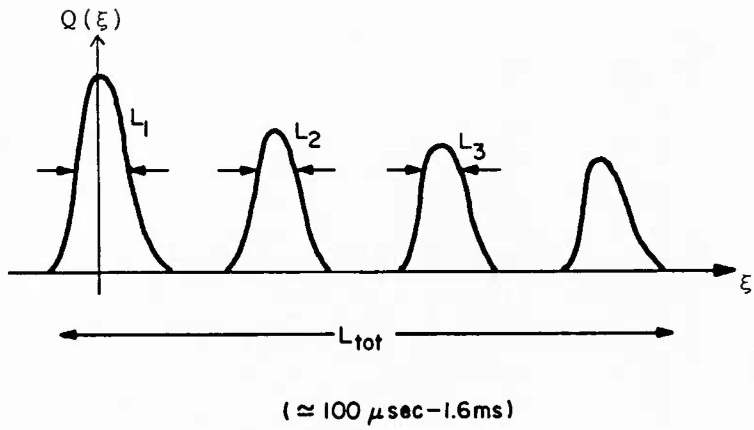


Figure 4 - Multimodal Delay Power Spectrum Q $(\xi)$

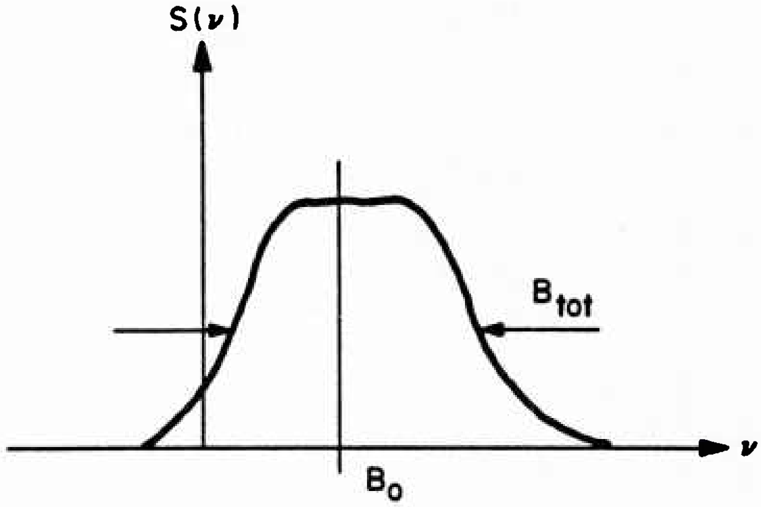


Figure 5 - Single Mode Doppler Power Spectrum S $(\nu)$.

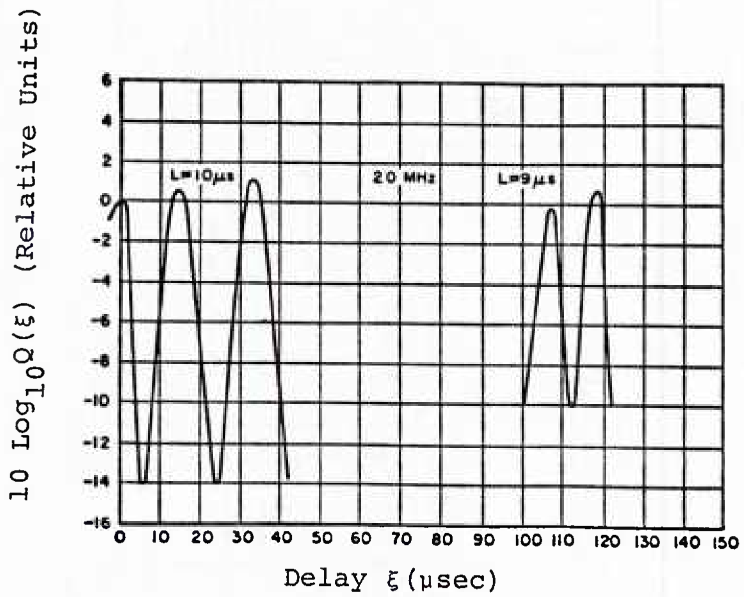


Figure 6 - Multipath Structure in a One-Hop Ionospheric Path in Mid-Latitude

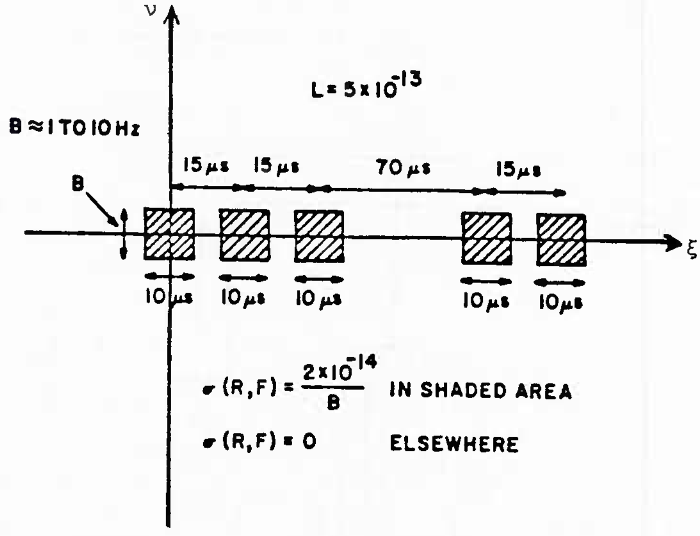


Figure 7 - Cross Section in the plane $(\xi, \nu)$ of the Scattering Function of the Path of Figure 6.

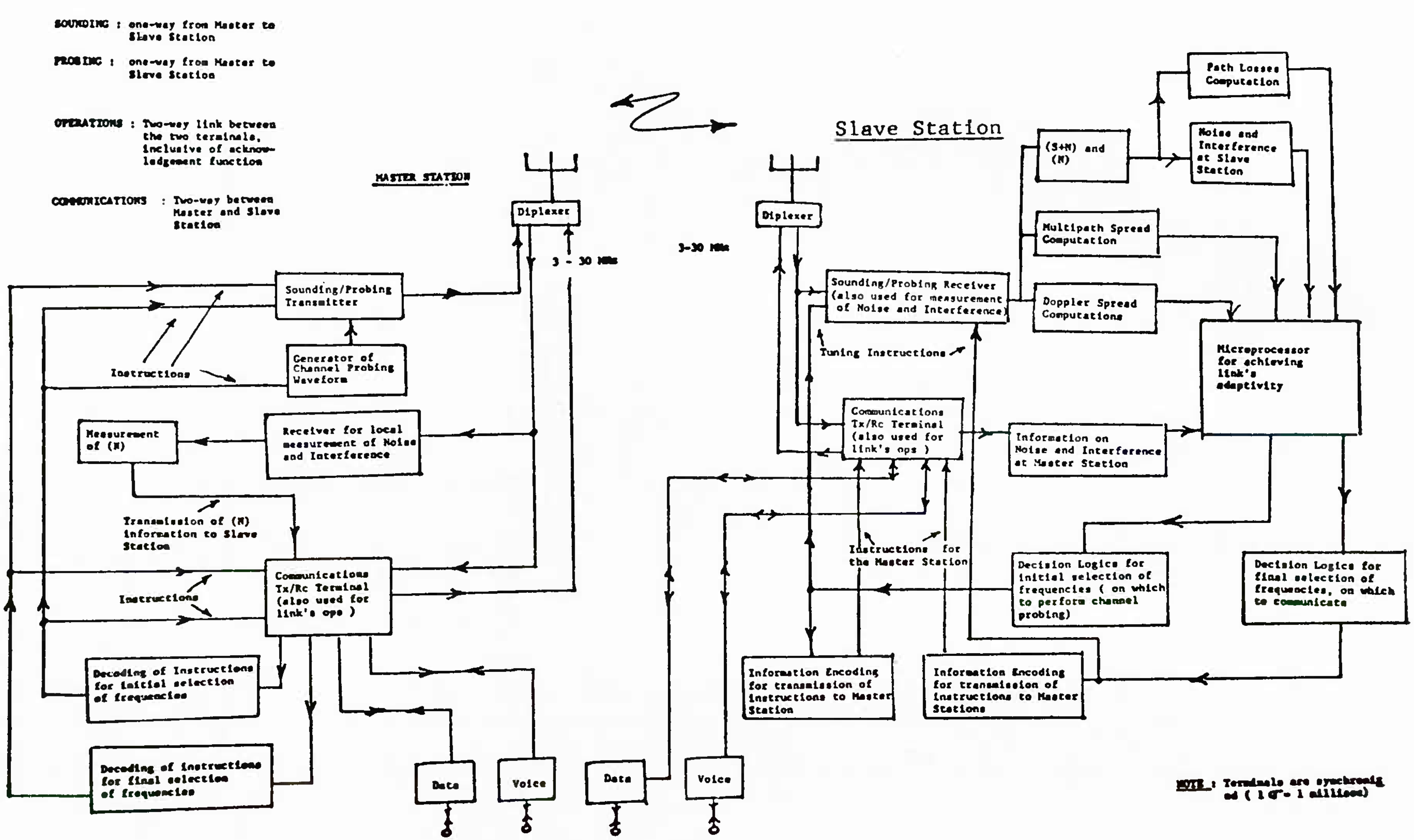


Figure 8 - Block Diagram of the Two Terminals of an Adaptive Link at HF
( taken from Gupta and Grossi, 1980, [1] )

# ADAPTIVE SYSTEMS IN OPERATION

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## ABSTRACT

The development and evolution of channel evaluation techniques is described. A recently developed fully automatic HF radio telephone system is discussed which automatically selects the suitable channel and also provides a telephone inter-connects. Also described is a HF message terminal which automatically requests, repeats and confirms message status for sender and receiver.

## 1. INTRODUCTION

In the early 1960's, real time channel evaluation (RTCE) systems were first used to improve the performance of operational HF radio systems (Jull et al,1962; Stevens, 1968). These systems were rather rudimentary as far as todays technology is concerned. The channel evaluation was done by equipment separate and independent of the communication transmitter and receiver. The RTCE equipment switched to each assigned frequency channel, measured the signal-to-noise or interference ratio and recorded the usable channels on a chart recorder. The radio operator selected the channel for communications by examining a history of the performance of all channels over a specified period of time. Since the 1960's, the RTCE systems have become more sophisticated and an integral part of the communication transmitter and receiver equipment. Microprocessors are now used to control the equipment operation and measure the parameters necessary for an adaptive system to operate effectively. With the microprocessor, additional features can be added, such as a calling system in which a base station can call a specific mobile terminal or a number of mobile terminals automatically. Various types of RTCE systems with different degrees of complexity are operational today. A recently developed HF radiotelephone system with automatic channel evaluation features is described in Section 2 and an HF message terminal that can be added to any type of system is described in Section 3.

## 2. RACE

A system named race (Radio Telephone with Automatic Channel Evaluation) was developed to improve the quality of telephone services provided by HF radio to remote areas (Chow and McLarnon 1982). This system not only evaluates the performance of each channel but also eliminates the requirement for an telephone operator. The race system consists of one Master and a number of Remote terminals. Each of these terminals consist of the following three subsystems as shown in Fig.1.

a) Controller Interface Unit (CIU):
The controller Interface is essentially a microprocessor which provides
   - an interface to the telephone system;
   - transmission and reception of dialled digits and supervisory data over the HF network;
   - channel evaluation and selection;
   - control of the HF transmitter receiver system.

b) HF Transmitter Receiver System:
This system consists of a conventional single side band transceiver with capability of a channel being selected remotely by means of a digital signal from the CIU. Broadband antennas are used to facilitate rapid switching of the channels.

c) Syncompex Unit:
This unit is a speech processor using digital techniques (Chow and McLarnon 1981) which improves the performance of the channel when the signal-to-noise or interference is low and provides a marginal service. Incorporated into this unit are 75 bps dual diversity FSK modems which also provide the data link for establishing a call to a subscriber.

In the RACE system the channel evaluation is done by transmitting data on available channels during the idle periods when the system is not occupied by radio telephone calls. Each Master station transmits a burst of FSK data, called an idle message, on each channel in turn, and all remote stations synchronize themselves to the Master station and evaluate the received data by assessing the error bit rate. If no idle message is received, the Remote station automatically steps to the next frequency channel maintaining local short term synchronization. The evaluation time is two seconds per channel.

When a call originates from a Master station the idle message is replaced by a call message directed to a specific Remote station. If the Remote station receives the call without error, it sends a message to the Master station. If an error free call message is not received, the sequence will be repeated on the next frequency. When the Master station receives a reply to its call, it also analyses the quality of the message to determine its agreement with the selected frequency. If it agrees, it sends as acknowledgement message or "handshake", to the Remote station. After the "handshake", the master sends a message to the Remote to ring the called subscriber. The Remote checks the status of the line and if the line is free it rings the subscriber. When the subscriber answers, a call connect message is sent to the Master which switches the call to the HF link. The average time taken to establish a call is 6 seconds with a maximum time for a eight channel system of 16 seconds. The steps taken by the system to establish a call from the Remote to the Master is similar to that just described from the Master to the Remote.

For a call between Remote stations, Remote A sends information to the Master on an idle frequency f1 requesting a call to a subscriber at Remote B. The Master passes control of the HF network to Remote A which sends a call request on another frequency f2 to Remote B. If the frequency is acceptable, Remote B sends a reply and Remote A checks the reply and if acceptable sends an acknowledgement. Remote A then contacts the Master on f1 and indicates the call will be made on another frequency on f2. Remotes A and B establish the call over the HF link and return to the idle condition when the call is completed. On the first clear idle message received by Remote A an "end of call" message will be sent to the Master who will check the message and if acceptable send an acknowledgement.

The data channel is in continuous use while the system is in operation and its performance is critical to the reliable operation of the system. The data channel uses a 75 bps modem employing binary FSK (85 Hz) shift with in band frequency diversity using 1105 Hz and 2125 Hz. The reliability of the channel is enhanced by the use of error detection coding and "stop and wait" ARQ protocol. The data channel serves the following functions which occur sequentially and are exclusive:

- Network synchronization and sounding
- Call set up
- Syncomplex control
- Call termination.

Details on the message format for the data channel are described by Derbyshire (1982). The data is organized into eight bit units with a minimum message length of 48 bits.

The real time channel evaluation involves the Master station transmitting a 48 bit idle message on each frequency every 2 seconds and each Remote station receiving and evaluating this message and keeping statistics on the channel quality. The evaluation uses real and pseudo errors detected on the incoming idle message. Real errors are accumulated over approximately four minutes with a weighting factor assigned to each measurement so as to follow rapidly changing channel conditions. When the is busy, a four minute time period is not of sufficient length to evaluate real errors. Under these conditions, an evaluation technique was selected based upon "pseudo error" analysis of the incoming data (Gooding, 1968). Pseudo error counts are found to be a suitable measure of channel quality. An algorithm is also incorporated in RACE to select the best channel for HF communications.

Field trials conducted in 1980 and 1981 for Master to Remote station distances of 65, 270, 490, and 965 kms confirmed the superiority of the dual diversity FSK over the single channel FSK for evaluating the best channel for voice communications. The single channel FSK selected channels with the smallest multipath spreads which are not necessary those with the best signal-to-noise ratio. The call completion rate during the test periods was estimated to be greater than 98% using a low power transmitter of 100 watts and simple non-directional broadband antennas. The availability of two and three channels for communications are as follows

| | |
|---|---|
| two channel availability | 96% |
| three channel availability | 86% |

These data on channel availability indicate the RACE system can support several simultaneous calls from a Master station. Even though the RACE system does not take into account non-reciprocity in propagation or different noise and interference levels at both ends of the circuit, it did not appear to be a major limitation in the performance of the system.

## 3. HF MESSAGE TERMINAL

A message terminal was developed at the Communications Research Centre in Ottawa, Canada to increase the capabilities of existing HF radio systems by permitting the transmission of text message. The message terminal can be connected directly to an existing system with a 600 ohm input/output ports. Coding and modulation techniques are incorporated in the HF data protocol to enable data communications when propagation conditions do not permit intelligible voice transmissions.

The system consists of a portable terminal with an alphanumeric keyboard, a hard copy printer and a single line display. The user types in a message have keyboard which appears on the terminal display panel. The outgoing message have a 1280 character buffer memory which hold the prepared text prior to transmission. The message can be corrected, updated or sent immediately. A typical message, about four lines in length, can be transmitted and confirmed on both sender and receiver terminals in 40 seconds. The destination terminal receives the incoming message with out operator assistance. The destination terminal receives the incoming message with out operator assistance. The outgoing and incoming messages are printed by a small hard copy printer in a 80 character by 10 line page format with a one inch gap between pages.

The terminal has a 75 bps dual channel FSK modem. The modem is implemented with a microprocessor and free from drift, aging does not require high precision components. The modem uses frequency diversity for reliable operation of the device during selective fading periods. The data transmission occupies 300 Hz of voice channel enabling more than one network to be operational in a 3 kHz bandwidth.

The following types of calls are possible with the message terminal;

a) Selective Call: Each terminal can call any other terminal on the same network, with ARQ protocol. This mode is very reliable for error free messages.

b) Broadcast Call: All terminals recognized the global address but do not answer the call. The message is transmitted several times and the terminals accept correct parts of the message. This mode is not as reliable as the selective type of call.

c) Privacy Call: When this option is selected, the terminal asks for the password which is used to start data randomization operation. The destination terminal upon receipt of this type of message only displays the calling stations call sign and "ENTER PASSWORD". The incorrect entry of the password lets the operator try three times and then erases the complete message. The correct entry of the password performs inverse of the randomization operation and the printing of the message.

The terminal has a RS-232 port for external equipment connection of the following equipment;

a) CRT and printer
b) Telephone line or short haul modem for remote control
c) Mass storage facilities

The message terminal increases the capabilities of the existing HF radio systems to transmit short text messages. The terminal can be connected easily to most conventional HF systems. The modulation and encoding are built into the data link protocol to allow data communications under conditions that do not permit intelligible voice communications. The terminals allow for unattended reception of the messages and operational communication privacy. The terminal has been tested over numerous short, medium and long range HF circuits, including transauroral paths, and the combination of dual diversity and selective-ARQ protocol has proven to be very effective.

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# RACE

## Subscriber Line Interface System

with the Remote Stations

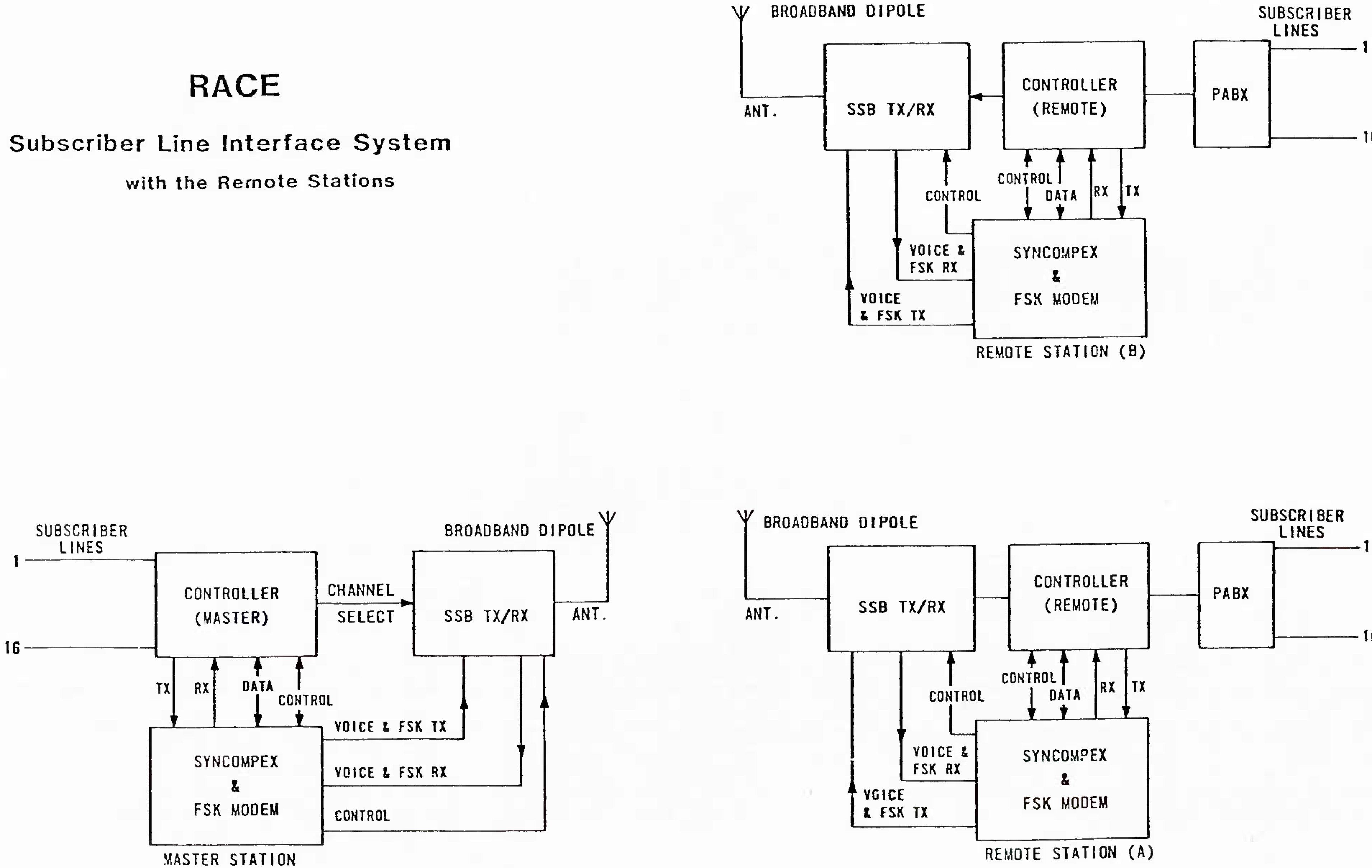



Figure 1

## HORIZONTAL HALF-WAVE DIPOLE ANTENNA ERECTED OVER THREE TYPES OF GROUND

poor ——— good - - - sea-water ........

Frequency - 14 MHz
Bearing - 0 degrees
Height - 20 meters
Length - 0.50 wavelengths

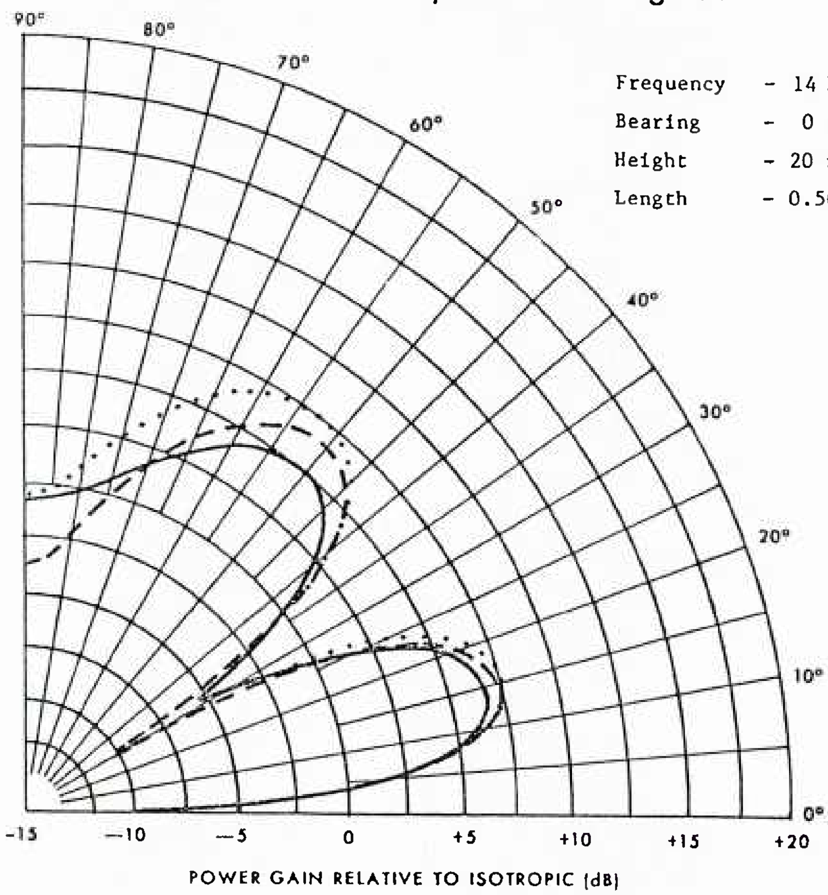



Fig. 2

## VERTICAL ANTENNA ERECTED OVER THREE TYPES OF GROUND

poor ——— good - - - - sea—water ....

Frequency - 6 MHz
Length - 0.25 wavelengths

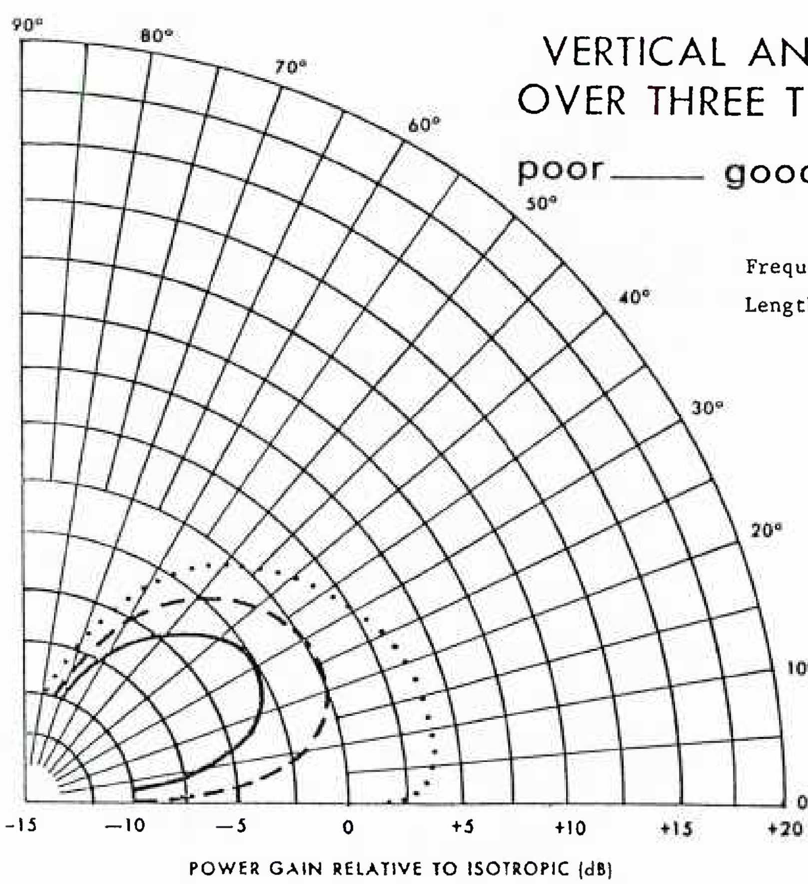
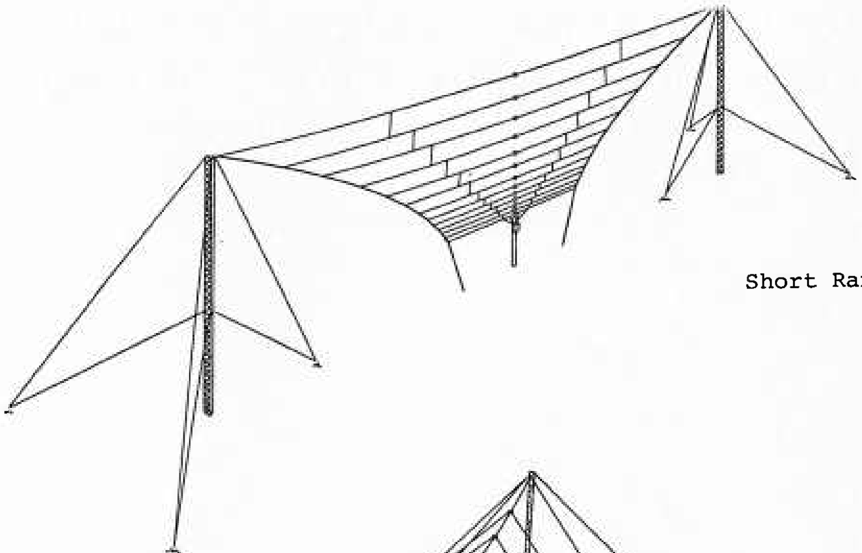
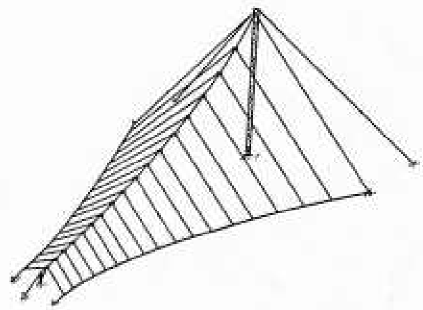



Fig. 3



Short Range



Intermediate
Range

Long Range

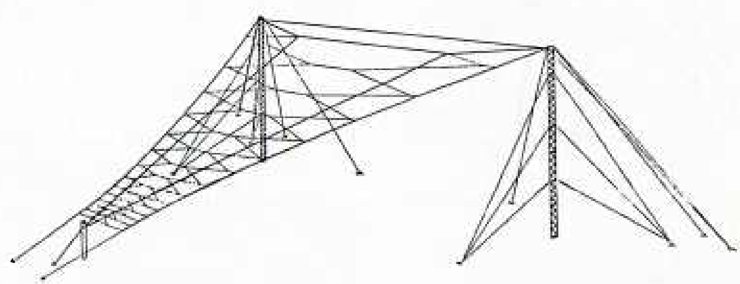


Fig. 4

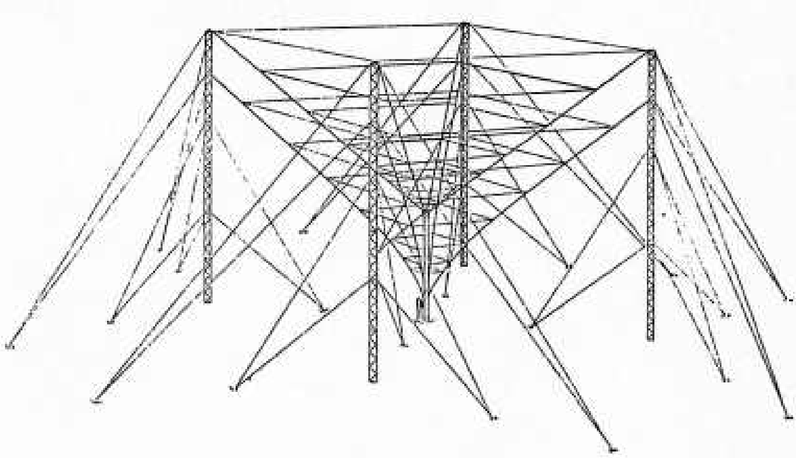


Fig. 5

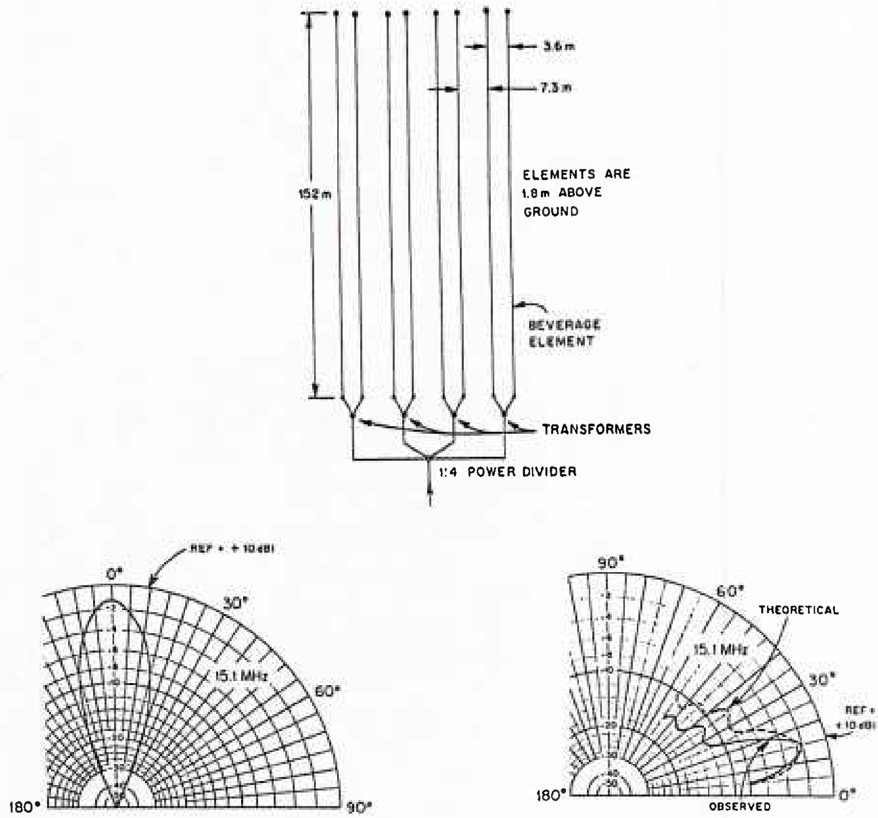

3.6 m
7.3 m
152 m
ELEMENTS ARE
1.8 m ABOVE
GROUND
BEVERAGE
ELEMENT
TRANSFORMERS
1:4 POWER DIVIDER
REF = +10 dBi
0°
30°
60°
90°
180°
15.1 MHz
THEORETICAL
OBSERVED

Fig. 6

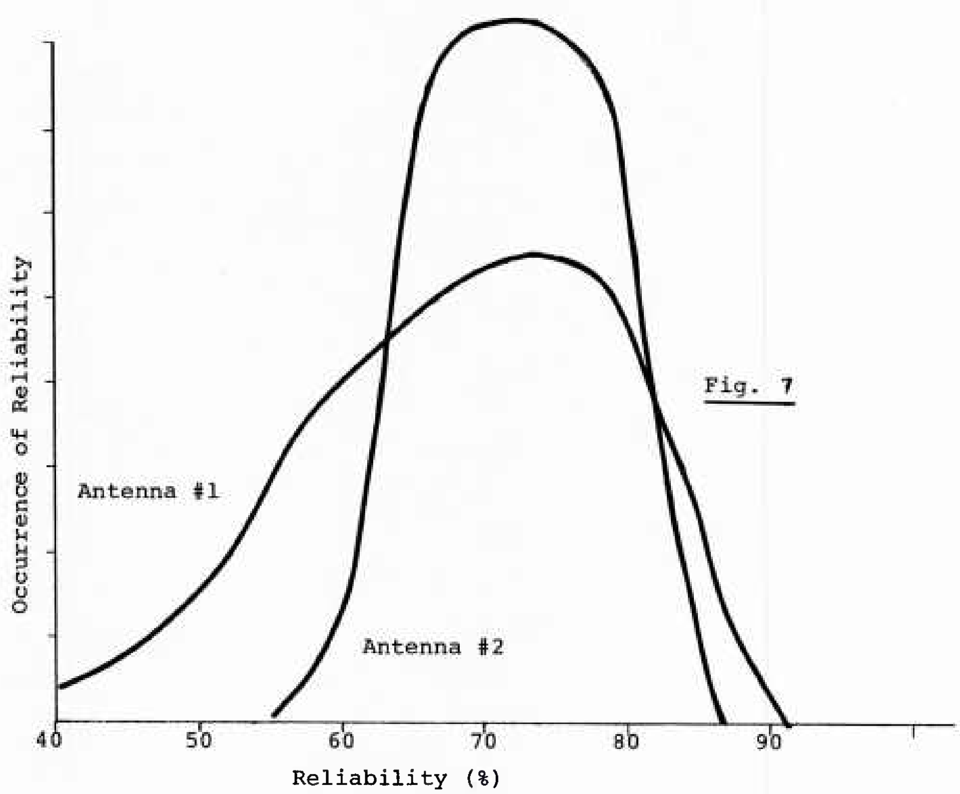

Occurrence of Reliability
Antenna #1
Antenna #2
40
50
60
70
80
90
Reliability (%)

Fig. 7

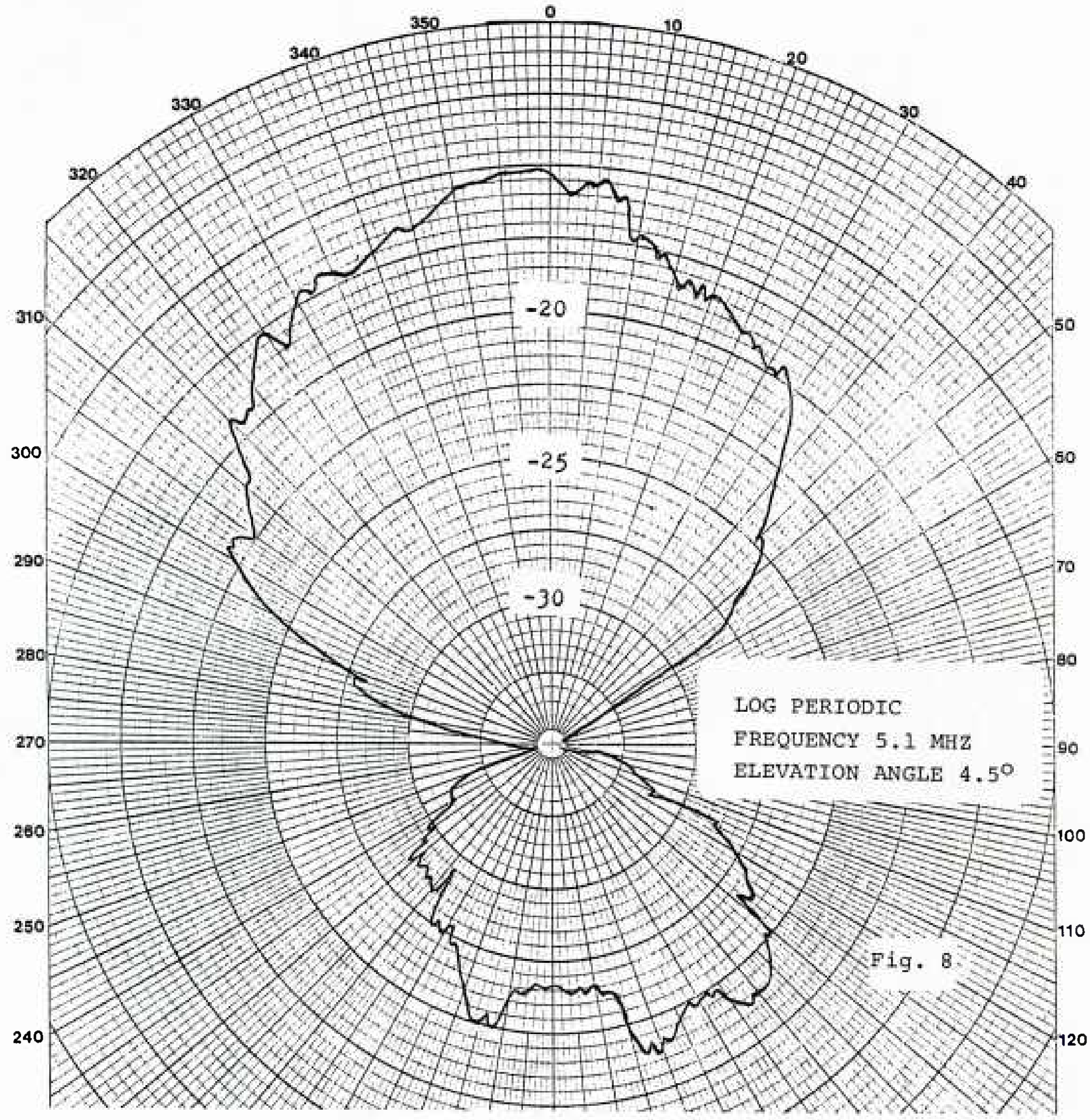

350
0
10
340
20
330
30
320
40
310
50
300
60
290
70
280
80
270
90
260
100
250
110
240
120
-20
-25
-30
LOG PERIODIC
FREQUENCY 5.1 MHZ
ELEVATION ANGLE 4.5°

Fig. 8

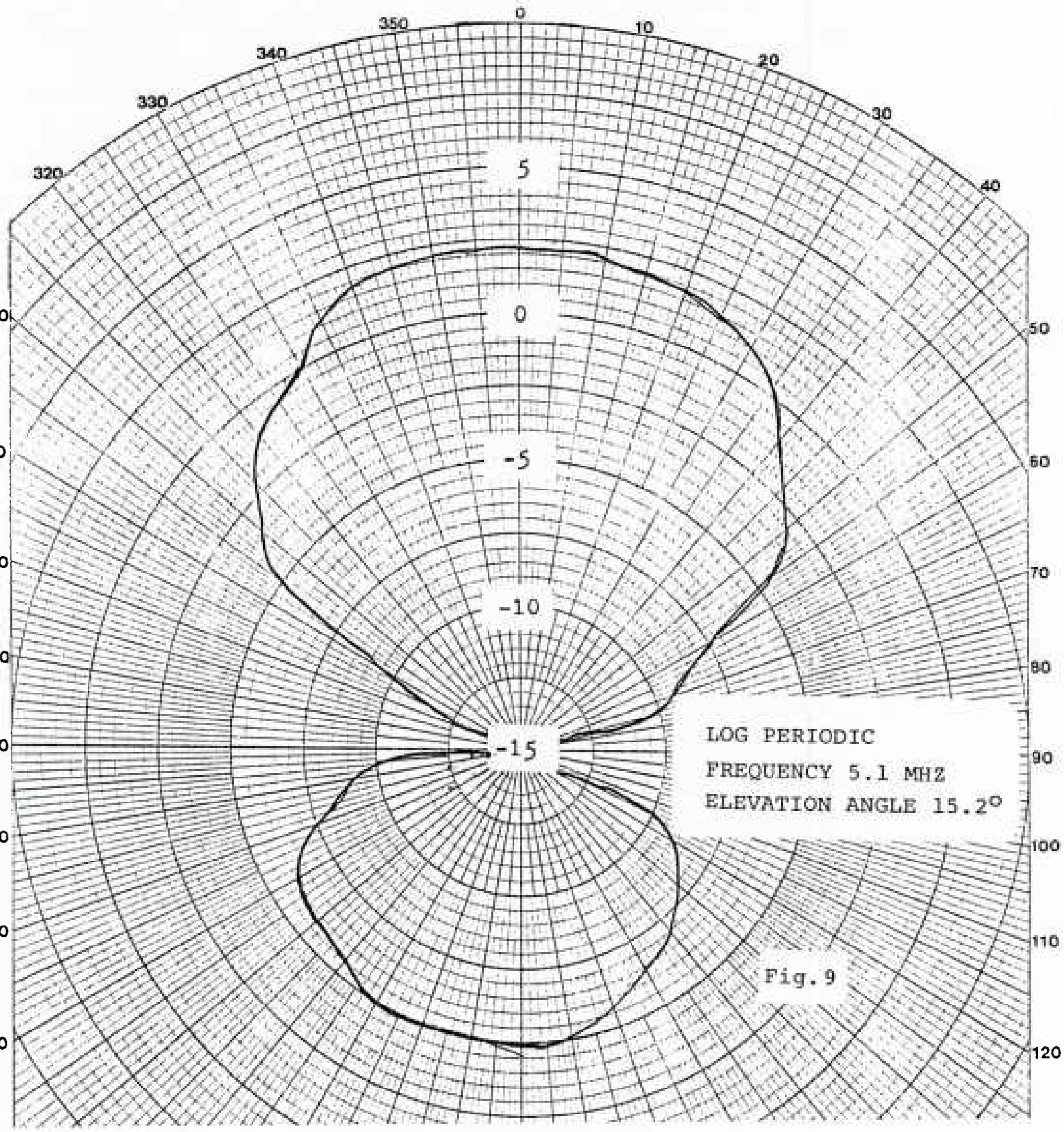

LOG PERIODIC
FREQUENCY 5.1 MHZ
ELEVATION ANGLE 15.2°
5
0
-5
-10
-15
0
10
20
30
40
50
60
70
80
90
100
110
120
240
250
260
270
280
290
300
310
320
330
340
350

Fig.9

Antenna Radiation Pattern

Elevation Plane

DELTA
FREQUENCY 5.1

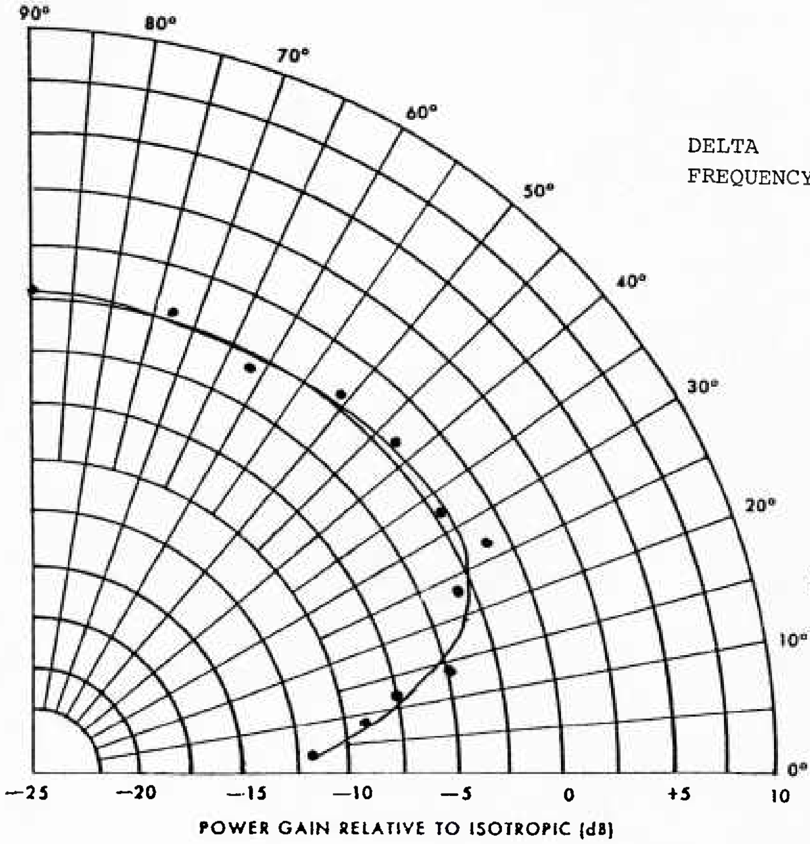



Fig 10

# SELECTIVE BIBLIOGRAPHY

This bibliography with Abstracts has been prepared to support AGARD Lecture Series No. 145 by the Scientific and Technical Information Branch of the US National Aeronautics and Space Administration, Washington, D.C., in consultation with the Lecture Series director, Dr. J. Aarons, Department of Astronomy, Boston University, Boston, Massachusetts.

PRINT 06/4/1-86 TERMINAL=68

UTTL: Radar-82; Proceedings of the International Conference, London, England, October 18-20, 1982

ABS: Topics related to radar systems are considered, taking into account intrapulse polarization agile radar, search strategies of phased array radars, design and performance considerations in modern phased array radar, a new generation airborne synthetic aperture radar system, results from a new dual band radar for sea surface and aircraft search, modular survivable radar for battlefield surveillance applications, and the Dolphin naval surveillance radar. Other subjects explored are concerned with sequential detection and MTI, adaptive processing techniques, HF/VHF radar, coherent radar processing, multisite radar operation, radar sea clutter, air traffic control, simulation and data processing, aspects of target recognition, low probability of intercept radar and passive operation, signal processing, low sidelobe antennas, and radar returns from weather and land. Attention is also given to beam forming with phased array antennas, optical fiber networks for signal distribution and control in phased array radars, and radar tracking systems. For individual items see A84-10752 to A84-10841 82/00/00 84A10751

UTTL: Experimental determination of wide-band propagation parameters in the HF spectrum CORP: Worth Research Associates, La Jolla, Calif.

RPT#: AD-A141936 AD-E800940 TM-129(DRAFT) 82/00/00 84N75156

UTTL: Design of an adaptive high-frequency data link

AUTH: A/AMES, J.; B/STEIN, S.; C/IRVINE, G. W. PAA: A/(SRI International, Menlo Park, CA); B/(SCPE, Inc., Newton Centre, MA); C/(Canadian Embassy, Washington, DC)

ABS: A design for a HF data link for automated over-the-ocean ATC is described. The design goals were to service 200 aircraft in flight over the North Atlantic, assuming 70 provide automatic position reports. The link includes several unattended ground stations and two control centers interconnected by satellite or cable with a full duplex 2400 pbs data rate. Ten bands covering 22 MHz are utilized to maximize network reliability given the variety of ionospheric, atmospheric and sea states, and distances faced. A polling protocol provides the bulk of the oceanic ATC data link requirements, with polling taking place in 10 sec epochs for individual aircraft, which could be switched among families of frequencies. Account is taken of interference, acquisition, and recovery mode problems and procedures. 83/00/00 84A49285

UTTL: Computer aided antenna design and frequency selection for HF communications

AUTH: A/AVERILL, W. P. CORP: Naval Postgraduate School, Monterey, Calif.

ABS: Efficient and reliable high frequency skywave communication can be obtained only if wave propagation paths are carefully analyzed and if an antenna which transmits and receives signals along the desired paths can be selected and designed. A micro-computer based program was developed to accomplish this task.

RPT#: AD-A150485 84/06/00 85N22869

UTTL: The design and simulation of a mobile radio network with distributed control

AUTH: A/BAKER, D. J.; B/FLYNN, J. A.; C/EPHREMIDES, A. PAA: B/(U.S. Navy, Naval Research Laboratory, Washington, DC); C/(U.S. Navy, Naval Research Laboratory, Washington, DC; Maryland, University, College Park, MD)

ABS: A new architecture for mobile radio networks, called the linked cluster architecture, is described, and methods for implementing this architecture using distributed control techniques are presented. It is illustrated how fully distributed control methods can be combined with hierarchical control to create a network that is robust with respect to both node loss and connectivity changes. Two distributed algorithms are presented that deal with the formation and linkage of clusters and the activation of the network links. To study the performance of the network structuring algorithms, a simulation model was developed. The use of Simula to construct software simulation tools is illustrated. Simulation results are shown for the example of a high frequency (HF) intratask force (ITF) communication network. 84/01/00 84A32823

UTTL: The diffraction of VLF radio waves by a patch of ionosphere illuminated by a powerful HF transmitter

AUTH: A/BARR, R.; B/RIETVELD, M. T.; C/STUBBE, P.; D/KOPKA, H. PAA: A/(Geophysical Observatory, Christchurch, New Zealand; Max-Planck-Institut fuer Aeronomie, Katlenburg, West Germany); D/(Max-Planck-Institut fuer Aeronomie, Katlenburg, West Germany)

ABS: The effects of heating a fragment of auroral ionosphere with powerful HF radiation on the short-path propagation parameters of VLF waves (12.1 and 16.4 kHz) are investigated. Typical amplitude and phase perturbations of + or - 0.05 dB and + or 0.5 deg were observed with effective radiated powers of 200 MW and an effective antenna beam width of 15 deg. The amplitude and phase perturbations at a fixed beam deflection angle (38 deg S) were found to reach 6 dB and 50 deg, respectively, during minimum VLF field strength values. Moreover, the impact of the diffractive effects of such heating of the ionosphere on the multimode propagation in the earth-ionosphere waveguide are assessed theoretically for daytime and nighttime conditions, and the normal and enhanced amplitude and phase perturbations are modeled and compared to observations. 85/03/01 85A26966

UTTL: Coordinated study of subkilometer and 3-m irregularities in the F region generated by high-power HF heating at Arecibo
AUTH: A/BASU, S.; B/BASU, S.; C/GANGULY, S.; D/GORDON, W. E. PAA: B/(Emmanuel College, Boston, MA); D/(Rice University, Houston, TX)
ABS: High-power HF transmitters near Arecibo were used to generate artificial ionospheric irregularities in the F region. Radio-star scintillation observations at 430 MHz were performed at Arecibo with the 305-m antenna, and radar-backscatter measurements at 50 MHz were simultaneously made from Guadeloupe Island to probe the subkilometer and 3-m irregularities in the heated volume. Scintillation studies indicate a low-frequency modulation of the faster intensity-fluctuation structure, which translates to spatial dimensions of 1-2 km. The frequency of the modulation envelope is found to be controlled by the heater power and is related to the dominant irregularity wavelength generated by the self-focusing instability. Scintillation spectra imply a steep power-law index of -5 in the scale-length range of about 300-150 m and a shallow index of -2 at less than 150 m. Simultaneous measurements of 430-MHz scintillations and 50-MHz radar backscatter from field-aligned striations show that subkilometer irregularities can be generated by both O and X-mode heating, whereas the 3-m irregularities are excited only by the O-mode heating, as is predicted by the theories of self-focusing and parametric instability. The width of the 50-MHz echo Doppler spectra is observed to be only about 2-3 Hz, and independent of the background plasma drift, implying that the frequency bandwidth of the scattered signal is probably controlled by the instability process.
RPT#: AD-A136606 AFGL-TR-83-0322 83/11/01 84A12906

UTTL: The effect of solar and lunar currents on simultaneous phase path, group path and amplitude measurements
AUTH: A/BAULCH, R. N. E.; B/BUTCHER, E. C. PAA: A/(Commonwealth Scientific and Industrial Research Organization, Div. of Textile Physics, Ryde, New South Wales; La Trobe University, Bundora, Victoria, Australia); B/(La Trobe University, Bundora, Victoria, Australia)
ABS: The solar and lunar variations in the phase path, group path and amplitude of a fixed frequency transmission were obtained at the September equinox over a slightly oblique path. The phase of the lunar semidiurnal tide in the phase path and amplitude were similar, the maxima occurring near 0200 lunar time, whereas the group path had a maximum near 0800 lunar time. These results were compared with other results obtained near the same location. The results suggest a complex situation in the E-region, where the height of the lunar current depends on season, and also suggest that the location and distribution of the solar and lunar currents may be different. 84/10/00 85A15382

UTTL: A simple sounder to measure the properties of ionospherically reflected radio waves
AUTH: A/BAULCH, R. N. E.; B/BUTCHER, E. C.; C/HAMMER, P. R.; D/DEVLIN, J. C. PAA: A/(Commonwealth Scientific and Industrial Research Organization, Div. of Textile Physics, Ryde, New South Wales; La Trobe University, Bundora, Victoria, Australia); C/(La Trobe University, Bundora, Victoria, Australia); D/(Andrew Antennas, Cambellfield; La Trobe University, Bundora, Victoria, Australia)
ABS: A system which measures the direction of arrival, amplitude, group path and phase path of high frequency radio waves reflected from the ionosphere is described. A CW double sideband modulated signal was used and the measurements were made over a slightly oblique path. Comparisons between the group height determined using this ionospheric sounder and an ionosonde located near the midpoint of the transmission are given and results of fast fluctuations in the measurements are presented. 84/10/00 85A15381

UTTL: Propagation and scattering considerations for the 0.2-3.0 GHz frequency range for a space-based radar
AUTH: A/BILOW, H. J. CORP: Naval Research Lab., Washington, D. C.
ABS: A survey is made of major phenomena of the lower

atmosphere and ionosphere which would introduce clutter and affect propagation in the 0.2-3.0 GHz frequency range in a space-based, ground-looking radar. Attenuation, refractive effects, and weather-related phenomena are included in the study. It is found that the major trade-off that must be made in the frequency range is the relative immunity of the low frequencies to weather related phenomena versus the relative immunity of the high frequencies to deleterious ionospheric effects.
RPT#: AD-A148504 NRL-MR-5453 84/12/18 85N17282

UTTL: Interaction of an acoustic wave of artificial origin with the ionosphere as observed by vertical HF sounding at total reflection levels
AUTH: A/BLANC, E. PAA: A/(Commissariat al'Energie Atomique, Laboratoire de Detection et de Geophysique, Bruyeres-le-Chatel, Essonne, France) 84/04/00 84A28220

UTTL: A new HF-link parameters/quality analysis approach
AUTH: A/BLISS, D. H. PAA: A/(Rockwell International Corp., Collins Telecommunications Products Div., Cedar Rapids, IA)
ABS: A new approach to real-time HF-link parameters measurement and quality analysis is presented, which is applicable to optimization of high-speed HF data links. The approach involves a technique based on digital signal processing for measurement of general HF parameters and uses simple in-band test signals and algorithmic estimation of data transmission performance. A feasibility model of an Advance Link Quality Analyzer was developed and extensively tested. The model performs spectral analysis calculations and initial software processing while sampling the analog signal at a maximum rate of 25 blocks/s. Experimental results are compared with non-real-time channel-simulation results and transmission tests. 83/00/00 85A34501

UTTL: Experimental evaluation of a high frequency radio data terminal in the ship-shore mode
AUTH: A/BODE, L.; B/CHOW, S. M.; C/SERINKEN, N.; D/TEPPER, B.; E/YAMANOTO, R. CORP: Communications Research Centre, Ottawa (Ontario). CSS: (Information Technology and Systems Branch.)
ABS: A field trial of a ship/shore direct printing HF radio message transmission system is described. The field trial was conducted between a Canadian Coast Guard ship in the Arctic and a radio station in Ottawa over a six week period. Four hundred and seventy thousand characters were passed with no errors due to the ARQ protocol. The average transmission rate attained was 300 to 500 characters per minute. It is indicated that the ARZ protocol used is both effective and efficient.
RPT#: CRC-1375 84/09/17 84N31458

UTTL: Radio receivers with a wide dynamic range
AUTH: A/BOGDANOVICH, B. M.
ABS: The present work explores the theoretical foundations, design principles and methods, and network solutions for HF receivers operating in conditions of relatively large changes of input-signal and noise levels. Factors determining the dynamic range of such receivers are examined, and principles for the optimization and adaptation of the receiver sections according to the dynamic-range criterion are presented. Examples of appropriate structural and network solutions are given. 84/00/00 84A48731

UTTL: Error correction coding for a self-adaptive modem: Theoretical and experimental results
AUTH: A/CHAVAND, F.; B/GOUTELARD, C.; C/HARARI, S. CORP: Paris-Sud Univ., Cachan (France). CSS: (Lab. d'Etude des Transmissions Ionospheriques.) In AGARD Propagation Influences on Digital Transmission Systems: Probl. and Solu. 16 p (SEE N85-19269 10-32)
ABS: An error correcting system designed for a 1200 bits/s self adaptive modem operating on the HF ionospheric channel is presented. Error statistics were analyzed by simulating a specific link whose characteristics had been measured previously by a HF backscatter probe. Some 30 millions of bits were processed which represent 27 different cases of propagation and scrambling. The coding system consists of a cascade of two codes: one corrects single error and small bursts (Reed Solomon codes); the other corrects long bursts of errors (Kasama code). The overall rate is 0.55. An error correcting simulation shows that least 10 -2 has been achieved. Limitations of these correction methods on such systems are discussed. The total error was reduced by a factor equal to or greater than 100 in all cases. 84/10/00 85N19303

UTTL: Principles of digital data transmision (2nd edition)
AUTH: A/CLARK, A. P. PAA: A/(Loughborough University of Technology, Loughborough, Leics., England)
ABS: A textbook on the principles and techniques of digital data transmission is presented. A nonmathematical survey of the properties of the voice-frequency

channels formed by telephone circuits and HF radio links, and of various techniques that have been used or proposed for the transmission of digital data over these channels is given. The different techniques are compared and descriptions are provided of the preferred data transmission systems. A theoretical analysis and comparison of various different signals that may be used for the transmission of digital data is then presented. Both baseband and modulated-carrier signals are studied, with particular emphasis on some of the baseband signals that may be obtained over partial-response channels and by correlative-level coding. Matched filter detection is studied in some detail as is the related topic of the optimum combination of transmitter and receiver filters.
83/00/00 84A11014

UTTL: The temporal evolution of 3-m striations in the modified ionosphere

AUTH: A/COSTER, A. J.; B/DJUTH, F. T.; C/JOST, R. J.; D/GORDON, W. E. PAA: A/(MIT, Lincoln Laboratory, Lexington, MA; Rice University, Houston, TX); B/(Aerospace Corp., Space Sciences Laboratory, Los Angeles, CA); C/(NASA, Johnson Space Center, Houston, TX); D/(Rice University, Houston, TX) CORP: LQ054005; Rice Univ., Houston, Tex.; Aerospace Corp., Los Angeles, Calif.; National Aeronautics and Space Administration. Johnson (Lyndon B.) Space Center.

ABS: Experiments were performed at Arecibo, Puerto Rico, to investigate the evolution times of 3-m field-aligned striations produced in the ionosphere by powerful high-frequency (HF) radio waves. The results of this investigation are now summarized. First, the striations' rise times are dependent on the HF electric field. The E region data suggest that this dependence is nonlinear. Second, the threshold value of the HF electric field required to produce detectable striations was experimentally determined. At threshold the component of the HF electric field perpendicular to the geomagnetic field is calculated to be 0.09 V/m in the F region and 0.37 V/m in the E region. Third, both the E and the F region data verify theoretical predictions that the striations' decay times are directly proportional to the electron diffusion across B. Finally, a one-to-one correspondence between the growth of the 3-m striations and the decline of the HF-enhanced plasma line during overshoot is sometimes observed.
85/03/01 85A26962

UTTL: Large simultaneous disturbances (LSDs) in the Antarctic ionosphere

AUTH: A/CROWLEY, G.; B/JONES, T. B.; C/DUDENEY, J. R.; D/RODGER, A. S. PAA: B/(Leicester, University, Leicester, England); D/(Natural Environment Research Council, British Antarctic Survey, Cambridge, England)

ABS: A new class of ionospheric disturbance has been observed in the 10 min-1 hr period range, during a high frequency Doppler experiment in the Antarctic Peninsula region whose three transmitted dual frequency network permitted horizontal and vertical propagation velocities to be estimated. The disturbances are characterized by unusually good correlation between perturbations on all available Doppler signals, although they are apparently nonpropagating and occur simultaneously at each reflection point. In virtue of the large Doppler shifts displayed, these events have been designated Large Simultaneous Disturbances (LSDs). Criteria for their evaluation are established, and a detailed analysis of one LSD is undertaken. Occurrence statistics are presented, including seasonal and diurnal distributions. 84/10/00 85A15384

UTTL: Finite element approach to propagation in elliptical and parabolic waveguides

AUTH: A/DALY, P. PAA: A/(Leeds University, Leeds, England)
84/04/00 84A38449

UTTL: Reflection of high-frequency radio waves in inhomogeneous ionospheric layers

AUTH: A/DAVIES, K.; B/RUSH, C. M. PAA: A/(NOAA, Space Environment Laboratory, Boulder, CO); B/(National Telecommunications and Information Administration, Institute for Telecommunication Sciences, Boulder, CO)

ABS: Formulas are given for the calculation of rays in model ionized layers in which the variation of the square of the refractive index is expressed separately in terms of the horizontal (x, y) and vertical (z) coordinates. Examples of three-dimensional paths in linear-parabolic layers are presented showing reflections in the horizontal, as well as the vertical, direction. To illustrate long-distance transequatorial propagation the equatorial anomaly is simulated by two elliptic cylinders with axes parallel to the equator. Sample calculations show that in this model the ground range can increase as the angle of elevation at the ground increases. 85/06/00 85A36530

UTTL: High-frequency ray paths in ionospheric layers with horizontal gradients
AUTH: A/DAVIES, K.; B/RUSH, C. M. PAA: A/(NOAA, Space Environment Laboratory, Boulder, CO); B/(National Telecommunications and Information Administration, Institute for Telecommunication Sciences, Boulder, CO)
ABS: The results of a study to determine the effects of horizontal gradients in the F2 region electron density on short-distance HF propagation paths (0-1000 km) are presented. Ray paths were obtained by employing specific models to represent the horizontal and vertical variations of the F2 region. These models permitted ray path characteristics to be expressed by algebraic relationships. The changes in ray path characteristics with height of reflection, frequency, and azimuth are presented for gradients typically encountered on short-distance HF propagation circuits over Europe. 85/02/00 85A21134

UTTL: An H.F. simulator for use with real time channel evaluation systems
AUTH: A/DAWSON, J. CORP: York Univ. (England). In AGARD Propagation Influences on Digital Transmission Systems: Probl. and Solu. 15 p (SEE N85-19269 10-32)
ABS: A high frequency (HF) simulator constructed at York University as an off-air test bed for an HF radio link using real time channel evaluation aided frequency selection is discussed. The simulator implements the basic channel model proposed by Watterson et al Watterson 69a with the addition of several novel features applicable to real time channel evaluation systems: rapid selection of channel parameters from stored sets; a realistic and repeatable interference simulation; and time variable channel statistics. A brief resume of channel modelling and atmospheric noise is given. 84/10/00 85N19293

UTTL: Observations and prognosis of HF wave propagation via the sporadic E layer
AUTH: A/DERBLOM, H. CORP: Uppsala Ionospheric Observatory (Sweden). Presented at RVK 84, Linkoeping, Sweden
ABS: Sporadic E propagation over a distance of 1000 km for 1 month was studied. Analysis of ionograms from a site close to the midpoint of the propagation path show that the median values of the maximum propagation observed (Es MOF) can be predicted accurately for daytime conditions. A great discrepancy between the observed and predicted values is reported for nighttime. Es MOF exceeds the maximum usable frequency (EsMUF) by several MHz. The reflection mechanism is possibly different during the night. Comparison between maximum transmission frequencies calculated by ray tracing using full wave theory, and by the secant law shows marginal differences only. Refraction in layers beneath the Es can be neglected in practical applications for EsMUF predictions.
RPT#: UIO-SR-84-04 ISSN-0349-2699 84/04/00 85N12294

UTTL: HF-enhanced plasma lines in the lower ionosphere
AUTH: A/DJUTH, F. T. PAA: A/(Aerospace Corp., Space Sciences Laboratory, Los Angeles, CA)
ABS: The ionospheric modification facility at Arecibo, Puerto Rico, has been used to study Langmuir waves excited in the lower ionosphere by a high-power HF radio wave. Measurements of HF-enhanced plasma lines have been made in the lower F region (less than 160 km altitude) and in the E region using the 430-MHz radar at Arecibo Observatory. These measurements complement observations made in the past at higher altitudes. Enhanced plasma line spectra observed in the lower F region peak at the so-called 'decay line' and contain additional spectral structure similar to that found in the upper F region. In the E region the spectra exhibit narrow peaks located at 430 MHz + or - f(HF), where f(HF) is the frequency of the modifying HF wave. While strong plasma line enhancements are commonly observed in sporadic E, only weak enhancements have been detected in the normal daytime E layer. Candidate mechanisms for the E region enhancements include the oscillating two-stream instability and direct conversion of the HF wave into Langmuir waves. 84/02/00 84A20784

UTTL: Ionospheric modification spawns new propagation
AUTH: A/DJUTH, F. T. PAA: A/(Aerospace Corp., Space Sciences Laboratory, Los Angeles, CA)
ABS: Transmissions at frequencies as high as a few gigahertz can be affected by electron-density irregularities in the ionosphere. The scattering of radio waves by these irregularities does not always result in degradation of the performance of a communications system. Under certain conditions, the generated scattering surfaces can be useful for low-loss ground-to-ground communications. In the case of high-power radio waves, the signal can modify the ionospheric medium in which it propagates. Modification of radio waves by the ionosphere was first noticed when a 625-kHz signal transmitted from Beromuenster, Switzerland, to Eindhoven, Netherlands, was inadvertently modulated by a signal from a radio station in Luxembourg. It was suggested that such cross-modulation of radio signals could be used to probe the lower ionosphere. Ground-based facilities were built for studying the interactions of high-power

waves in the ionosphere. The experiments conducted are discussed along with the obtained results. 85/08/00 85A45324

UTTL: High-resolution studies of the HF ionospheric modification interaction region
AUTH: A/DUNCAN, L. M.; B/SHEERIN, J. P. PAA: A/(Los Alamos National Laboratory, NM); B/(Iowa, University, Iowa City) CORP: Los Alamos Scientific Lab., N. Mex. ; Iowa Univ., Iowa City.
ABS: The use of the pulse edge analysis technique to explain ionospheric modifications caused by high-power HF radio waves is discussed. The technique, implemented at the Arecibo Observatory, uses long radar pulses and very rapid data sampling. A comparison of the pulse leading and trailing edge characteristics is obtained and the comparison is used to estimate the relative changes in the interaction region height and layer width; an example utilizing this technique is provided. Main plasma line overshoot and miniovershoot were studied from the pulse edge observations; the observations at various HF pulsings and radar resolutions are graphically presented. From the pulse edge data the development and the occurrence of main plasma line overshoot and miniovershoot are explained. The theories of soliton formation and collapse, wave ducting, profile modification, and parametric instabilities are examined as a means of explaining main plasma line overshoots and miniovershoots. 85/09/01 85A49217

UTTL: A design for an automatic HF radio system
AUTH: A/ELVY, S. J. PAA: A/(S.H.A.P.E., Technical Centre, The Hague, Netherlands)
ABS: Microprocessor control systems, coding schemes, and modern communications equipment suggest new possibilities for the operation of HF radio systems. This paper describes a design concept for an automatic radio system, intended to overcome some of the traditional difficulties experienced with HF communications and to provide new capabilities including selective calling and automatic connectivity. The design incorporates a passive frequency selection technique using built-in ionospheric prediction data and real time monitoring of channel interference levels. 85/06/00 85A36526

UTTL: The HF-induced plasma line below threshold
AUTH: A/FEJER, J. A.; B/SULZER, M. PAA: B/(Arecibo Observatory, Arecibo, PR)
ABS: Observations at Arecibo of the HF-induced plasma line for low powers of the HF modifying transmitter are described and interpreted. The observed spectra were centered on the radar frequency minus the HF modifying frequency and had the same shape and width as the ion line of incoherent backscatter. The importance of the standing-wave nature of the HF pump wave for such observations is stressed. 84/04/00 84A28222

UTTL: Novel ways for tracking rays
AUTH: A/FELSEN, L. B. PAA: A/(New York, Polytechnic Institute, Farmingdale, NY)
ABS: Rays no longer just represent the trajectories traversed by a beam of light but they describe the general transport properties of high-frequency fields. Viewed within the framework of wave spectra, ray fields are localized spectral objects synthesized by constructive spectral interference. The spectral flesh within these objects can be trimmed to the bare bones for the simplest type of ray field, a local plane wave, but must be augmented in ray transition regions near caustics, shadow boundaries, diffraction centers, etc., where more-intricate wave phenomena occur. Thus ray tracing along ray paths generally involves a succession of bare-bones and fleshed-out (uniformized) spectral entities that can furnish in a highly effective manner the amplitude and the phase of the high-frequency field in a complex environment. Relevant spectral concepts are discussed, including those for complex rays that describe evanescent fields and Gaussian beams and those leading to collective treatment of multiple reflected or diffracted contributions. 85/06/00 85A37058

UTTL: On the use of virtual reflection height information to interpret single-frequency A1 absorption measurements
AUTH: A/FERGUSON, B. G. PAA: A/(Department of Science and Technology, Ionospheric Prediction Service, Darlinghurst, New South Wales; New South Wales, University, Kensington, Australia)
ABS: It the A1 data collected at Camden, Australia, is sorted into 2 km virtual height bins, then the echo amplitudes are observed to increase as the overall virtual height decreases. Model calculations show that this behavior can be simulated by introducing a sharp increase in electron-density gradient near the reflection level. This appears to truncate the extent of the ray path in the deviative region and leads to a

fall in absorption. Since the deviative component of the absorption is a large fraction of the total absorption (at the observing frequency of 1.91 MHz), this reduction can be dramatic. This mechanism is believed to be the main source of variability in single-frequency A1 absorption measurements. 84/02/00 84A28358

UTTL: Characteristics of ionospheric ELF radiation generated by HF heating
AUTH: A/FERRARO, A. J.; B/LEE, H. S.; C/ALLSHOUSE, R.; D/CARROLL, K.; E/LUNNEN, R.; F/COLLINS, T. PAA: F/(Pennsylvania State University, University Park, PA)
ABS: This paper describes new observations of the characteristics of ELF generation produced by modulation of the dynamo current system from HF heating of the ionospheric D-region. A model of the ELF antenna structure embedded in the D-region is described and stepped ELF frequency observations are shown to support the model assumptions. Presented are data on the phase height of the ELF ionospheric antenna versus ELF frequency, polarization of the downgoing wave and relationship to the dynamo current direction, correlation of ELF field strength with per cent cross-modulation, power linearity tests and duty cycle results. All observations used the high power heater facility of the Arecibo Observatory. 84/10/00 85A15377

UTTL: A comparison of multipath arrival structures observed in the presence of a surface duct with predictions obtained using classical ray techniques and the parabolic equation method
AUTH: A/FREESE, H. A. CORP: Naval Underwater Systems Center, New London, Conn.
ABS: Observations of multipath arrival structures made at sea using short pulses have indicated that energy leaking into and out of surface ducts can play a significant role in the propagation of acoustic energy. This leakage phenomenon was first reported with reference to leakage into and out of refracted-surface reflected (RSR) paths. Recent experience has indicated that this leakage phenomenon can also be of importance for energy propagating via bottom bounce modes. Observed underwater acoustic multipath arrival structures are compared with theoretical predictions based on historic sound speed profiles for the vicinity of the Mid-Atlantic Ridge (3500 m water) and the Blake Plateau (1000 m water). Examining these comparisons for surface duct environmental conditions resulted in the conclusion that wave theory is necessary for predicting the distribution of received energy and that classical ray theory (i.e., ray propagation using the high frequency assumption without sound leakage correction) is inadequate.
RPT#: AD-A156024 AD-E900451 NUSC-TM-811061 85/02/14 85N35696

UTTL: Simultaneous observation of HF-enhanced plasma waves and HF-wave self-focusing
AUTH: A/FREY, A.; B/DUNCAN, L. M. PAA: A/(Max-Planck-Institut fuer Aeronomie, Katlenburg, West Germany); B/(Los Alamos National Laboratory, Los Alamos, NM)
ABS: Intense HF-radiowaves of the ordinary mode transmitted from the ground enhance plasma waves near the reflection height. These have been extensively studied in the past by the use of incoherent-scatter radars. Intense HF-radiowaves propagating in the ionosphere also produce electron density irregularities with scale sizes much larger than the HF wavelength of about 60 m. These have been observed by radio star intensity scintillations. For the past 2 years a new method was used at Arecibo, PR which allows radar and scintillation measurements at 430 MHz to be performed simultaneously along the same line of sight. The scale sizes deduced from the scintillation measurements are shorter than the scale sizes observed with the radar and are inconsistent with the HF-power density thresholds predicted by existing theories. 84/07/00 84A42013

UTTL: First experimental evidence of HF produced electron density irregularities in the polar ionosphere, diagnosed by UHF radio star scintillations
AUTH: A/FREY, A.; B/STUBBE, P.; C/KOPKA, H. PAA: C/(Max-Planck-Institut fuer Aeronomie, Katlenburg, West Germany)
ABS: Observations made on May 10 and September 9, 1983, during which large scale irregularities in the ionosphere were excited by the HF facility at Tromso, Norway, are discussed. The observations were detected using radio star scintillations at 933 MHz, and the HF-beam was pointed 10-15 deg south. Observations made on September 9 showed the HF-power density threshold to be 22 microW/sq m. It was determined that shorter scale irregularities appear to be preferentially produced during lower HF-power and are not the result of cascading from larger scale irregularities. Consequently, HF-power density thresholds for short scale irregularities do not concur with values proposed by Cragin et al. (1977). Also noted were the irregular growth times (about 10-40 sec) and decay

times (1-3 min). 84/05/00 84A34315

UTTL: The calibration of an HF radar used for ionospheric research
AUTH: A/FROM, W. R.; B/WHITEHEAD, J. D. PAA: B/(Queensland, University, Brisbane, Australia)
ABS: The HF radar on Bribie Island, Australia, uses crossed-fan beams produced by crossed linear transmitter and receiver arrays of 10 elements each to simulate a pencil beam. The beam points vertically when all the array elements are in phase, and is steerable by up to 20 deg off vertical at the central one of the three operating frequencies. Phase and gain changes within the transmitters and receivers are compensated for by an automatic system of adjustment. The 10 transmitting antennas are, as nearly as possible, physically identical as are the 10 receiving antennas. Antenna calibration using high flying aircraft or satellites is not possible. A method is described for using the ionospheric reflections to measure the polar diagram and also to correct for errors in the direction of pointing. 84/02/00 84A20788

UTTL: Sporadic E movement followed with a pencil beam high frequency radar
AUTH: A/FROM, W. R. PAA: A/(Queensland, University, Brisbane, Australia)
ABS: Several types of sporadic E are observed using the 5.80 and 3.84-MHz Bribie Island pencil-beam high-frequency radar. Blanketing Es takes the form of large flat sheets with ripples in them. Non-blanketing Es is observed to be small clouds that drift across the field of view (40 deg). There is continuous gradation of sporadic E structure between these extremes. There are at least four different physical means by which sporadic E clouds may apparently move. It is concluded that non-blanketing sporadic E consists of separate clouds which follow the movement of the constructive interference between internal gravity waves rather than being blown by the background wind. 83/12/00 84A16428

UTTL: Global considerations for utilization of real-time channel evaluation systems in HF spectrum management
AUTH: A/GOODMAN, J. M.; B/DAEHLER, M.; C/REILLY, M. H.; D/MARTIN, A. J. PAA: D/(U.S. Navy, E. O. Hulburt Center for Space Research, Washington, DC)
ABS: According to the CCIR, there are three classes of Real-Time Channel Evaluation (RTCE) systems. These classes are related to 1-remote transmitted signal preprocessing, 2-base transmitted signal preprocessing, and 3-remote received signal processing. Oblique Incidence Sounding (OIS) and Channel Evaluation and Calling (CHEC) are included in the first class. The present investigation is principally concerned with the attributes of the first class of RTCE. The OIS system is considered, taking into account the chirpsounder variety of OIS. However, the specific remarks may apply equally to all classes and subclasses of RTCE. It is pointed out that the chirpsounder system is currently of interest for application in quasi-real time spectrum management at HF because of a lower transmit power requirement (as compared to pulse OIS). 83/00/00 85A34480

UTTL: A commentray on the utilization of real-time channel evaluation systems in HF spectrum management
AUTH: A/GOODMAN, J. M.; B/DAEHLER, M.; C/REILLY, M. H.; D/MARTIN, A. J. CORP: Naval Research Lab., Washington, D. C.
ABS: Real-Time Channel Evaluation (RTCE) devices are coming into increased use in modern management of HF communications systems. The earliest techniques used in channel evaluation exploited vertical incidence pulse sounders (VIS) and these devices are still being used for some applications. An oblique incidence sounding technology has also been developed employing both pulse and chirp waveform approaches. A variety of RTCE devices are identified in the report but the emphasis is placed on the OIS chirp sounder. Using this device as a canonical channel evaluator, the implications for global, theater, and local HF resource management may be addressed. As one might suspect, there are both advantages and disadvantages which may accrue from construction of a sounder network. Issues include: data applicability, data collection and dissemination, network size and cost, network ECM vulnerabilities, optimum network architecture, and system component reliabilities, to name a few. These issues are outlined in the report.
RPT#: AD-A148616 NRL-MR-5454 84/11/28 85N17289

UTTL: An HF phased-array radar for studying small-scale structure in the high-latitude ionosphere
AUTH: A/GREENWALD, R. A.; B/BAKER, K. B.; C/HUTCHINS, R. A.; D/HANUISE, C. PAA: C/(Johns Hopkins University, Laurel, MD); D/(Toulon et Var, Universite, Toulon, France; Johns Hopkins University, Laurel, MD)
ABS: Since October 1983, a new coherent backscatter radar has been in operation at Goose Bay, Labrador, for the purpose of studying small-scale electron-density

structure in the high-latitude ionosphere. This radar operates over a frequency band that extends from 8 to 20 MHz, and it uses an electronically phased array of 16 log-periodic antennas for both transmission and reception. The radar transmits a seven-pulse pattern that permits the 17-lag complex autocorrelation functions of the backscattered signals to be determined as a function of range and azimuth. A complete description of the radar is provided, including explanations of the operation of the phasing matrix, the techniques of data acquisition and analysis as implemented in the radar micromputer, and the possible on-line and automatic operating modes that may be instituted. Examples of some of the initial results obtained with the radar during the afternoon and late evening hours are presented. These examples include images of the two-dimensional distribution of small-scale structure and of their associated mean Doppler motion and F-region Doppler spectra derived from the complex autocorrelation functions. These Doppler spectra show interesting differences from those of high-latitude E-region irregularities. 85/02/00 85A21132

UTTL: Initial studies of small-scale F region irregularities at very high latitudes
AUTH: A/GREENWALD, R. A.; B/BAKER, K. B.; C/VILLAIN, J. P. PAA: B/(Johns Hopkins University, Laurel, MD); C/(Toulon et Var, Universite, Toulon, France)
ABS: A recently developed HF radar system designed to study F region irregularities at high latitudes. The radar utilizes a seven-pulse multipulse sequence enabling it to determine 17-lag autocorrelation functions of E and F region irregularities. These autocorrelation functions and the associated Doppler spectra are not subject to the range-frequency ambiguity that occurs with other spectral techniques. The radar has been operated from the Cleary field site near Fairbanks, Alaska, for a short period in February 1982. The primary field of view of these measurements was poleward of Inuvik in northwestern Canada. Initial results from these measurements indicate a great deal of spatial and temporal variability in the irregularity Doppler characteristics; however, the overall pattern is generally consistent with the nominal pattern of very high latitude convection. Some anomalies existed, including evidence of an appreciable component of dawn-dusk drift near the most poleward latitudes of observation in the dawn local time sector. 83/12/00 84A13778

UTTL: High-frequency radiowave probing of the high-latitude ionosphere
AUTH: A/GREENWALD, R. A. PAA: A/(Johns Hopkins University, Laurel, MD)
ABS: The Applied Physics Laboratory program for probing the ionosphere using HF radars to study coherent scatter effects is described. The selection of appropriate frequencies permits the perpendiclar reflection of the signals from the ambient magnetic field. A radar station has been operating at Goose Bay, Canada since late 1983, and features an antenna array, phasing matrix and an electronics shed. One antenna array serves for transmission and two for reception. Switching electronics permit setting four patterns of time delay in the array, which enhances the signal phasing. The duplicate receiving array improves the capabilities of identification of the direction of the sources of received signals. Sample results are provided of measurements of the formation, dispersal and decay of F region irregularities. 85/03/00 85A31104

UTTL: Comparison of IONCAP predictions with AN/TRQ-35(V) oblique-incidence sounder measurements for a one-month Conus test during July and August 1982
AUTH: A/HARNISH, L. O.; B/HAGN, G. H.; C/ALVAREZ, W. T. PAA: B/(SRI International Telecommunications Sciences Center, Arlington, VA); C/(U.S. Army, Communications-Electronics Engineering Installation Agency, Fort Huachuca, AZ)
ABS: A one-month test of high-frequency (HF) propagation predictions was conducted by the U.S. Army Special Forces during July and August 1982. Several frequency selection techniques were employed, including one based on the frequency reliability tables (FRTs) generated by the U.S. Army Communications-Electronics Engineering Installation Agency (USACEEIA) using the Ionospheric Communications Analysis and Prediction (IONCAP) model. This paper presents the results of the analyses regarding the relationship between the frequencies predicted two months before the test using IONCAP and the frequencies obtained using the oblique-incidence sounders during the test. 83/00/00 85A34478

UTTL: Decision feedback equalizer test results for HF skywave communications. Design trade-offs and performance data are presented for Kalman and LMS-Decision feedback equalizers
AUTH: A/HOFF, L. E.; B/KING, A. R. CORP: Naval Ocean Systems Center, San Diego, Calif. CSS: ( Communications Systems and Technology Dept.)

RPT#: AD-A132651 NOSC-TD-471 82/09/30 84N72407

UTTL: HF modification of the auroral D-region detected by a partial reflection experiment
AUTH: A/HOLT, O.; B/BREKKE, A.; C/HANSEN, T.; D/KOPKA, H. ; E/STUBBE, P. PAA: C/(Nordly Observatoriet, Tromso, Norway); E/(Max-Planck-Institut fuer Aeronomie, Katlenburg-Lindau, West Germany)
ABS: During an auroral event with strong ionization in the lower D-region, the ionosphere was modified by a powerful (260 MW ERP) transmitter operating at 4 MHz in the extraordinary mode. The effects on the received partial reflection in an experiment using 2.75 HMz waves were observed. The results indicate a very abrupt increase in electron temperature with a factor of about 3 at the level of maximum heating. This seems to be followed by a change in electron density, probably due to electron temperature variation of recombination reaction rates. The effect of the heating wave on the irregularities giving rise to the partial reflections is discussed. 85/06/00 85A44514

UTTL: The problems of short-wave radio service in the region in which propagation modes are changing
AUTH: A/HORTENBACH, K. J. PAA: A/(Deutsche Welle, Cologne, West Germany)
ABS: The coverage area of a short-wave radio transmitter is considered under the assumption that interferences by other transmitters can be disregarded. It is found that this area is limited at large distances by propagation loss, while near the transmitter the skip zone has to be taken into account. However, observations of the received signal show that unsufficiently covered regions can also exist between these limits. Possible reasons for the occurrence of such regions are explored, taking into account the concept of the reliability of an ionospheric connection. If the operational frequency is high enough, zones of silence are found in the areas between the first and the second mode, and the second and the third mode. It is concluded that it is generally impossible to cover an area involving large distances without any gaps by using a single frequency. Such an objective requires an employment of additional frequencies. 84/00/00 84A26528

UTTL: Statistics at HF skywave ground backscatter Doppler spectra
AUTH: A/JONES, R. M.; B/RILEY, J. P.; C/GEORGES, T. M. CORP: National Oceanic and Atmospheric Administration, Boulder, Colo. CSS: (Wave Propagation Lab.)
ABS: Cumulative distribution functions (CDF's) are presented for Doppler spectra from skywave (ionospherically propagated) ground backscatter measured on 17 October 1980. The variance of the CDF's is much greater than would be expected if ionospheric reflection coefficient and ground backscatter coefficient were both Gaussian random variables. In contrast, skywave echoes from the sea show a normalized variance between 2 and 3, in agreement with a model that assigns a unit normalized variance to both the ionospheric reflection coefficient and the sea-echo backscatter. Furthermore, the CDF's for skywave sea-echo backscatter spectra show qualitative agreement with the Hankel distribution that would be expected if both ionospheric reflection coefficient and sea-echo backscatter were Gaussian random variables. The large variance in skywave ground backscatter is probably caused by observed trends of ground backscatter coefficient with range and azimuth. In addition, a new method is demonstrated for estimating the CDF of an ensemble that has only a few samples.
RPT#: PB84-204973 NOAA-TM-ERL-WPL-116 NOAA-84061103 84/05/00 84N32669

UTTL: Frequency correlation of HF skywave ground backscatter Doppler spectra
AUTH: A/JONES, R. M.; B/RILEY, J. P.; C/GEORGES, T. M. CORP: National Oceanic and Atmospheric Administration, Boulder, Colo. CSS: (Wave Propagation Lab.)
ABS: The frequency correlation of Doppler spectra from ionospherically propagated ground backscatter measured on 17 October 1980 is presented. As with direct sea-echo backscatter spectra, there is no significant frequency correlation other than that caused by the window (weighting function) used for the spectral analysis. This result supports a previous assumption used to analyze skywave sea-echo backscatter that ionospheric reflection does not introduce frequency correlation.
RPT#: PB84-204353 NOAA-TM-ERL-WPL-115 84/05/00 84N32668

all the alternatives considered. Life cycle costs were developed for four configuration alternatives, and ranged from $1.1M to $4.1M. The implementation strategy is also described, focusing on its four key phases: (1) evaluation and design; (2) procurement; (3) installation; and (4) operation.

RPT#: PB85-174167 DOT-SAP-85-1 85/01/00 85N29141

UTTL: Simplified estimation of ray-path mirroring height for HF radiowaves reflected from the ionospheric F-region

AUTH: A/LOCKWOOD, M. PAA: A/(Science and Engineering Research Council, Rutherford Appleton Laboratory, Didcot, Oxon, England) 84/04/00 84A28156

UTTL: Stability enhancement of a flexible robot manipulator

AUTH: A/LOOKE, T. D. CORP: Naval Academy, Annapolis, Md. CSS: (Dept. of Weapons and Systems Engineering.)

ABS: A computer software programming technique was developed to compensate a highly oscillatory robot system controlled by a bang-bang input. The assumptions that the system was linear and had lumped parameter characteristics allowed a fifth order, simplified dynamic model to be derived. Analysis using frequency response methods led to further simplification of the model to a third order system. Based on the third order model, a technique was developed which would compensate the system with a form of deadbeat control. Simulation of the model driven by the compensated bang-bang input verified the deadbeat response. The technique was implemented on an 8080-based microcomputer system which controlled the input. Actual system response to the compensated input was observed to be essentially free of the undesirable oscillatory motions, thus yielding an apparently rigid system.

RPT#: AD-A134185 USNA-TSPR-126 83/06/24 84N16807

UTTL: Standing wave pattern of HF radio waves in the ionospheric reflection region. I - General formulas

AUTH: A/LUNDBORG, B.; B/THIDE, B. PAA: B/(Uppsala, Ionospheric Observatory, Sweden)

ABS: General formulas for the accurate computing of the field strength near the reflection point of an electromagnetic wave impinging vertically upon a horizontally stratified ionosphere are derived by means of a uniform approximation valid throughout the whole reflection region. Smoothly varying profiles of general shape but with the gross features determined by the presence of one or two transition points are assumed. Coupling and nonlinear effects are not explicitly included. Since the uniform approximation method is analytic in its essence, it is in many respects more versatile than existing brute force numerical methods. 85/08/00 85A42024

UTTL: Standing wave pattern of HF radio waves in the ionospheric reflection region. Part 1: General formulas

AUTH: A/LUNDBORG, B.; B/THIDE, B. CORP: Uppsala Ionospheric Observatory (Sweden).

ABS: Formulas for computing the field strength near the reflection point of an electromagnetic wave impinging vertically upon a horizontally stratified ionosphere are derived by a uniform approximation valid throughout the whole reflection region. Smoothly varying profiles of general shape but with the gross features determined by the presence of one or two transition points are assumed. Coupling and nonlinear effects are not explicitly included. The formulas can be used to calculate the field structure of HF radio waves transmitted into the ionosphere and to calculate the swelling of reflected waves.

RPT#: UIO-SR-84-05-PT-1 ISSN-0349-2699 84/10/03 85N18253

UTTL: Detection of radiation from a heated and modulated equatorial electrojet current system

AUTH: A/LUNNEN, R. J.; B/LEE, H. S.; C/FERRARO, A. J.; D/COLLINS, T. W.; E/WOODMAN, R. F. PAA: D/(Pennsylvania State University, University Park, PA) ; E/(Instituto Geofisico del Peru, Lima, Peru)

ABS: The characteristics of the local signal that would be radiated from a strong equatorial electrojet when heated and modulated has been investigated. At the geomagnetic equator, the characteristics were similar to, but less intense than, those observed at Arecibo, Puerto Rico during experiments involving high frequency heating of the lower D-region of the ionosphere. This radiation is believed to be the first detected from a heated and modulated equatorial electrojet current in the Western Hemisphere. 84/09/13 85A10367

UTTL: HF satellite sound broadcasting - A preliminary assessment

AUTH: A/MILLER, J. E. PAA: A/(ORI, Inc., Landover, MD)

ABS: The characteristics of telescoping antennas and the necessary minimum field strength for acceptable consumer portable radio receivers linked to a satellite broadcast segment are detailed. Attention is

UTTL: Frequency dependence of anomalous absorption caused by high power radio waves
AUTH: A/JONES, T. B.; B/ROBINSON, T.; C/STUBBE, P.; D/KOPKA, H. PAA: B/(Leicester, University, Leicester, England); D/(Max-Planck-Institut fuer Aeronomie, Lindau ueber Northeim, West Germany)
ABS: High power radio waves can modify the ionospheric electron density distribution to produce field aligned plasma irregularities which give rise to the anomalous absorption of HF radio waves. The coefficient of anomalous absorption of a vertically propagated radio wave due to scattering from field aligned irregularities has been calculated, taking into account the effects of the geomagnetic field on electron motions. These results are compared with those of other theoretical models. Furthermore, the scale lengths of field aligned irregularities produced by a high power radio wave during recent high latitude modification experiments have been determined from measurements of the anomalous absorption by means of this theory. 84/02/00 84A28357

UTTL: Multiple phase-screen simulation of HF waves propagation in the turbulent stratified ionosphere
AUTH: A/KIANG, Y.-W.; B/LIU, C. H. PAA: A/(National Taiwan University, Taipei, Republic of China); B/(Illinois, University, Urbana, IL)
ABS: HF wave propagation in the turbulent stratified ionosphere is investigated using the multiple phase-screen simulation method. For waves incident along the axis of the background stratification, generalized parabolic equations for the modified complex amplitude u are derived which are applicable even in the strong scattering case. Based on these parabolic equations, several one-dimensional random phase screens are numerically generated to represent the effect of the random irregularities on the wave. For plane wave incidence, various statistical properties of the field, such as correlation, power spectrum, scintillation index, etc., can be directly computed using spatial averaging under the ergodicity assumption. The simulation method is also used to examine the effect of the velocity profile on the drift pattern on the ground. 85/06/00 85A36567

UTTL: Local time distribution of HF Doppler frequency deviations associated with storm sudden commencements
AUTH: A/KIKUCHI, T.; B/ISHIMINE, T.; C/SUGIUCHI, H. PAA: C/(Ministry of Posts and Telecommunications, Radio Research Laboratories, Koganei, Tokyo, Japan)
ABS: Consideration is given to the variations of the frequency of a standard HF signal, referred to as SCF, at the time of sudden storm commencements (SSC), based on the observations performed at Kokobunji and Okinawa, Japan, during the period of March 1981 to August 1983. It is shown that the nighttime main frequency deviations of the SCF are caused by the dawn-to-dusk electric field, which coincides with the electric field associated with the DP (polar-originated geomagnetic disturbances) part of the main impulse of SSC. The daytime preliminary frequency deviations of the SCF are explained in terms of the induction electric field associated with the abrupt magnetic increase due to SSC. 85/05/01 85A33521

UTTL: Global distribution of atmospheric radio noise derived from thunderstorm activity
AUTH: A/KOTAKI, M. PAA: A/(Ministry of Posts and Telecommunications, Radio Research Laboratories, Koganei, Tokyo, Japan)
ABS: Global atmospheric radio noise levels are calculated on the basis of global maps of thunderstorm activity derived from Ionosphere Sounding Satellite-b observations. Global distribution maps are presented for 2.5, 5.0, 10.0, and 20.0 Hz atmospheric radio noise frequencies, in Universal Time, for the autumn season. The results calculated exhibit better agreement with world-wide measurements than those given by the International Radio Consultative Committee Report No. 322. In addition, a better explanation of some atmospheric radio noise characteristics is obtained by the present method. 84/10/00 85A15378

UTTL: Skywave radar
AUTH: A/KOTAKI, M.
ABS: Some characteristics of skywave radars are explained and an outline of the most advanced radar system in the world is presented. Some examples of skywave radar applications are shown and the developement of radar systems around the world is described. 83/06/00 84A33821

UTTL: Field antenna handbook
AUTH: A/KUCH, J. A. PAA: A/(IIT Research Inst., Chicago)
CORP: Electromagnetic Compatibility Analysis Center, Annapolis, Md.
ABS: This handbook presents basic propagation theory, the fundamentals concerning antennas, and the design and use of tactical high frequency and very high frequency antennas. It is a field reference for basic antenna facts and a usage guide for antennas.

RPT#: AD-A155204 ECAC-CR-83-200 84/06/00 85N32242

UTTL: A theoretical model of artificial spread F echoes
AUTH: A/KUO, S. P.; B/KUO, S. C.; C/LEE, M. C. PAA: B/(New York, Polytechnic Institute, Farmingdale, NY); C/(MIT, Cambridge, MA) CORP: Polytechnic Inst. of New York, Farmingdale.; Massachusetts Inst. of Tech., Cambridge.
ABS: A theoretical model is developed for artificial spread F echoes elicited by irradiating the ionospheric F region with signals from ground-based HF transmitters. Account is taken of the irregularity polarizations, scale length and the magnetic dip angle of the echo. Ray tracing equations are defined for wave propagation in a horizontally stratified ionosphere which has been bathed with HF signals and therefore contains wavelike structures. Irregularities polarized within the meridian plane are found to cause the spread F echoes, while perpendicularly polarized irregularities do not. A magnetic dip angle of 5 deg must be exceeded for the spread F to become strong. The irregularities need scale lengths exceeding 100 m. Modes of operation are identified for inducing the required spread F echoes using ground-based radiotelescopes. 85/06/00 85A36558

UTTL: A laboratory investigation of the high-frequency Farley-Buneman instability
AUTH: A/KUSTOM, B.; B/DANGELO, N.; C/MERLINO, R. L. PAA: C/(Iowa, University, Iowa City, IA) CORP: Iowa Univ., Iowa City.
ABS: A laboratory investigation of the high-frequency Farley-Buneman instability is described. The instability was studied theoretically and is predicted to occur in the low E region of the ionosphere when the E/B drift velocity of the electrons relative to the ions is several times Cs, the ion-acoustic speed. In the experiments, an increase of the electric field well above the Lee threshold merely enhances the general power level of the fluctuations but does not effect appreciably their spectral shape. The observed frequency spectra fall-off in all cases very rapidly with increasing frequency, with a spectral shape of the type P(f) alpha 1f to the 3.5 power. This result has negative implications for a recently proposed mechanism of anomalous wave electron heating in the lower E region of the ionosphere. 85/02/01 85A23373

UTTL: Artificial ionospheric disturbances caused by powerful radio waves
AUTH: A/LEE, M. C.; B/KUO, S. P. PAA: B/(Polytechnic Inst. of New York, Farmingdale) CORP: Regis Coll., Weston, Mass. Presented at the Ionospheric Effect Symp., Alexandria, Va., 1-3 May 1984
ABS: Artificial ionospheric disturbances evidenced as fluctuations in plasma density and geomagnetic field can be caused by powerful radio waves with a broad frequency band ranging from a few KHz to several GHz. The filamentation instability of radio waves with a broad frequency band ranging from a few KHz to several GHz can produce both large-scale plasma density fluctuations and large-scale geomagnetic field fluctuations simultaneously. The excitation of this instability is examined in the VLF wave injection experiments, the envisioned MF ionospheric heating experiments, the HF ionospheric heating experiments and the conceptualized Solar Power Satellite project. Significant geomagnetic field fluctuations with magnitudes even comparable to those observed in magnetospheric (sub)systems can be excited in all of the cases investigated. Particle precipitation and airglow enhancement are expected to be the concomitant ionospheric effects associated with the wave-induced geomagnetic field fluctuations.
RPT#: AD-A148077 AFGL-TR-84-0297 84/11/19 85N17473

UTTL: Adaptive techniques to correct for effects of ionospheric refraction in navigation surveillance and communication systems
AUTH: A/LEE, M. C.; B/DONATELI, D. E. CORP: Regis Coll., Weston, Mass.
RPT#: AD-A128986 AFGL-TR-83-0051 83/02/03 84N74681

UTTL: US Coast Guard High Frequency Emergency Network (HFEN) implementation plan
AUTH: A/LENYEL, D.; B/KALEN, D.; C/WOLF, S. PAA: A/(Arinc Research, Inc.); B/(Arinc Research, Inc.) CORP: Coast Guard, Washington, D.C. CSS: (Office of Command, Control and Communications.)
ABS: In the event of national disasters or emergencies. The Coast Guard HFEN will provide minimum essential communications required to support critical command, control, and communications functions and operational command decisions without the use of telephones, teletypes, and data network systems. An overview of the U.S. Coast Guard's approach to development and implementation of the needed system is presented. It has selected an alternative for implementation that builds on existing HF facilities, which rates highest from a combined coast and performance perspective of

focused on the 26 MHz band and a minimum 26 dB SNR at the receiver for 90 percent of the time. A carrier modulated 30 percent by a 400 Hz tone is chosen. A 0.5-1 m whip antenna will have a 0.1 wavelength electrical length at 26 MHz, a situation which requires a minimum usable field strength of 48.8 dB for a linearly polarized wave or 51.8 dB for a circularly polarized wave. A GEO satellite will need an equivalent isotropic radiated power (EIRP) of 82.2 dB, i.e., twice the power of present GEO satellites. It is concluded that innovative or advanced designs will be needed if the necessary EIRP levels are to be met. Ongoing design efforts to meet the requirements are noted. 85/01/00 85A29621

UTTL: Comparison of HF oblique transmissions with ionospheric predictions
AUTH: A/MILLMAN, G. H.; B/SWANSON, R. W. PAA: B/(General Electric Co., Syracuse, NY)
ABS: High-frequency oblique transmissions made for a short period of time during the maximum phase of the past solar cycle are compared with the frequencies predicted utilizing the ITS-78 and the IONCAP (ionospheric communications analysis and prediction) ionospheric prediction programs developed by the Institute for Telecommunication Sciences. The external atmospheric noise levels predicted by the ITS-78 program are also compared with experimental data. The measurements conducted in the northeastern U.S. reveal that, for undisturbed solar-geophysical conditions, the experimental data are in good agreement with the theoretical predictions. 85/06/00 85A36532

UTTL: Experimental investigation of modal power distribution in a duct at high frequency
AUTH: A/MORFEY, C. L.; B/BAXTER, S. M. PAA: A/(Southampton, University, Southampton, England) 85/02/00 85A21853

UTTL: RF sky-wave backscatter radar for over-the-horizon detection
AUTH: A/NELSON, G. R.; B/MILLMAN, G. H. PAA: B/(General Electric Co., Syracuse, NY)
ABS: The environmental factors that impact the design and peformance of over-the-horizon backscatter (OTH-B) radars are reviewed in this paper along with the quantitative characteristics at higher frequency (HF) of targets, external noise, and deleterious propagation effects. The performance as limited by these environmental factors is also examined. Key design considerations in the synthesis of cost effective systems are reviewed. 82/00/00 84A10769

UTTL: Lunar tidal variation of ionospheric absorption on different frequencies at a tropical latitude during low and high solar activity years
AUTH: A/PATEL, D. B.; B/KOTADIA, K. M. PAA: B/(Gujarat University, Ahmedabad, India) 85/02/00 85A32997

UTTL: Kinematics of a stochastic field of internal waves in a shear current
AUTH: A/PETERS, H. CORP: Kiel Univ. (West Germany). CSS: (Ist. fuer Meereskunde.)
RPT#: DE83-901054 NP-3901054 81/00/00 84N71416

UTTL: Observation of high-frequency turbulence induced by an artificial ion beam in the ionosphere
AUTH: A/POTTELETTE, R.; B/ILLIANO, J. M.; C/BAUER, O. H.; D/TREUMANN, R. PAA: B/(CNRS and CNET, Centre de Recherches en Physique de l'Environnement Terrestre et Planetaire, Saint Maur-des-Fosses, Val-de-Marne, France); D/(Max-Planck-Institut fuer Physik und Astrophysik, Garching, West Germany)
ABS: Accurate measurements of some parameters of an artificial xenon ion beam propagating in an ionospheric plasma have been obtained by using a radiofrequency probe. The electron density profile inside the beam is studied as a function of the distance from the xenon source as it travels across the magnetic field lines. For distances small compared to the ion gyroradius, the electron density follows an analytical law which expresses conservation of flux of the number of ions through the beam cross section, proving that the beam is neutralized at very short distances from the source. At short distances from the source, the HF turbulence is concentrated at the front edge of the beam; with increasing distance, the density gradients are smoother and the turbulence weaker. Another type of HF turbulence generated by a destabilization of the harmonics of the electron gyrofrequency is detected in regions of low-density plasma preceding the beam edge. 84/04/01 84A30771

UTTL: Ionospheric true height profiles from oblique ionograms
AUTH: A/REILLY, M. H. PAA: A/(U.S. Navy, E. O. Hulburt Center for Space Research, Washington, DC)
ABS: An improved direct technique in which HF oblique ionograms are reduced to ionospheric true height profiles is introduced. The benefits of this method result principally from the use of a more accurate

Breit-Tuve relation to curved earth and ionosphere geometries. By comparing the results of calculations on known cases, the extent of improvement with this technique relative to the techniques by Gething and Maliphant (1967), George (1970), and Smith (1970), is demonstrated. 85/06/00 85A36528

UTTL: Ohmic heating of the polar F region by HF pulses
AUTH: A/SHOUCRI, M. M.; B/MORALES, G. J.; C/MAGGS, J. E. PAA: C/(California, University, Los Angeles, CA)
ABS: A study of the modifications produced by high-power, high-frequency (HF) waves in the F region of a quiet (low solar cycle) polar ionosphere is presented. In the polar geometry, maximum ohmic heating occurs producing significant changes to the zero-order electron temperature and density profiles. The temporal and spatial aspects of these changes are calculated, and their effects on relevant ionospheric parameters, such as the dc conductivity tensor, are found. An interesting result is the generation of a density depression at the reflection point of the HF waves. The dependence of ohmic heating on HF pump parameters and peak background density is considered. 84/05/01 84A34625

UTTL: The present state of continental and intercontinental ionospheric radio communications
AUTH: A/SILLENI, S.
ABS: Ionospheric characteristics which affect radio wave propagation are discussed, with an emphasis on techniques for using the ionosphere to extend the range of radio links. Interference has been proven highly correlated with the onset of solar disturbances. The ionosphere, however, by bending and reflecting radio signals, is useful for beyond-the-horizon radio links. ELF frequencies permit communication with submerged submarines. HF bands are employed for mobile surface communications around 75 MHz. The VLF bands from 3-30 kHz are applied for long-range links and radionavigation, and with controlled polarization furnish a high SNR. Continuity of broadcasts from earth-earth and earth-satellite depends on constant monitoring of ionospheric conditions and the spectra of the disturbances. 84/00/00 85A12527

UTTL: Limited vocabulary voice recognition and synthesis for HF communication
AUTH: A/SMITH, R. PAA: A/(USAF, Rome Air Development Center, Griffiss AFB, NY)
ABS: Since most military and commercial transmissions usually implement only a 256 word vocabulary for communications, then actual voice communications are not required, particularly in military jamming environments. The voice can be synthesized at the receiving end after transmission of a code sequence. A look-up table will suffice for synthesis of analog baseband signals. The system could also recognize the phonetic-letters-spelled-out (PLSO) system (e.g., Tango, Charley, Able, etc.). The technique permits HF utilization with an ionospheric hop and acceptable reception in all but very extreme jamming situations. It is noted that the simplified coding permits translation between operators who do not speak the same language. 83/00/00 85A14461

UTTL: Time and space domain filtering for improved HF communication
AUTH: A/SMITH, R. N.; B/MOSES, R. L. PAA: B/(Virginia Polytechnic Inst. and State Univ.) CORP: Rome Air Development Center, Griffiss AFB, N.Y. In AGARD Propagation Influences on Digital Transmission Systems: Probl. and Solu. 15 p (SEE N85-19269 10-32)
ABS: High speed data transfer has until recently been generally unattainable over HF channels because of the highly disturbed nature of the channel. The development of real-time signal processors has done much to mitigate effects of channel disturbances, making it possible to substantially increase throughput. Data rates as high as 9600 bits per second can be achieved by implementing signal processors as adaptive channel equalizers. This recent increase in data rates has rendered feasible HF spread spectrum communication systems. Spread spectrum systems are attractive in many applications because they are resistant to narrowband interference. However, channel equalizers necessary for high data rates are susceptible to narrowband interference. Spatial or temporal prefiltering has commonly been employed to prewhiten the received data, providing improved equalizer performance. The more popular one-dimensional techniques for temporal and spatial prewhitening of HF communication signals are reviewed. Also, combined space-time prewhitening techniques are proposed. Algorithms for designing and implementing these two-dimensional whitening filters are presented. Advantages and disadvantages of the two design strategies are discussed. Particular attention is focused on system performance, computational

requirements, and cost. Computer simulated results for these signal processing algorithms are presented.
84/10/00 85N19298

UTTL: Delta wings with shock-free cross flow
AUTH: A/SRITHARAN, S. S. CORP: National Aeronautics and Space Administration. Langley Research Center, Hampton, Va.
ABS: In order to have a high level of maneuverability, supersonic delta wings should have a cross flow that is free of embedded shock waves. The conical cross flow sonic surface differs from that of plane transonic flow in many aspects. Well-known properties such as the monotone law are not true for conical cross flow sonic surfaces. By using a local analysis of the cross flow sonic line, relevant conditions for smooth cross flow are obtained. A technique to artificially construct a smooth sonic surface and an efficient numerical method to calculate the flow field are used to obtain cones with smooth cross flow.
RPT#: NASA-CR-172297 NAS 1.26:172297 REPT-84-6 84/02/00
84N19282

UTTL: Stimulated electromagnetic emission - A new technique to study the parametric decay instability in the ionosphere
AUTH: A/STUBBE, P.; B/KOPKA, H.; C/THIDE, B.; D/DERBLOM, H. PAA: B/(Max-Planck-Institut fuer Aeronomie, Katlenburg, West Germany); D/(Uppsala, Ionospheric Observatory, Uppsala, Sweden)
ABS: A powerful HF wave, transmitted in the O mode with a frequency not exceeding the critical F region frequency, gives rise to secondary electromagnetic radiation, filling a frequency band of several 10 kHz around the frequency of the primary wave. The spectrum of these secondary waves is richly structured. The systematically occurring spectral features are identified and described. The majority of these features can be understood on the basis of scatter processes involving Langmuir waves and low-frequency density perturbations excited by the parametric decay instability. Other features, including a broad spectral maximum at 20 to 40 kHz on the upshifted side, are not as yet fully understood. 84/09/01
84A48272

UTTL: HF-enhanced ion and plasma line spectra with two pumps
AUTH: A/SULZER, M.; B/IERKIC, H. M.; C/FEJER, J. A.; D/SHOWEN, R. L. PAA: C/(Arecibo Observatory, Arecibo, PR); D/(SRI International, Menlo Park, CA)
ABS: Spectra of the HF-enhanced ion line and upshifted and downshifted plasma lines were obtained by a multiple-pulse technique (9 pulses) and by a technique using pulse-to-pulse correlation (512 pulses) with a short (about 1 ms) interpulse period for high frequency resolution. Observations using the multiple-pulse technique suggest the presence of two different types of spectral features. One type has a spectral width typically greater than 1 kHz, and another type a width much less than 400 Hz, the frequency resolution of the multiple-pulse technique used. Observations using pulses with a short interpulse period show the width of the narrow spectral features to be in the 20- to 60-Hz range. The results are in good qualitative agreement with predictions of the linear theory of parametric instabilities excited by two pumps. Observations of the narrow spectral features can be used to measure the line-of-sight component of ionospheric drift velocities of the electron gas with an accuracy of about 2 m/s. 84/08/01 84A44070

UTTL: HF-Doppler observations of acoustic waves excited by the Urakawa-Oki earthquake on 21 March 1982
AUTH: A/TANAKA, T.; B/ICHINOSE, T.; C/OKUZAWA, T.; D/SHIBATA, T.; E/SATO, Y.; F/NAGASAWA, C.; G/OGAWA, T. PAA: B/(Doshisha University, Kyoto, Japan); D/(University of Electro-Communications, Chofu, Tokyo, Japan); E/(Tohoku Institute of Technology, Sendai, Japan); F/(Tokyo Metropolitan University, Tokyo, Japan); G/(Kyoto University, Uji, Japan)
ABS: Ionospheric disturbances caused by the Urakawa-Oki earthquake at 0232 UT on 21 March 1982 have been detected by a network of HF-Doppler sounders in central Japan. The HF-Doppler data, together with the seismic data, have been used to formulate a mechanism whereby ionospheric disturbances are produced by an event of relatively small epicentral distance. Comparison of the dynamic spectra of these data has revealed experimentally that the atmosphere acts as a low-pass filter for the upward-propagating acoustic waves. The cut-off periods of this filter are estimated by applying a digital filter technique to the up-down component of the seismograms and are found to be 10 s from the ground up to 156 km and 20 s from 156-195 km. Considering the transfer function of the atmosphere derived from the theory of Pitteway and Hines, the observed result does not contradict the

prediction that the atmospheric filter mechanism is mainly attributable to viscosity. 84/03/00
84A30992

UTTL: The influence of large-scale TIDs on the bearings of geographically spaced HF transmissions
AUTH: A/TEDD, B. L.; B/JONES, T. B.; C/STRANGEWAYS, H. J. PAA: B/(Leicester, University, Leicester, England); C/(Sheffield, University, Sheffield, England)
ABS: An investigation is made of the adequacy of the simple corrugated reflector model to simulate the effects of large-scale travelling ionospheric disturbances (TIDs) on the bearings of HF transmissions. Model results are compared with the quasi-periodic variations of bearings measured simultaneously for a number of geographically spaced transmission paths and with results obtained from 3D ray tracing studies incorporating a more realistic TID model (the 'wave' model). Although the bearing error signature is, at some times (particularly during the day), similar to that predicted by the corrugated reflector model, ray tracing calculations with the 'wave' model give generally better results and predict bearing error signatures which correspond much better to some of those observed in the experimental data than those predicted by the simpler model. However, the prediction accuracy of both models is found to be limited by temporal variation of both the TID waveform and the ambient ionosphere. 84/02/00 84A28353

UTTL: HF prediction using adiabatic invariant theory
AUTH: A/TICHOVOLSKY, E. J. CORP: Rome Air Development Center, Griffiss AFB, N.Y.
ABS: Gurevich and Tsedilina's adiabatic invariant theory for the potential well model of ionospheric HF radiowave propagation is reviewed and applied to a specific round-the-world (RTW) propagation path. Propagation modes (ducting, ground-hop and chordal) are qualitatively deduced from charts of the adiabatic invariants I(F), I(FE), I(FE+E) and I(1) vs path length. Quantitative results are presented in the form of a simulated RTW ionogram between an elevated source and receiver. In support of RADC's orbiting receiver experiment, the robustness of global HF propagation is displayed in the form of synoptic maps of I(E), I(F), I(FE) and I(1), using the IONCAP medium model ionosphere. These maps show the abundance of various propagation modes on a world-wide basis.
RPT#: AD-A143239 AD-E750949 RADC-TR-84-59 84/03/00
84N32652

UTTL: Real-time update of two well-known models of the maximum usable frequency
AUTH: A/UFFELMAN, D. R. PAA: A/(SRI International Telecommunications Sciences Center, Arlington, VA)
ABS: Models of the high frequency (HF) radiowave propagation environment have been developed with the aim to provide system designers and frequency managers with the information needed to optimize HF systems to perform a required set of tasks. An important outstanding problem to be solved by the technical community is the accurate assessment and short-term prediction of the HF propagation environment. Urgently needed is a highly accurate short-term HF propagation predictive capabil,ity for both the quiescent and disturbed environment. A description is given of work which offers a potential solution to a significant part of the considered problem. The attempt is made to employ an update technique in the MUF (maximum usable frequency) computation of IONCAP (Ionospheric Communications Analysis and Prediction). 83/00/00
85A34479

UTTL: Serial data transmission with large output on the HF channel
AUTH: A/VANUFFELEN, J. P. CORP: Telecommunications Radioelectriques et Telephoniques, Le Plessis-Robinson (France). IN AGARD Propagation Influences on Digital Transmission Systems: Probl. and Solu. 8 p (SEE N85-19269 10-32)
ABS: A modem is described that uses series transmission and a self-adaptive equalizer for correcting propagation distortions. The modem permits data transmission to a maximum output of 2400 b/s as well as in a continued mode with brief messages of one-second normal duration. Characteristics of the modem and the signal processing applied are discussed. Laboratory measurements of modem performance are compared with results of tests during simulated multipath propagation or with frequency shifting. Still other tests carried out on a 500 km link are discussed. Results were analyzed on a computer and are presented in the form of long- and short-term statistics.
84/10/00 85N19305

UTTL: HF ray tracing at high latitudes using measured meridional electron density distributions
AUTH: A/VILLAIN, J. P.; B/GREENWALD, R. A.; C/VICKREY, J. F. PAA: A/(Toulon et du Var, Universite, Toulon, France); B/(Johns Hopkins Univesity, Laurel, MD); C/(SRI International Radio Physics Laboratory, Menlo Park, CA)
ABS: Through the development of incoherent scatter radar

facilities it is now possible to determine the ionospheric electron density distribution in both latitude and altitude. Measurements based on the use of such facilities may be included in appropriately modified versions of existing HF ray tracing programs to determine HF propagation paths in a realistic high-latitude ionosphere. A description is presented of the results of a simulation which has utilized this approach. The conducted analysis demonstrates the importance of HF propagation of sharp horizontal density ionization gradients present in the auroral ionosphere. 84/02/00 84A20782

UTTL: High-resolution probing of the HF ionospheric skywave channel - F2 layer results
AUTH: A/WAGNER, L. S.; B/GOLDSTEIN, J. A. PAA: B/(U.S. Navy, Naval Research Laboratory, Washington, DC)
ABS: A computer-controlled, coherent, coded-pulse, oblique sounder designed to probe the HF communication channel (2-30 MHz) to a bandwidth of 1 MHz, is described. The HF channel prober permits simultaneous investigation of the time and frequency characteristics of channel behavior and provides a view of channel propagation conditions. A series of measurements made, during selected periods in 1982 and 1983, on a short-baseline (126 km) mid-latitude path between an island off the coast of southern California and the mainland, are reported. To illustrate the unique capabilities of the instrument and the unusual effects that are observed with a high-resolution probing pulse, the response of the one-hop F2-layer sky wave mode to a 1-microsecond probing pulse is used. 85/06/00 85A36529

UTTL: Observations of an ionospheric perturbation arising from the Coalinga earthquake of May 2, 1983
AUTH: A/WOLCOTT, J. H.; B/SIMONS, D. J.; C/LEE, D. D.; D/NELSON, R. A. PAA: B/(Los Alamos National Laboratory, Los Alamos, NM); D/(SRI International, Menlo Park, CA)
ABS: An ionospheric perturbation that was produced by the Coalinga earthquake of May 2, 1983, was detected by a network of high-frequency radio links in northern California. The ionospheric refraction regions of all five HF propagation paths, at distances between 160 and 285 km (horizontal range) from the epicenter, were affected by a ground-motion-induced acoustic pulse that propagated to ionospheric heights. The acoustic pulse was produced by the earthquake-induced seismic waves rather than the vertical ground motion above the epicenter. These observations appear to be the first ionospheric disturbances to be reported this close to an earthquake epicenter. 84/08/01 84A44075

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