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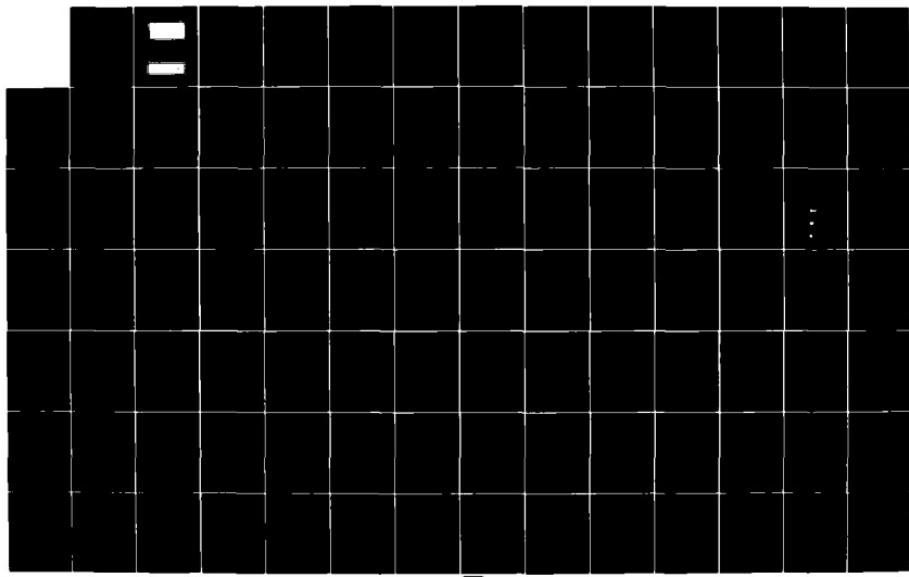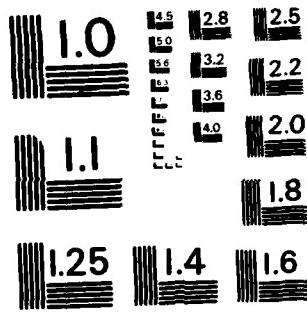
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ADVISORY GROUP FOR AEROSPACE RESEARCH & DEVELOPMENT

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### AGARD LECTURE SERIES No.127

Modern HF Communications

NORTH ATLANTIC TREATY ORGANIZATION

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NORTH ATLANTIC TREATY ORGANIZATION

ADVISORY GROUP FOR AEROSPACE RESEARCH AND DEVELOPMENT

(ORGANISATION DU TRAITE DE L'ATLANTIQUE NORD)

AGARD Lecture Series No.127

MODERN HF COMMUNICATIONS

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Exchanging of scientific and technical information;

Continuously stimulating advances in the aerospace sciences relevant to strengthening the common defence posture;

Improving the co-operation among member nations in aerospace research and development;

Providing scientific and technical advice and assistance to the North Atlantic Military Committee in the field of aerospace research and development;

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THEME

Lectures Series 127 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° 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 fore-
casting. Finally we hope to discuss the research and development efforts necessary to fully exploit HF
communications.

MILITARY SYSTEM REQUIREMENTS

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For the military in many nations, if not in all nations, HF turns out to be a primary means of communications rather than a backup. In some countries no other communications 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 us consider HF communications as a primary military communications medium even in those nations making use of satellites in their military complex.

The requirements of a military system range from repetitive 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.

In the area of architecture some military needs are as follows:

a. Connectiveness: The requirements may range from many users to many users to 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: This is always a difficult area but 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 margir.
d. Security
e. Interoperability
f. Coverage: ground and skyware
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 radiocommunications 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 buildings 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 operations 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 station such as aircraft or ships;
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 smaller ailine 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 a high quality and the performance is satifactory 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 higher 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.

HF SYSTEM DESIGN PRINCIPLES

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SUMMARY

The lecture deals with the general principles of HF communication system design, using as a framework a generalised communication system comprising:

- propagation path
- information source and sink
- source encoder/decoder
- channel encoder/decoder
- RF equipment.

The basic properties of the medium relevant to the design, control and operation of HF systems are considered. In particular, the problems of HF system control are examined in depth.

The lecture is intended to provide a link between the more detailed lectures concentrating upon specific aspects of HF system design.

1. INTRODUCTION

In this lecture, the general topic of HF system design will be considered from the viewpoints of present practice and future trends. The two major aims of the lecture are:

(i) To examine the fundamental technical problems of HF system design, control and operation together with potential solutions;

(ii) To provide an introduction to the more detailed aspects of HF system design to be examined in later lectures.

The key to the effective use of the HF propagation medium is an appreciation of its basic strengths and weaknesses so that a system design can exploit the strengths and minimise the effects of the weaknesses. Above all, a high level of performance requires that 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.

For convenience, the main strengths and weaknesses of the medium are summarised in Table 1 below:

Table 1

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

Table 1 (continued)

Weaknesses
(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.

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 later, 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. Two factors preventing this form of adaptation at the present time are:

> (i)   The slow reaction time of manually-controlled systems;
> (ii) The lack of an adequate real-time model for propagation and manmade interference.

Methods of overcoming these restrictions by improved HF system design are discussed in this paper.

2. HF SYSTEM CONTROL

The techniques necessary for the effective control of HF communication systems are essentially the same as those which have already been developed within the discipline of automatic control. Two communications scenarios which are encountered in practice are illustrated in Figs. 1(a) and 1(b). Fig. 1(a) shows a 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 make 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 two ways:

> (i)   By using a priori knowledge of the nature of the path, eg from previous practical experience and via off-line propagation analysis programs;
> (ii) By multiplexing channel probing signals with the traffic signals so that the receiver can derive information to model the path and hence can adjust its parameters accordingly.

In the closed loop system, the feedback link enables information extracted from the transmitted probing signals 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 process of probing, or identifying, the channel is known as real-time channel evaluation (RTCE) and is discussed in detail in a subsequent lecture (Darnell, 1983).

3. THE GENERALISED COMMUNICATION SYSTEM

Fig. 2 shows the elements of a generalised communication system. To illustrate the functions of the indvidual elements of Fig.2, an example of a digitized speech communication system will be considered.

The average information content of human speech is relatively low - comparable with that of a low-speed teleprinter link (50 - 100 bits/s). However, if PCM is applied to the speech waveform, typically a digitised rate of 64 kbits/s will be necessary for toll quality speech reproduction. The contrast between the 100 bits/s information rate and the 64 kbits/s PCM rate is marked and is due to the fact that natural speech reproduction requires the transmission of many more parameters than simply its information content, eg loudness, pitch, intonation, etc. It is possible

to digitise speech at a rate much less than 64 kbits/s and still to obtain acceptable quality by employing a more efficient model of speech production. PCM is a waveform encoding technique which digitises a 4 kHz wide speech bandwidth without regard to any speech generation mechanisms. Vocoders are systems which achieve lower digitisation rates by modelling the mechanisms of speech production more accurately.

3.1 Source Encoder Decoder

Much of the redundancy in speech is deterministic, ie its form is predictable; this allows its potential reduction or removal. A generalised source encoder has the function of reducing the redundancy in an information source signal, as shown in Fig. 3. Perfect source encoding would result in R_2 being zero and thus I would be a completely unpredictable random data stream which could not be compressed further. In practice, efficient source encoding makes $R_2 \ll R_1$.

A specific example of a source encoder for speech is a "formant" vocoder analyser. The formant model of speech depends upon the fact that a particular speaker will tend to have 4 or 5 spectral regions in which the energy of speech is concentrated. These high energy regions correspond physically to resonances of the vocal tract/cavity and a typical baseband speech spectrum might appear as illustrated in Fig. 4. The vocal tract cavity is therefore modelled in terms of the parameters of 4 or 5 resonant circuits. The centre frequencies and response amplitudes of the circuits will vary within defined ranges and the values of these parameters are sampled at regular intervals, digitised and formatted, together with digitised values of pitch and an indication as to whether the sound at the sampling instant is voiced or unvoiced. The overall data rate required to specify the digitised speech by means of the parameters of the formant model is normally in the range 1.2 - 2.4 kbits/s, thus representing a considerable reduction in comparison with the 64 kbit/s directly digitised PCM. Other vocoder models allow a similar reduction in rate to be achieved. Thus, some knowledge of the mechanism of speech production enables source encoding procedures to be developed which can remove much of the redundancy of speech. At the receiver site, the source decoder employs a formant synthesiser, based upon the same speech generation model, to which the decoded parameter values are applied to reintroduce the deterministic redundancy and hence reconstitute an estimate of the original source speech waveform. Fig. 5 shows the operation of the source decoder diagramatically.

3.2 Channel Encoder/Decoder

The function of the channel encoder in Fig. 2 is to add redundancy to the compressed source encoder output, as shown in Fig. 6. At first sight, the addition of more redundancy seems counter-productive since the source encoder has just been employed to reduce redundancy in the source signal. However, R_3 is now of a form specifically designed to combat the types of noise and perturbation likely to be encountered during transmission over the propagation path, eg error protection coding - as illustrated in Fig. 7. In the formant vocoder system under consideration, a compressed source rate of 1.2 kbits/s at the output of the vocoder analyser might typically be expanded to 2.4 kbits/s by the addition of parity check digits, thus allowing various transmission error patterns to be detected and in some cases corrected. A composite diagramatic representation of the processes of source and channel encoding, showing the relative compression and expansion of data rate at each stage, is given in Fig. 8. In general, at the receiver site, the channel decoder makes use of the redundancy inserted by the channel encoder in order to correct for the various types of transmission distortions and produce an estimate of the modified source signal, as shown in Fig.9. Clearly, the more that is known of the channel behaviour, the more effective can the channel encoding/decoding procedures be made - hence the requirement for channel identification or RTCE. The nature of the HF channel will be considered in more detail in the following section of the lecture.

3.3 RF Units

The function of the RF units of the generalised communication system is to provide the interface between the channel itself and the channel encoder and decoder. For most types of practical system, this interface will involve processes such as frequency conversion, filtering, amplification, etc. The rôle of the RF elements in the overall HF system design will be discussed later.

4. A SIMPLIFIED HF CHANNEL MODEL

In this section, a qualitative model of the HF skywave channel is described and used as a vehicle on which to base a discussion of the design of HF communication systems.

The problem of maximising the availability and reliability of an HF system is essentially involves the optimisation of a number of system parameters. These parameters, which will be classified as "primary" and "secondary", are listed in Table 2 below:

Table 2

Primary Parameters

(i)   Operating frequencies

(ii)   Transmission start time(s) and duration(s)

(iii)   Geographical location of the transmitter and receiver terminals - should this not already be determined by other considerations

Secondary Parameters

(i)   System control strategies

(ii)   Source encoding/decoding algorithms

(iii)   Channel encoding/decoding algorithms

(iv)   Radiated power levels

(v)   Antenna characteristics

It is seen that optimisation of the primary parameters will depend for the most part upon the state of the transmission medium at any time, whilst secondary parameter optimisation will also depend upon many other factors, eg equipment available, type of traffic required, etc.

The simplified channel model is based upon the primary parameters of frequency, time and position/distance, which form the axes of the diagram shown in Fig. 10. For a given HF system employing a specified set of secondary parameters, the channel can be considered as giving rise to a set of 3-dimensional transmission "windows" in which the signal can be placed in order to yield a specified level of fidelity at the receiver, eg in terms of say data error rate or speech intelligibility. The fidelity requirement can also be interpreted as a minimum signal-to-noise ratio (SNR) within the windows. Evidently, if the required SNR can be achieved by simple optimisation of primary parameters, then less attention need be given to the task of secondary parameter optimisation.

In Fig. 10, the times t_1, t_2, t_i,etc are the instants at which the windows become available to the system user; the values of t and the window durations Δt are normally dictated by factors such as fading characteristics, diurnal path variability, atmospheric noise, etc. In the frequency dimension, Δf represents the frequency range open for transmission over which the required SNR can be maintained; again, Δf is influenced by many factors such as time of day, season and sunspot number. Δd indicates the range of distance or position over which the transmitter and receiver locations can be varied, whilst again maintaining the specified received signal fidelity criterion. In practice, the available window structure would be considerably more complex than that shown in Fig. 10: multiple windows in both f and d dimensions could well exist for a given value of t - as illustrated in Fig. 11. Also, for the sake of simplicity, regularly-shaped windows are shown in Figs. 10 and 11; in reality, the windows would tend to be irregularly-shaped volumes.

As the required SNR is increased, the windows available for transmission will tend to reduce both in number and volume. A decrease in the required SNR would result in an increase in window dimensions causing many of the windows to merge together until, for very low SNR's (eg for manual morse telegraphy), they may well persist for many hours along the t axis. For transmission rates of a few kbits/s, the window durations may on occasions be only a few seconds, or even milliseconds.

4.1 Effects of Noise and Interference in the Model

The simplified channel model described previously can be extended to incorporate the effects of natural noise and manmade interference. In the same way as the communications transmitter/receiver system gives rise to transmission windows as shown in Fig. 11, the path from a noise source or interfering transmitter to the communications receiver will also have associated transmission windows.

The problem for the designer and operator of a communication system is thus to minimise the probability of coincidence between noise/interference and communication windows. If partial or complete coincidence does occur, there is a possibility that the communications traffic will be disrupted, ie the received SNR will be degraded. Fig. 12 shows a situation where partial coincidence between a communications window and an interfering signal window occurs, thus effectively reducing the volume of the window available to the communicator.

The remainder of this lecture will now be devoted to a discussion of HF system design methods which can

    (i) Maximise the size of the communications windows

and

    (ii) Minimise the probability of any coincidence between noise/interference and communications windows.

These design techniques will be discussed in the context of the primary and secondary parameters defined previously.

5. PARAMETER OPTIMISATION AND SYSTEM DESIGN

In the majority of existing HF systems, the extent to which any parameter optimisation can be carried out is severely limited by the basic system architecture. From the nature of the channel model presented in Section 4, it is obvious that the HF path will have an information carrying capacity which will vary with time over a wide range - a characteristic already noted in Section 1. This can be seen by considering the classical expression for the maximum error-free transmission capacity, C, for a channel having total bandwidth B Hz, signal power S and noise power N:

$$C = B \log_2 [1 + S/N] \quad \text{bits/s} \qquad (1)$$

All of the three parameters B, S and N as observed at the communications receiver will be functions of the three primary parameters of frequency, distance and time - assuming a fixed set of secondary parameters. Normally the position of the transmitter and receiver terminals will be determined by logistic or other factors, and therefore only the frequency and time variability need be considered. An exception to this simplification is a system employing geographical diversity with a number of spaced receiving stations where distance is an important parameter to be taken into account; the benefits of this form of diversity will be discussed later in this lecture. If distance variability is neglected, expression (1) can be modified to give the channel capacity at any instant as:

$$C = \int_B \log_2 [1 + S(f)/N(f)] \, df \quad \text{bits/s} \qquad (2)$$

and the total maximum error-free information throughput, I, in a time interval T is thus:

$$I = \int_T C \, dt = \int_T \int_B \log_2 [1 + S(f,t)/N(f,t)] \, d^f \, dt \quad \text{bits} \qquad (3)$$

Expression (3) is still approximate in that it represents the information throughput for a memoryless channel: in practice, HF channels may have memory in the form of multipath. However, expression (3) serves to illustrate the potential variability of the transmission capacity of an HF channel.

To clarify the concept of the variable capacity channel further, Fig. 13 illustrates one possible physical mechanism giving rise to such a variation: the magnitudes of two multipath component modes for the wanted signal are shown as functions of time and frequency, together with the amplitude of an unwanted interfering signal. The fading characteristics of the three signals are assumed independent in both time and frequency, to the extent that either of the wanted signal modes or the interfering signal may be dominant. When one wanted signal component has a significantly greater amplitude than either the other wanted signal component or the interfering signal, the possibility exists for relatively high quality, quasi single mode reception - albeit for a short time interval - as indicated by the "window" shown in Fig. 13.

It is interesting at this point to note a degree of similarity between HF channels and meteor burst paths (Bartholomé and Vogt, 1968). Communication using ionised meteor trails can take place only for short periods of time (typically 0.5 - 1 s) when a usable trail occurs; the time between the occurrence of such trails is relatively long - of the order of 30 - 60 s. However, when the path does exist high rate transmission is possible since the propagation is essentially single mode. The HF propagation model described previously also gives rise to high capacity windows, as shown in Fig. 13, but in contrast to the meteor burst path, having zero capacity between windows, the HF path has a residual, lower and variable capacity between windows.

The majority of existing HF systems tend to operate at a constant transmission rate with fixed bandwidths and signal formats - a situation which clearly forms a fundamental mismatch between the techniques employed and the essential nature of the propagation medium. For a constant transmission rate, R, if the range of available capacities is from C_{min} to C_{max} then, for error-free transmission to be possible

$$R \leqslant C_{min} \qquad (4)$$

Thus, for much of any given transmission interval, a constant rate HF system will be operating at well below its potential capacity. This simple argument leads logically to the concept of an adaptive system whose parameters can change in response to changes in available capacity. However, if constant rate transmission is mandatory, the source and channel encoding procedures must be made adaptive to counter changes in the received SNR. The practical design implications of adaptive operation are discussed in the following two sections dealing with primary and secondary parameter optimisation.

5.1 Primary Parameter Optimisation

In order to achieve a certain HF transmission capacity reliably, it is important to be able to monitor and select appropriate values of the primary parameters of frequency, time and distance/position. In this way, usable transmission windows can be selected and matched to the type of traffic which it is required to pass. Having optimised the process of primary parameter selection, the secondary parameters of the system can also be optimised - the topic considered in Section 5.2 of this lecture.

As has been indicated previously, the position and separation of the communication system transmitting and receiving terminals will normally be determined by factors other than the characteristics of the propagation path; thus, the problem of primary parameter optimisation 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 (Darnell, 1983). RTCE procedures have been developed for both the open loop and closed loop scenarios introduced in Section 2. Currently, 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 4, 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 must 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 channels. 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 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 into the communication system design and hence to employ the same basic RF and signal processing equipment. 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.

5.2 Secondary Parameter Optimisation

Assuming that RTCE has been employed to optimise the primary parameters of the HF system, design and control techniques for optimising the secondary parameters of the system will now be considered by reference to the elements of the generalised communication system introduced in Section 3 and shown in Fig. 2.

(a) System Control Strategies

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.

(b) Source Encoding/Decoding Algorithms

The function of the source encoder shown in Fig. 2 is to compress or modify the
basic source data, ie to reduce its redundancy. In the case of low rate telegraphy
data and analogue speech, the requirement for source encoding is minimal since these
forms of traffic can be transmitted with reasonable fidelity using relatively simple
channel encoding schemes. However, one of the major technical problems associated
with HF communication which has yet to be solved satisfactorily is that of
transmitting digitised 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 the majority of HF channels - 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. The latter option may not
be achievable in the shorter term without resorting to buffering of the input speech
and subsequent transmission at a lower rate over an extended period of time, thus
involving the system users in significant waiting times until the data has been
passed.

In general, one of the most effective contributions which the system user can
make to transmission reliability is to minimise the amount of traffic to be
transmitted over a given period by the communications system, which is in itself the
most fundamental form of source encoding. This then enables greater protection to be
applied to the residual data by more powerful channel encoding procedures.

(c) Channel Encoding/Decoding Algorithms

The channel encoder shown in Fig. 2 is intended to condition the compressed or
modified data from the source encoder so that it can withstand the noise and
perturbations likely to be encountered during transmission over the channel. The
channel encoding/decoding techniques which offer promise of economic and effective
implementation with variable capacity HF sytems will now be examined.

Diversity Processing

Diversity processing at HF can potentially provide a significant improvement in
the effective received SNR. Fig. 14 shows the theoretical improvement in the SNR in
dB for a dual diversity system in comparison with a non-diversity system as a
function of bit error rate (BER) under Rayleigh fading conditions with additive
Gaussian white noise (GWN). Clearly,in a practical dual diversity combining system, a
performance improvement of 10 dB in received SNR is achievable.

When the traffic signal is transmitted on more than one frequency
simultaneously, frequency diversity combining can be employed at the receiver
providing that the separation between frequencies is great enough to ensure that the
received components behave independently in terms of their fading characteristics
(Alnatt et al, 1957).

It is well established that, if the receiving antennas are appropriately
separated, space diversity can provide one of the most effective ways of enhancing HF
system performance. If space diversity is not feasible, polarisation diversity can in
some situations be used as a somewhat less efficient alternative.

Time diversity (McCarthy, 1975) and the related technique of interleaving
(Douglas & Hercus, 1971) have been shown to be effective in combatting burst errors
due to interference and fading. However, the improvement in performance is only
achieved at the expense of greater transmission bandwidth and increased processing
time.

If an HF communication system is structured appropriately, geographical diversity in which several versions of the transmitted signal are received at widely separated receiving sites can potentially provide one of the most effective counters to propagation fluctuations and interfering signals. The probability of interfering signal windows coinciding with communication windows for all links in a geographically dispersed network of the type shown in Fig. 15 is relatively small compared with the coincidence probability for any single link in the network. However, combination of the received signals to produce an overall maximum likelihood estimate of the transmitted signal does require that there are reliable interconnections between the various sites - a situation which will only apply in sophisticated networks where HF is one component of an integrated mix of communications techniques.

Signal Processing Techniques

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.

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.

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, as illustrated in Fig. 16. 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.

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).

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 (Darnell, 1983).

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. Fig. 17 shows the principle of this type of encoding for a transmitted signal comprising three identical data streams, each clocked at a different rate. At the end of every data block in the slowest stream, the receiver indicates to the transmitter which of the data streams it has been able to receive successfully in the previous transmission interval; the transmitter then re-aligns the transmitted data blocks accordingly over the subsequent transmission interval. This form of encoding has yet to be developed and implemented for HF applications.

With the advent of cheap and powerful computing capacity, it is now possible to consider the introduction of post reception processing a_ 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.

(d) 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.

(e) 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.

6. NETWORK AND FREQUENCY PLANNING CONSIDERATIONS

In this lecture, attention has so far been concentrated upon the design of single HF communication systems. However, when a network comprising many such systems is being considered, other important design criteria should be addressed. A network will typically comprise a mixture of groundwave and ionospherically refracted skywave paths. At the lower end of the frequency band, short-distance paths of the type used primarily for tactical communications will employ predominantly groundwave propagation or mixed skywave and groundwave propagation with consequent interference between groundwave and skywave - particularly at night. Longer distance point-to-point circuits will only employ skywave modes.

6.1 Frequency Sharing and Re-Use

From the viewpoint of spectrum utilisation, it is advantageous to be able to re-use or share frequency assignments between circuits wherever possible, ie to allocate the same operating frequencies for simultaneous use by two or more independent links. This requires that there is negligible interference between the links. Considering the schematic arrangement shown in Fig. 18(a) where two short-distance HF links, AA' and BB', are separated by a long HF path A"B", A" and B" being the mid-points of the two short-distance links. Fig. 18(b) shows typical plots of lowest usable frequency (LUF) and maximum usable frequency (MUF) for the three paths. The usable frequency ranges between the LUF and MUF for each of the three paths are shown hatched. If it assumed that the two short-range paths will employ groundwave propagation, then the frequency chosen can be the same providing that it does not intersect the usable skywave range for the path A"B"; the distance A"B" is too great for any groundwave propagated interference to occur between AA' and BB'. In Fig. 18(b), two possible groundwave frequencies, $f_{GW(L)}$ and $f_{GW(U)}$, are shown; the lower frequency is chosen to be less than the LUF's for the two short-range paths to avoid the possibility of interference between groundwave and skywave. The upper frequency is chosen to be above the MUF for the path A"B" to eliminate skywave interference between the two circuits. It should be noted that the upper frequency will normally only be usable over very short distances because of the poorer groundwave propagation at higher frequencies; this poorer propagation is to some extent offset by lower natural noise levels at higher frequencies.

If the two short-range paths are of such a length and nature that only skywave propagation is possible, the individual skywave signals using a common frequency should again not interfere with each other via path A"B". It is usual, wherever possible, to operate skywave circuits on a "2-frequency" basis, ie one frequency for night-time operation and another for day-time. Fig. 18(b) also shows how this type of operation can be arranged to fall within the usable ranges for the two short paths and yet below the LUF for path A"B".

For any given geographical situation involving tactical short-range links separated by relatively long distances, the above technique may be applicable. This example serves only to illustrate the potential value and complexity of HF frequency sharing. If the terminals of the tactical circuits are mobile, their predicted operational areas must be taken into account in any frequency sharing algorithms, together with any other regions to which it is desirable to restrict propagation. The frequency management of a network of automatic and adaptive HF communication links, each having integrated RTCE, will be a significant practical problem. H wever, the spectrum congestion in the HF band is such that the frequency management of individual networks and links must be carried out in a more co-ordinated manner, and must be considered at the design stage of any new system.

6.2 Frequency Band Extension

Returning again to the simplified model for HF propagation described in Section 4: a similar representation of propagation by the mechanisms of sporadic E-layer refraction and meteor burst is also valid, with the difference that in these two cases the coincidence probability between communication and interference windows is much less. RTCE can also enable the selection of sporadic E modes, which are

relatively localised phenomena, thus tending to employ frequencies which are somewhat higher on average than for more conventional HF modes. In this situation, interference probabilities will be reduced significantly.

Meteor burst systems have the advantage that the path between communications transmitter and receiver is virtually unique with an extremely low interference probability. This type of link, if integrated into the HF system, could be used for EOW purposes. Thus, the concept evolves of an HF type of communication system, with an operating frequency range of say from 2 - 50 MHz, which would make use of any convenient propagation mechanism available at the required transmission time.

7. CONCLUSIONS

From this introductory lecture, covering current HF communication system design and future design trends, certain general conclusions can be drawn:

(a) The key to significant improvement in HF system performance is the effective incorporation and application of appropriate RTCE techniques;

(b) That the design aim for an HF system should be to achieve the desired level of performance by the application of improved system control and signal generation/processing algorithms, rather than by the use of traditional transmission techniques, high radiated power levels and large antennas;

(c) That the variable capacity of the HF path should be recognised and exploited.

Against this background, the major technical problems to be solved in future HF designs can be identified as:

- The development of RTCE techniques which can be integrated with the communications function of the system, employing a common range of RF and processing equipment;

- The development of automatic system control algorithms and associated processing equipment;

- The design of adaptive/re-configurable signal generation and processing equipment for use in the variable capacity environment;

- The provision of highly-reliable EOW facilities for system control purposes;

- The development of techniques to allow the transmission of reliable digitised speech over all types of HF paths;

- Evaluation of the benefits of designing HF equipment to cover an extended frequency range above 30 MHz to exploit more transient propagation mechanisms such as meteor burst;

- Design techniques to minimise system operation and maintenance costs.

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a) OPEN LOOP

INPUT x(t) — TX — PROPAGATION PATH — RX — OUTPUT y(t)

b) CLOSED LOOP

INPUT x(t) — TX — PROPAGATION PATH — RX — OUTPUT y(t)

FEEDBACK PATH
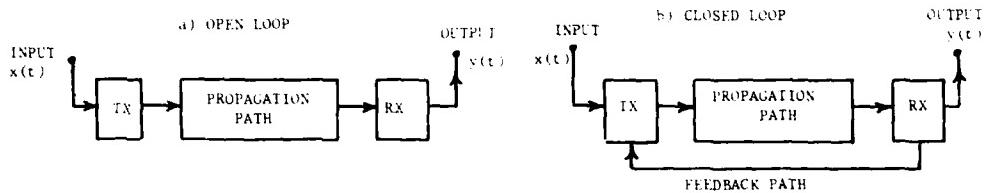
FIGURE 1: BASIC COMMUNICATION SCENARIOS

TRANSMITTER SITE

RECEIVER SITE

NOISE PERTURBATIONS

SOURCE — SOURCE ENCODER — CHANNEL ENCODER — RF UNITS — CHANNEL — RF UNITS — CHANNEL DECODER — SOURCE DECODER — SINK

SYSTEM CONTROL
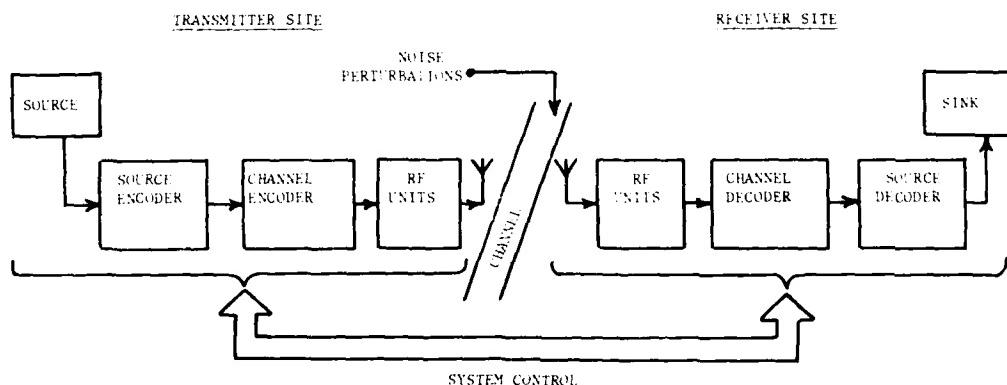
FIGURE 2: GENERALISED COMMUNICATION SYSTEM

INFORMATION & REDUNDANCY $(I + R_1)$

(SOURCE SIGNAL)

SOURCE ENCODER

INFORMATION & REDUNDANCY $(I + R_2)$

(MODIFIED/ COMPRESSED SOURCE SIGNAL)

$R_2 < R_1$
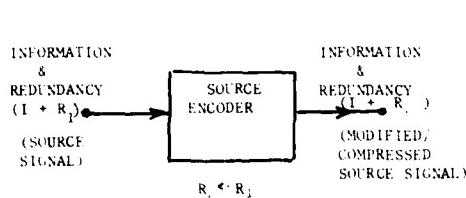
FIGURE 3: FUNCTION OF SOURCE ENCODER

AMPLITUDE

f_1 f_2 f_3 f_4 FREQUENCY

FORMANT FREQUENCIES
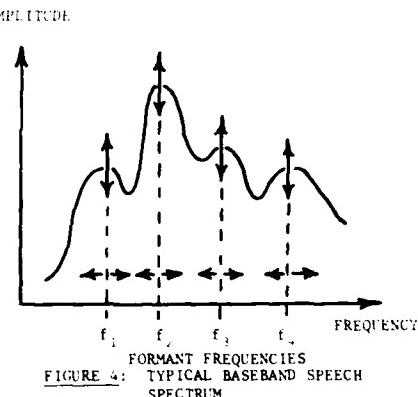
FIGURE 4: TYPICAL BASEBAND SPEECH SPECTRUM

MODIFIED/ COMPRESSED SOURCE SIGNAL ESTIMATE $(I + R_2)$

SOURCE DECODER

SOURCE SIGNAL ESTIMATE $(I + R_1)$
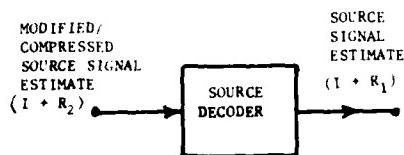
FIGURE 5: FUNCTION OF SOURCE DECODER

INFORMATION & REDUNDANCY $(I + R_2)$

(MODIFIED/ COMPRESSED SOURCE SIGNAL)

CHANNEL ENCODER

INFORMATION & REDUNDANCY $(I + R_3)$

(TRANSMITTED SIGNAL)
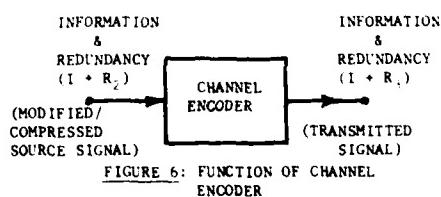
FIGURE 6: FUNCTION OF CHANNEL ENCODER

$(I + R_i)$

MODIFIED/ COMPRESSED SOURCE SIGNAL DIGITS

ERROR
PROTECTION
CODING
ALGORITHM

$(I + R_i)$

MODIFIED COMPRESSED SOURCE SIGNAL DIGITS	PARITY DIGITS
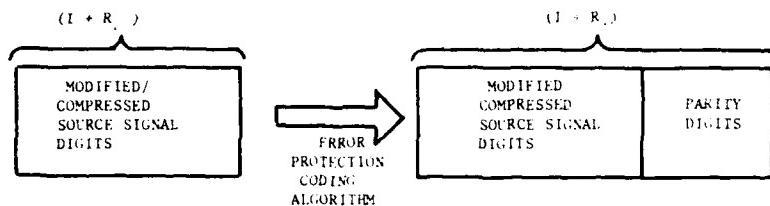
FIGURE 7:   ERROR PROTECTION CODING AS AN EXAMPLE OF CHANNEL ENCODING

SOURCE	SOURCE ENCODING PROCESS	COMPRESSED/ MODIFIED SOURCE SIGNAL	CHANNEL ENCODING PROCESS	CHANNEL ENCODED SIGNAL

DATA
RATE

I

R

I

R

I

R

DIRECTION OF DATA FLOW THROUGH
SYSTEM
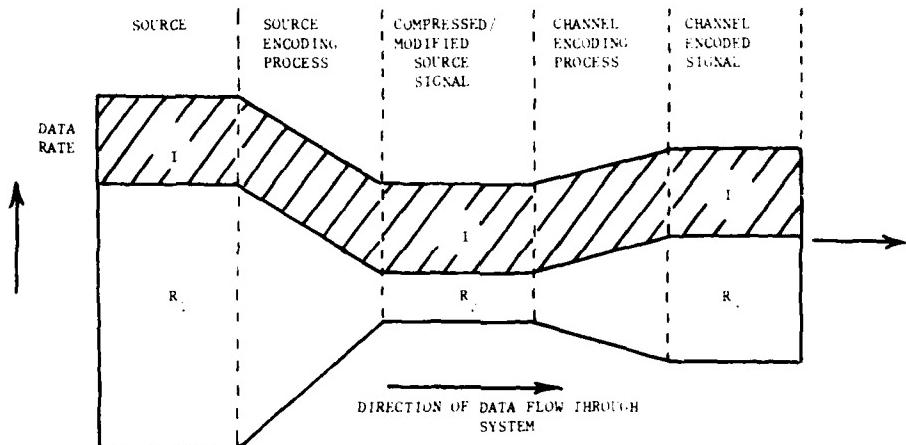
FIGURE 8: COMBINED EFFECTS OF SOURCE & CHANNEL CODING

CHANNEL
ENCODED SIGNAL
ESTIMATE

$(I + R_i)$

CHANNEL DECODER

MODIFIED/
COMPRESSED
SOURCE SIGNAL
ESTIMATE

$(I + R_i)$
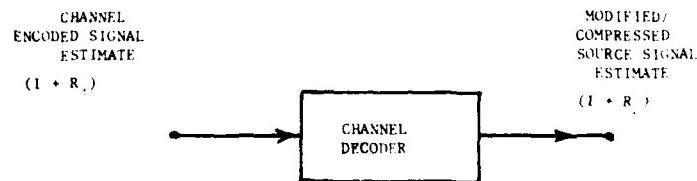
FIGURE 9: FUNCTION OF CHANNEL DECODER
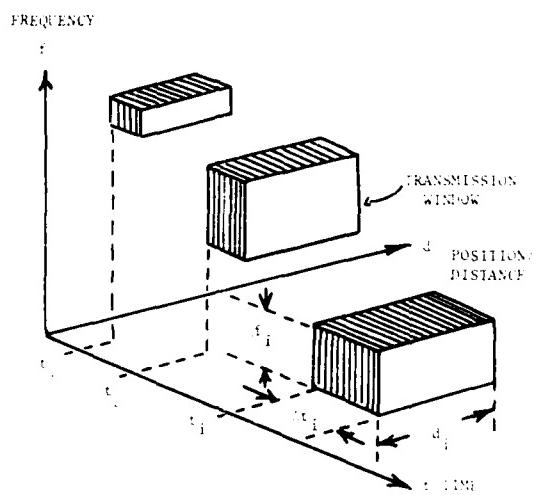
FREQUENCY

FIGURE 10: SIMPLIFIED HF PATH MODEL
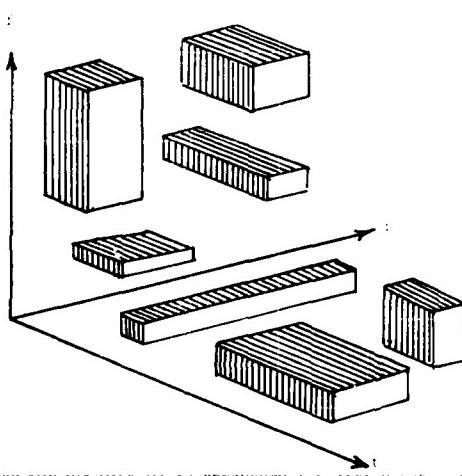
FIGURE 11: SIMULTANEOUS WINDOWS IN FREQUENCY & POSITION FOR A GIVEN TIME
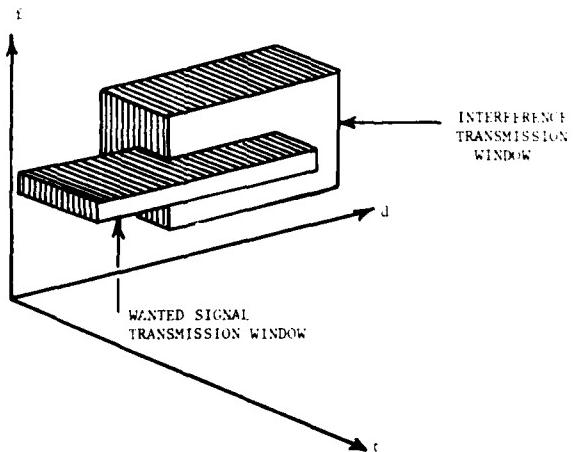
FIGURE 12: EFFECT OF INTERFERENCE IN THE SIMPLIFIED HF PATH MODEL

4-14

AMPLITUDE

FREQUENCY

WINDOW

f_1

f_2

f_3

TIME

——— MODE 1   —·—·—· MODE 2

- - - - - INTERFERING SIGNAL
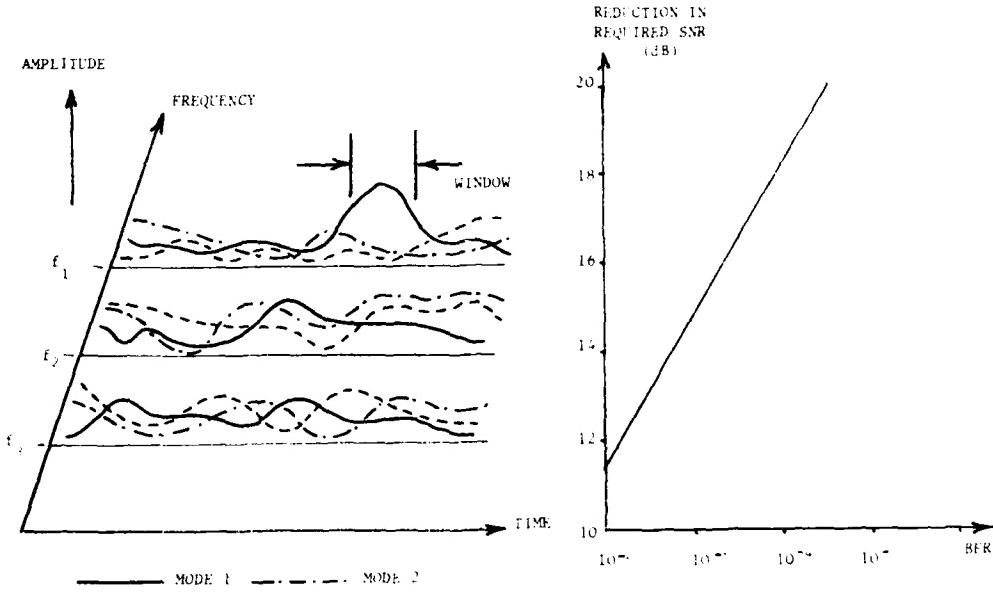
FIGURE 13: THE NATURE OF A VARIABLE CAPACITY
HF PATH

REDUCTION IN
REQUIRED SNR
(dB)

20

18

16

14

12

10

10^{-2} 10^{-3} 10^{-4} 10^{-5} BER

FIGURE 14: REDUCTION IN MEDIAN SNR REQUIRED
FOR A GIVEN BER RELATIVE TO
A NON-DIVERSITY SYSTEM,
ASSUMING IDEAL DUAL-DIVERSITY
COMBINING

COMMUNICATIONS
TX

RX1

PATH 1

PATH 2

RX2

PATH i

PATH 3

INTERFERING
TX

RX3

RXi
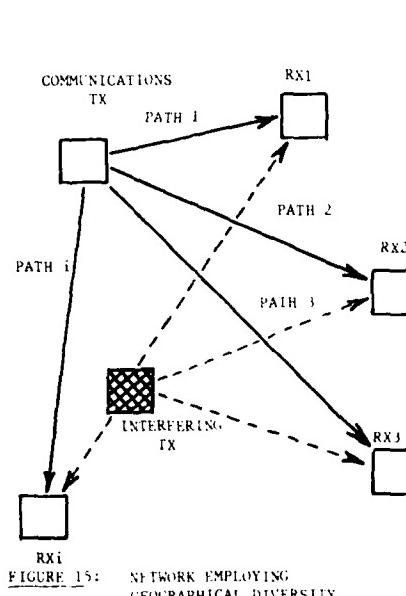
FIGURE 15: NETWORK EMPLOYING
GEOGRAPHICAL DIVERSITY

INTERVAL OVER
WHICH SYMBOL
IS PRESENT

NON-QUENCHED

FILTER
INITIALISED
BY NOISE

DETECTION
DECISION

INPUT
SIGNAL
+
NOISE

QUENCHED

GATE
ON

GATE
OFF

SYMBOL
TIMING

TIME
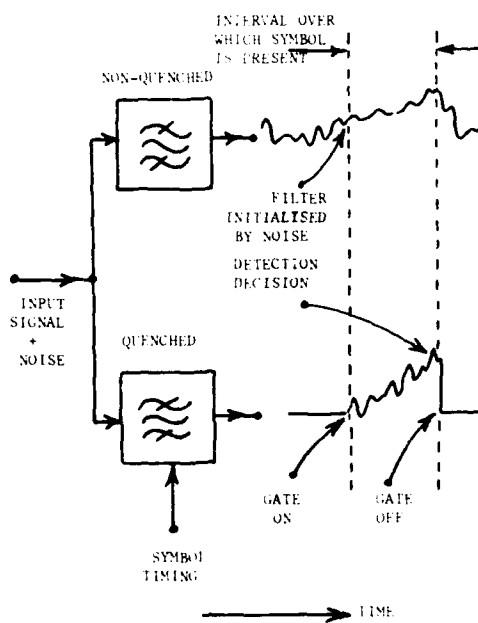
FIGURE 16: PRINCIPLE OF QUENCHED
FILTERING

STREAM 1

STREAM 2

STREAM 3

RECEIVER
RESPONSE

ACCEPT
STREAM

ACCEPT
STREAM

ACCEPT
NONE

ACCEPT
STREAM

ACCEPT
STREAM

ACCEPT
STREAM

TIME
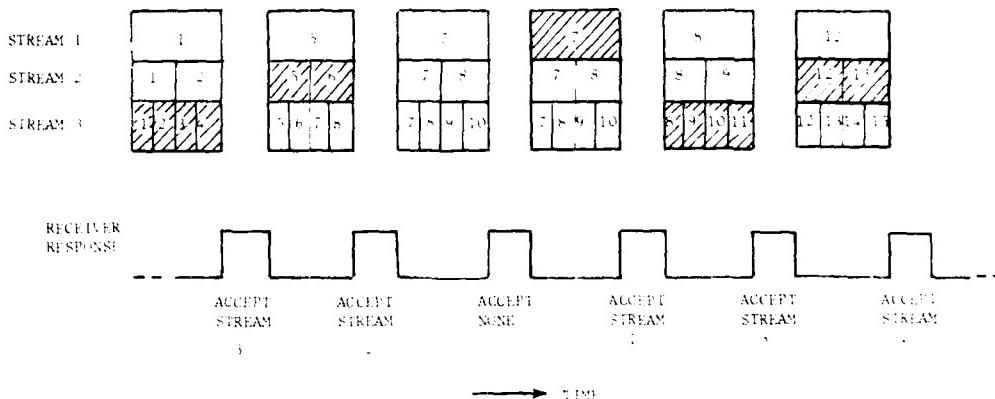
FIGURE 17: EMBEDDED DATA ENCODING SCHEME

(a)

A'

A''

SHORT
PATH

A

LONG
PATH
A''-AA'
OR
BB'

(b)

FREQUENCY
(MHz)

16

B   B''   B'

SHORT
PATH

NIGHT   DAY   NIGHT

f_{GW}

0   6   12   18   24 TIME
(HOURS)
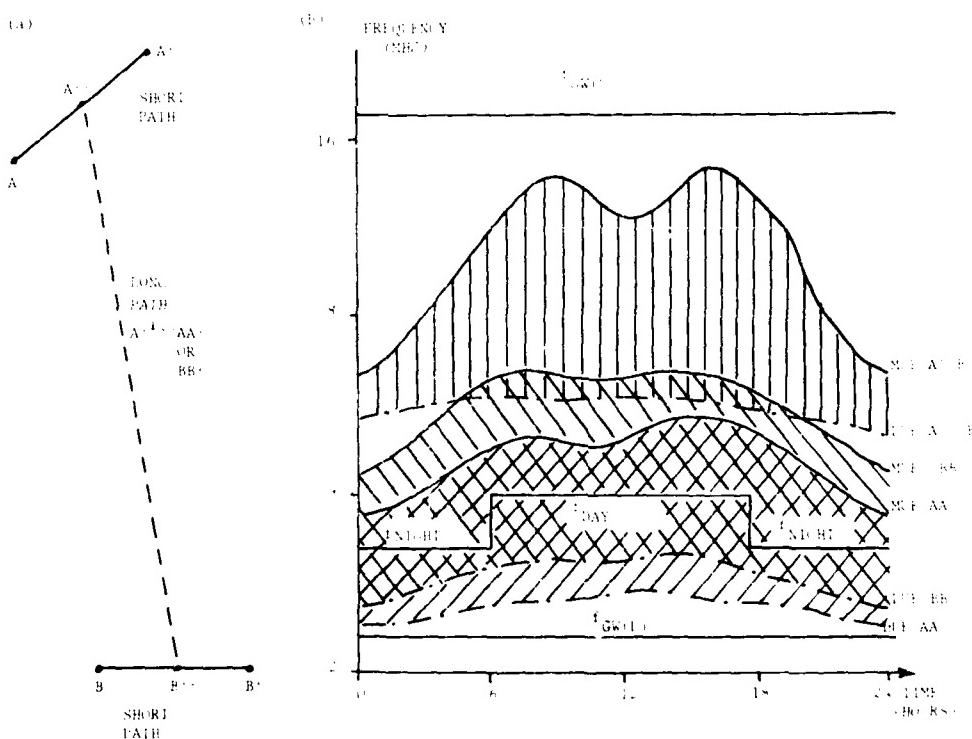
FIGURE 18: PRINCIPLE OF FREQUENCY SHARING

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:

$$\frac{dN_e}{dt} = q(h,\chi) - \alpha_{eff} N_e N_+ \qquad (1)$$

where N_+ is the positive ion density and α_{eff} is a constant, the effective recombination rate, χ 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 ν with which an electron collides with neutral molecules and ions. The collision frequency ν is proportional to atmospheric pressure. Typical height variations of N_e, ν 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 μ is the real part of the refractive index and I_n is the angle of incidence upon the layer labelled $n+1$. Then

$$\mu_0 \sin I_0 = \mu_1 \sin I_1 = \mu_2 \sin I_2 ... = \text{constant} \qquad (2)$$

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

$$\mu = \sqrt{1- \frac{N_e e^2}{m_e f^2 4\pi^2 \epsilon_0}} \qquad (3)$$

where N_e is the number density of free electrons, m_e the electron mass, f the wave frequency and ϵ_0 the permittivity of free space. The plasma frequency of the propagation medium is defined as

$$f_N^2 = \frac{N_e e^2}{m_e 4\pi \epsilon_0} \qquad \text{so that} \qquad \mu = \sqrt{1- \frac{f_N^2}{f^2}}$$

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>

$$f_0 = 9\sqrt{N_{e_{max}}} \qquad (5)$$

where $N_{e_{max}}$ is the maximum electron density. Frequencies $f > f_0$ will penetrate the layer. A wave incident obliquely at an angle I_0 will be reflected at a height where

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

and the highest frequency which can be reflected is $f_{max} = 9\sqrt{N_{e_{max}}} / \cos I_0 = f_0 / \cos I_0$. This frequency is, of course, always greater than the critical frequency f_0.

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_0 is reflected back to the ground at a distance d. The maximum frequency which can reach this distance is $f_{max} = f_0 / \cos I_0$. If there are no denser ionospheric layers at greater heights, this frequency is the Maximum Useable Frequency (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_0. 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 can be derived from simple considerations, if the critical frequency of the layer at the reflection is known.

Figure 6 shows the results of raytracing for three frequencies in a model layer with critical frequency $f_0 = 4$ MHz and $N_{e_{max}}$ at a height of 300 km. The figure clearly demonstrates that for a frequency $f > f_0$ 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

$$A \propto \frac{1}{f^2} \int_s \frac{1}{\mu} \, N_e \nu ds \qquad db$$

where ν is the collision frequency of an electron, μ 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 μ 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 $f \mp f_L$ 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

$$n^2 = \epsilon - i \frac{1800}{f} \sigma$$

where ϵ is the permitivity, f is given in MHz and σ the conductivity in Siemens m^{-1}, ϵ and σ 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 (1925) 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 pulse' 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. There 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 χ, 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 specific 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)

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

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

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

where $A(f)$ is a constant when $\mu \rightarrow 1$ (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

$$L(f) = \frac{677.2 \ I \ sec \ \phi_0}{(f+f_L)^{1.98}+10.2} \qquad \{dB\} \qquad 11)$$

where ϕ_0 is the angle of incidence at 100 km, and the frequencies are given in MHz

$$I = -0.4+e^{-2.937+0.8445 \ f_o f}$$

The factor I thus contains the solar zenith angle and sunspot cycle dependence through the E-layer critical frequency $f_o F$. 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 F-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:

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

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

$$P_r = P_t + G_R - L_b \qquad (13)$$

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 $1 \ \mu V \ m^{-1}$) is given in terms of P_r by

$$E = P_r + 20 \ \log f + 107.2 \qquad (14)$$

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

$$E_n = F_a - 65.5 + 20 \ \log f + 10 \ \log b \qquad (15)$$

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, YLF 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 F layer, FS 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 YLF 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
ITT 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 Cyber 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°.

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 78.05

		MAR	1983			SSN =	77.					
OSLO TO LONDON							AZIMUTHS			N. MI.		KM
59.92 N	10.75 E - 51.50 N				.08 W	229.52	51.54	622.6	1153.0

MINIMUM ANGLE   5.0 DEGREES

ITS- 1 ANTENNA PACKAGE

XMTR   2.0 TO 30.0 HORZ. DIPOLE H   -.25 L   -.50 A   0.0 OFF AZ   1.0
RCVR   2.0 TO 30.0 HORZ. DIPOLE H   -.25 L   -.50 A   0.0 OFF AZ   0.0
POWER =   1.000 KW S 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	.02	.00	.00	.00	.00	.00	.00	.78
4.0	4.7	5.5	.15	.94	.87	.21	.00	.00	.00	.00	.10	.00	.00	.80
6.0	6.7	7.5	.52	.91	.72	.73	.32	.00	.00	.00	.00	.00	.00	.79
8.0	8.7	11.1	.00	.01	.75	.85	.35	.61	.23	.01	.00	.00	.00	.81
10.0	10.7	15.4	.00	.10	.52	.73	.91	.36	.69	.44	.14	.00	.00	.84
12.0	12.7	14.4	.00	.00	.15	.79	.85	.84	.73	.58	.50	.01	.00	.75
14.0	14.7	14.5	.01	.00	.51	.79	.92	.52	.70	.45	.11	.00	.00	.81
16.0	16.7	14.7	.00	.15	.71	.79	.74	.74	.67	.58	.27	.00	.00	.82
18.0	18.7	12.5	.48	.76	.85	.81	.75	.70	.59	.52	.12	.00	.00	.79
20.0	20.7	9.7	.25	.97	.95	.47	.74	.49	.16	.01	.00	.00	.00	.88
22.0	22.7	7.1	.07	.98	.14	.85	.57	.22	.05	.00	.00	.00	.00	.85
24.0	.7	6.5	.05	.97	.95	.77	.45	.16	.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 5°. Horizontal half wave dipoles are specified for both transmitter and receiver with heights above ground 1/4 wavelength. (H = 0.25 λ, L = 0.5 λ). 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 78.05

MAR   1985   SSN = 77.
OSLO TO LONDON   AZIMUTHS   N. MI.   KM
59.92 N   10.75 E - 51.50 N   . 8 W   227.52   51.54   622.6   1155.0
                          MINIMUM ANGLE   5.0 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 CBW   REQ. REL = .90 REQ. SNR = 55.0

UT MUF

12.0 14.4   2.0   5.0   5.0   7.5 10.0 12.5 15.0 17.5 20.0 25.0 30.0 FREQ
     1F2   1 5   1 E   2F1 1F2   1F2   1F2   1F2   1F2   1F2   1F2   1F2   MODE
     29.7   5.4   6.0 49.4 51.4 25.0 25.3 29.7 29.7 29.7 29.7 29.7 ANGLE
     575.   82. 88. 262. 411. 600. 273. 575. 575. 575. 575. 575. V HITE
     69. -56. -21. 45.   55.   58.   69.   68.   55.   44.   5. -15. SNR
     12. 115.   48.   77.   5.   5.   4.   18.   22.   52.   78.   42. RPWRG
     .75   .00   .01   .15   .29   .85   .84   .73   .58   .50   .11   .00 REL

24.0   6.5   5.0   5.0   5.0   7.5 11.0 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
     54.4 45.7 25.6 27.1 54.4 54.4 54.4 54.4 54.4 54.4 54.4 54.4 ANGLE
     457. 317. 317. 357. 457. 457. 457. 457. 457. 457. 457. 457. V HITE
     78.   70.   74.   71.   68.   55.   29.   2.   -1.   -1.   -0.   0. SNR
      5.   -5.   -5.   -2.   9.   25.   52.   57.   67.   67.   66.   66. RPWRG
     .85   .95   .97   .85   .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 PPWRG 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.
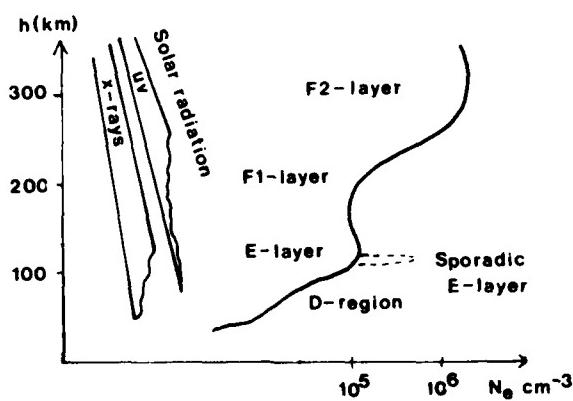
Figure 1   A typical undisturbed daytime electron density distribution. To the right the figure
            indicates which wavelength bands are absorbed at different heights. The E- and F-regions
   normally show clear maxima, whereas the F1 region is a ledge in the profile.
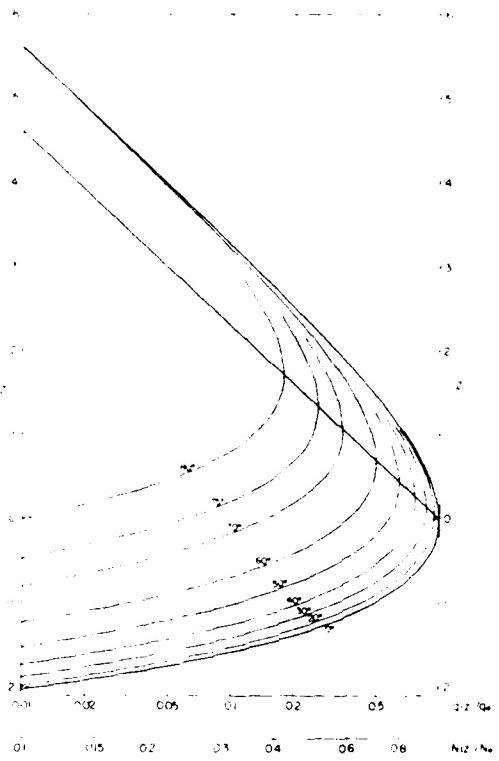
Figure 2   Normalized photoionization rate and electron density in a Chapman layer as a function of normalized height z and solar zenith angle χ. $z = \frac{h - h_0}{H}$ where h_0 is a reference height and H is the atmospheric scale height.
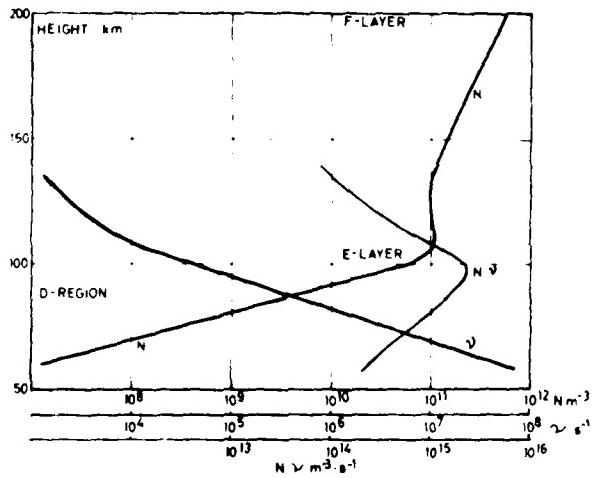

Figure 3   Typical height distributions of N_e, ν and $N_e\nu$ for daytime conditions.
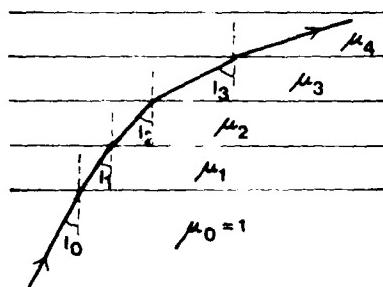

Figure 4   Refraction in a series of slabs according to Snell's law.
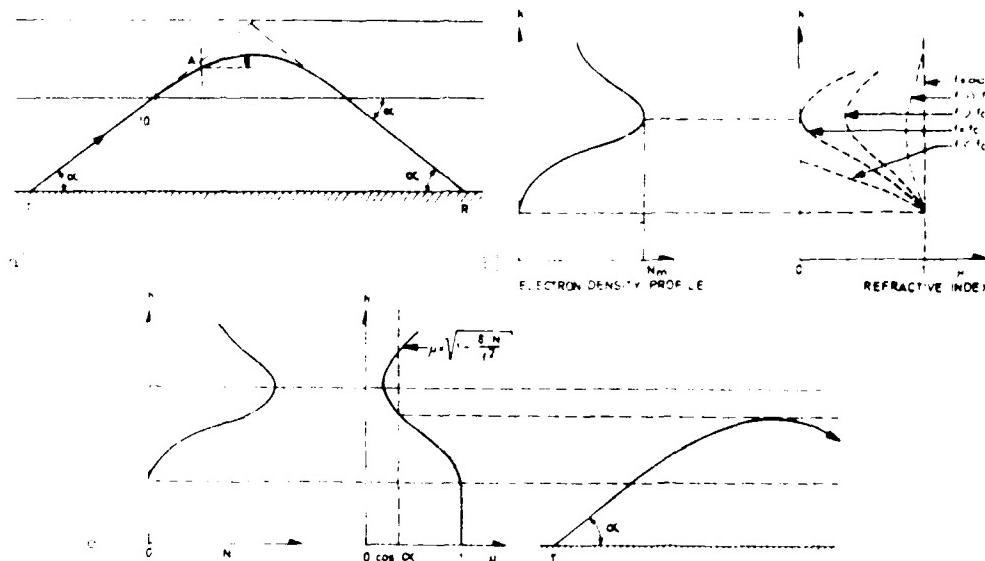
ELECTRON DENSITY PROFILE

REFRACTIVE INDEX

$$\mu = \sqrt{1 - \frac{8.1\,N}{f^2}}$$

Rays at the critical frequency 4 MHz

Rays at 5 MHz in the same ionosphere

Rays at 6 MHz

Figure 6   Raypaths for propagation at three different frequencies via a simple 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 μ.

a) 2 F2 mode

b) M - mode

c) 1E + 1F2 mode

d) Mode in an ionosphere with tilts
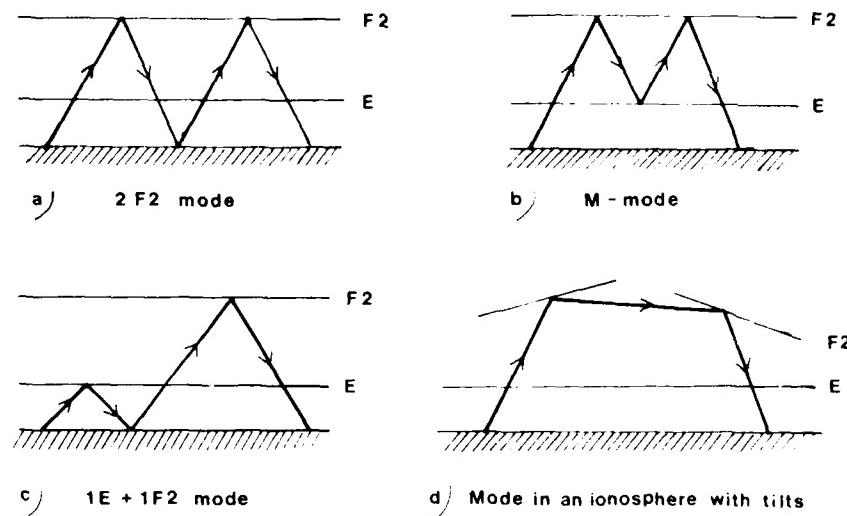
Figure 7   Examples of possible ionospheric modes.
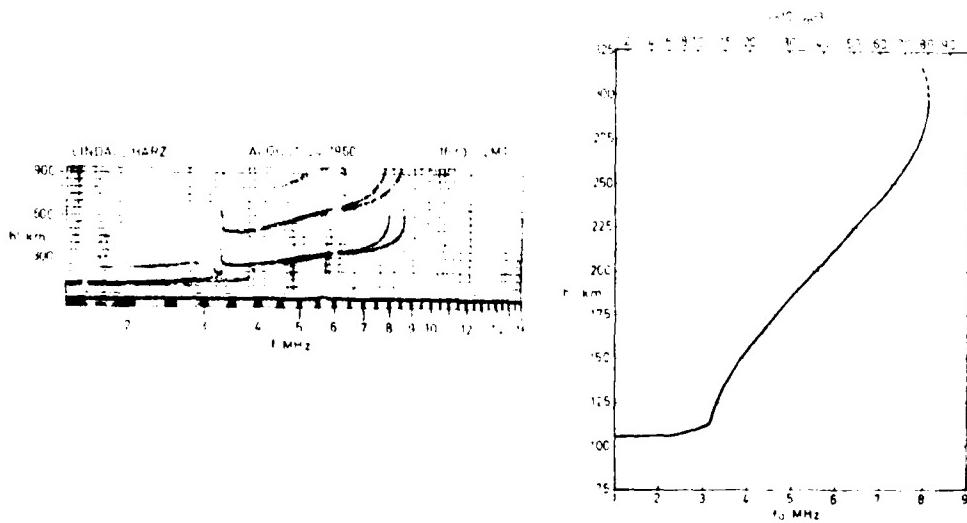
Figure 8   Typical ionosonde recording with corresponding electron density profile. Piet et al. 1967.

IONOSPHERIC VERTICAL SOUNDING STATIONS

• = 1959 DATA IN WDC-A
□ = 1961 DATA IN WDC-A
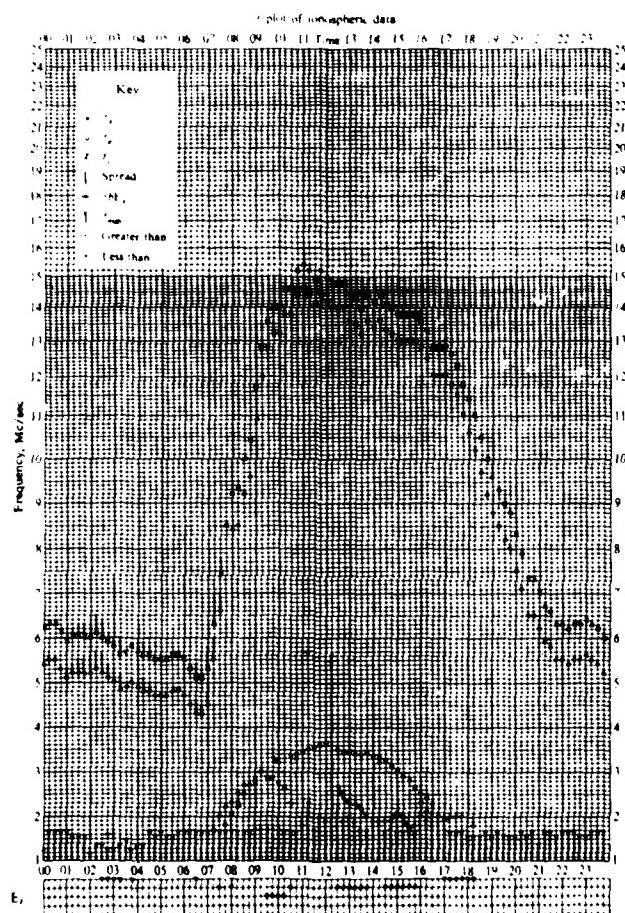
Figure 10.   F plot of critical frequencies for F- and E-layer as a function of time of day. Davies
1966.
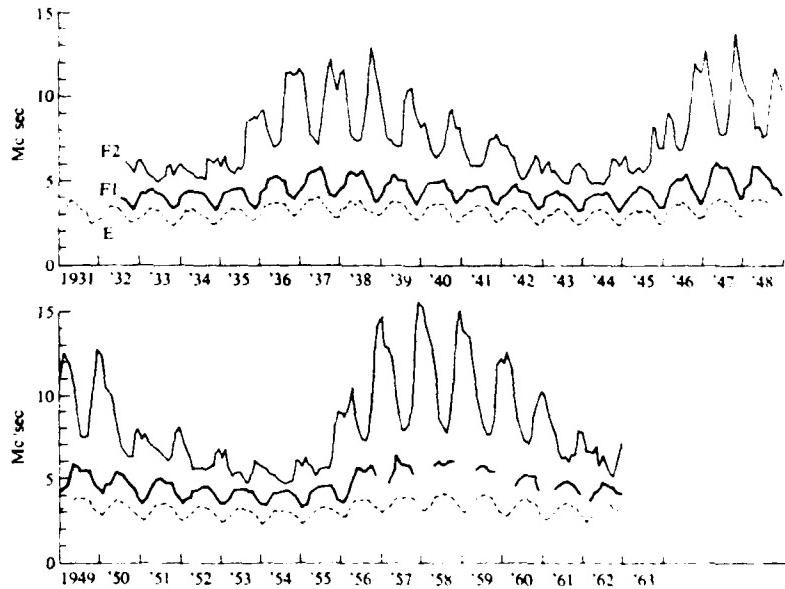
Figure 11   Long term variations of critical frequencies at Slough, England (Davies 1967).
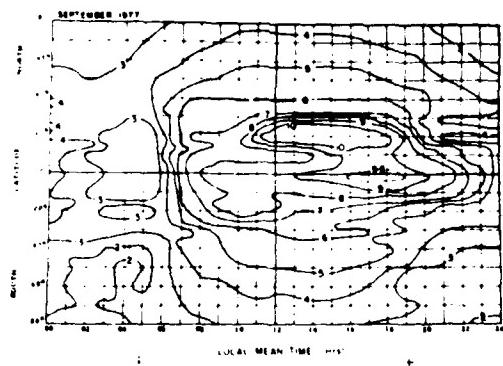
Figure 12   A contour map of foF2 in MHz for the east zone (50°E – 170°E) for September 1977 (Reddy et al 1979).
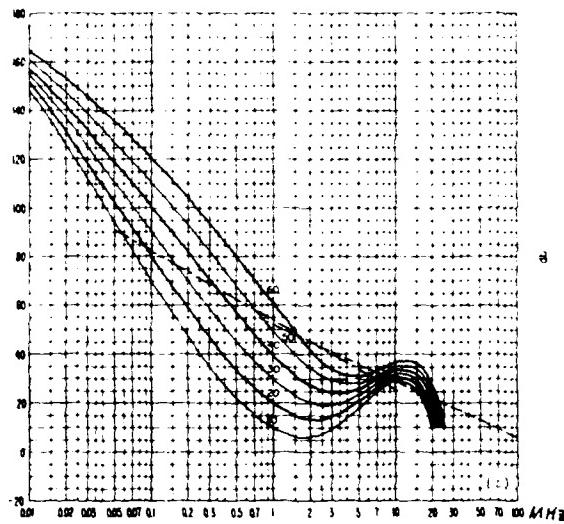
Figure 13   Variation of radio noise parameter F_a with frequency. F_a is the power available from a lossless antenna in dB over kT_0. T_0 is a reference temp = 288.39 K.
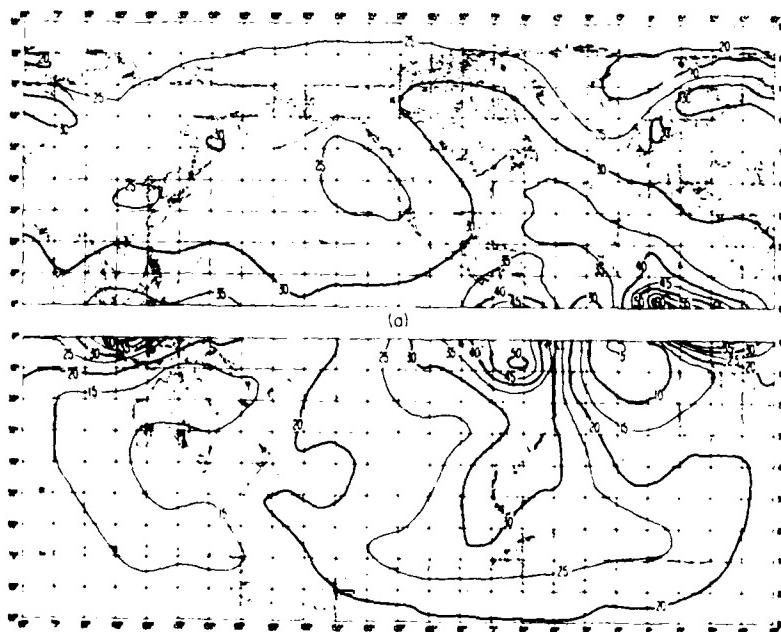

Figure 14   Contour map of F_a at 1 MHz for Winter 0800-1200 Local time.
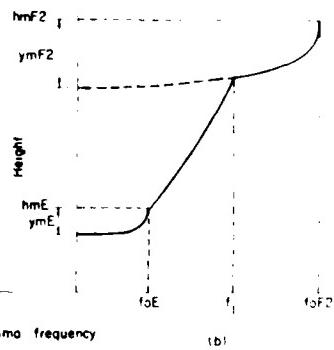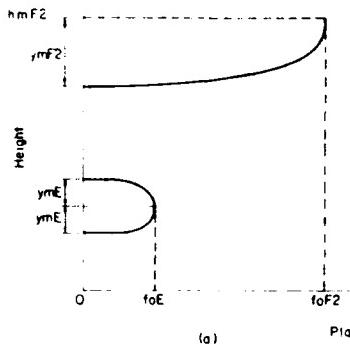
Figure 15a) $N_e(h)$ model for CCIR first procedure   Figure 15b) $N_e(h)$ model for CCIR second procedure.
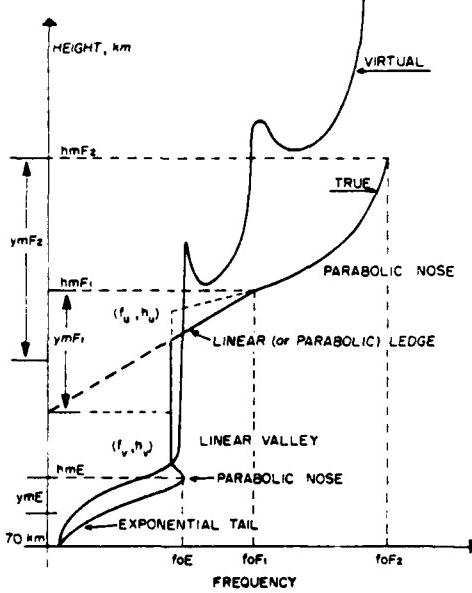

Figure 15c) $N_e(h)$ model used in the IONCAP procedure
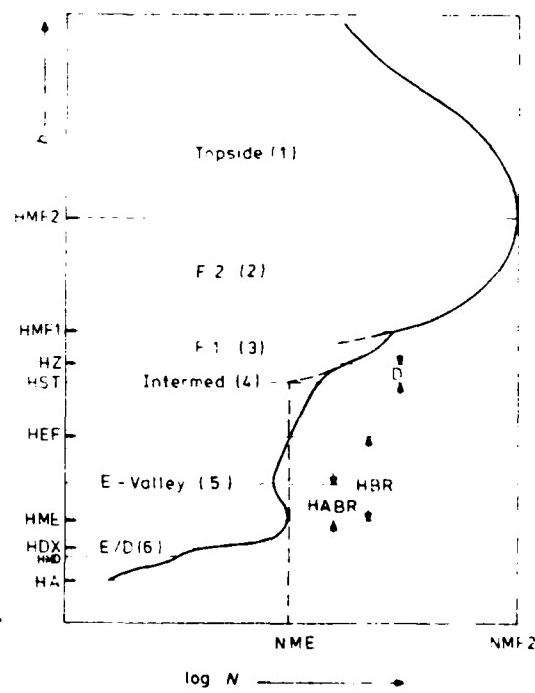
Figure 15d) $N_e(h)$ model used in the International Reference Ionosphere.
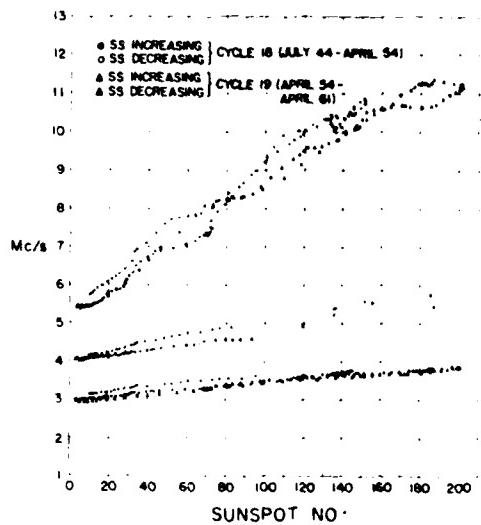
Figure 16   Critical frequencies of ionospheric
   layers versus smoothed sunspot
              number (Davies 1965).
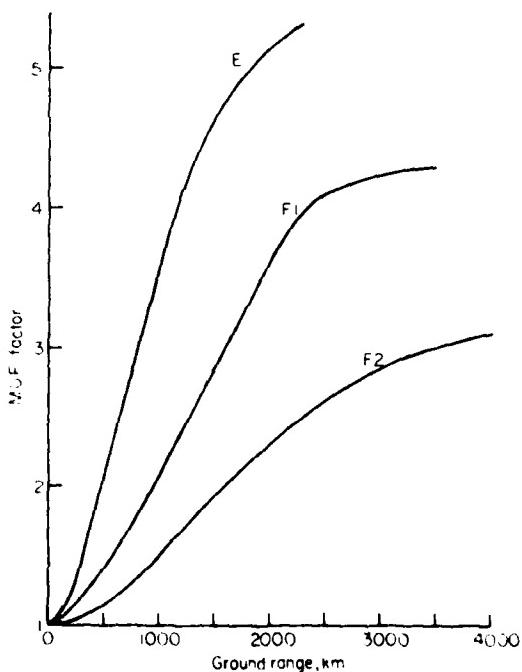
Figure 17   MUF factors for single hop E, F1 and
              F2 reflections (Bradley 1979).

METHOD 28   IONCAP 78.03

OSLO TO LONDON   MAR   1985   SSN = 77.
                                             AZIMUTHS   N. MI.   KM
59.92 N   10.75 E - 51.50 N   .08 W   220.52   51.54   622.6   1155.0
                              MINIMUM ANGLE   3.0 DEGREES
TTS- 1 ANTENNA PACKAGE
XMTR   2.0 TO 30.0 HORZ. DIPOLE H   -.25 L   -.57 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

MUF(....)   FOT(xxxx)   LUF(++++)

| 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	11.8	-
14.0	14.5	11.1	-
16.0	14.0	11.0	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.
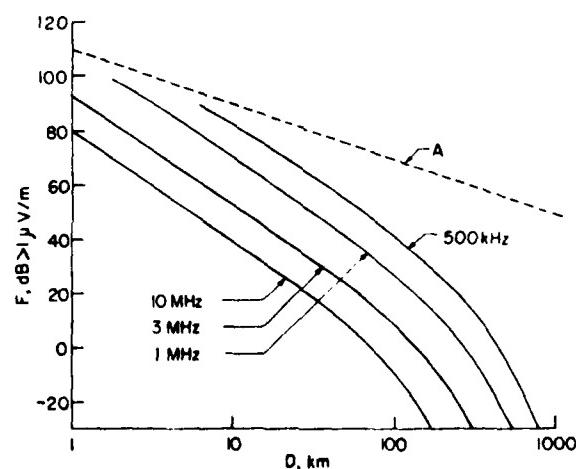
Figure 19   Ground wave field strength versus distance for ε = 4, and σ = 10^{-3} 8 m^{-1} (Bradley 1979).

REAL-TIME CHANNEL EVALUATION

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SUMMARY

The requirement for real-time channel evaluation (RTCE) in HF systems is identified and discussed in detail. Various scenarios in which RTCE is applicable are examined and classified.

Specific RTCE techniques and systems are then described including:

- Pulse/modulated pulse sounding
- Chirp sounding
- Limited channel monitoring
- Interference characterisation
- In-band RTCE

The application of RTCE to practical HF systems is discussed and the benefits accruing from its use quantified.

INTRODUCTION

As has been pointed out in a previous lecture (Darnell, 1983), real-time channel evaluation (RTCE), in the context of HF communications, corresponds to the process of system identification, or propagation path modelling, which must be carried out prior to the application of optimal control strategies. 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 provide 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% probability 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 than 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.

2. THE NATURE OF RTCE

There are two basic sources of signal distortion associated with HF paths, ie
time and frequency dispersion. The combined effects of these two distortion
mechanisms can be represented by the "channel scattering function", a typical example
of which is shown in Fig. 1. In essence, this function shows the dispersive effects
of the channel on an ideal impulse in the time domain and on a single frequency CW
tone in the frequency domain. In the example illustrated, three discrete propagation
modes are shown: these might arise from refraction by different ionospheric layers
and/or multiple refractions from individual layers. A given scattering function will
only be a valid description of the channel for a specific transmission frequency and
over an interval during which the "dispersion surfaces" remain sensibly constant.
However, it is not necessary for a communicator or HF system controller to know what
detailed, fundamental, physical principles give rise to the propagation phenomena
which affect the performance of the communication system - that is the task of the
ionospheric physicist; rather, that person should have access to the parameters of an
appropriate real-time model which describes the path behaviour adequately. This model
data can then be used to control the operation of the system and to adapt optimally
the signal generation and processing algorithms.

A definition of RTCE (Darnell, 1978), now adopted by CCIR (CCIR, 1981), is:

"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".

As a result of the above definition, the following points should be noted:

(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. In an earlier lecture
(Darnell, 1983), it was shown that RTCE is an essential
prerequisite for the application of automatic system control
procedures.

(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, 1983), 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.

Most channel models generated by practical RTCE systems are related to the channel scattering function and tend to fall into either the time domain or frequency domain category, depending upon the nature of the signals to be passed over the associated communication link.

3. A GENERALISED RTCE ALGORITHM

It is assumed that the state of any HF communications channel can be characterised by a set of n distinct measurable parameters which are functions of both frequency f and time t, ie

$$p_1(f, t), p_2(f, t), \ldots \ldots , p_n(f, t) \tag{1}$$

This set of parameters can be expressed as a column vector $(\underline{P}(f, t))$, where

$$(\underline{P}(f, t)) = \begin{pmatrix} p_1(f, t) \\ p_2(f, t) \\ \cdot \\ \cdot \\ \cdot \\ p_n(f, t) \end{pmatrix} \tag{2}$$

The definition of RTCE implies that the channel parameters will be sampled at a constant rate consistent with the minimum response time of the communication system; let this sampling interval be T and the number of alternative channels in the available set be m. Hence, at the k th sampling instant when

$$t = kT \tag{3}$$

and for the i th frequency channel of the set for which

$$f = f_i \qquad (i \text{ integer } 1 \leqslant i \leqslant m) \tag{4}$$

expression (2) becomes:

$$(\underline{P}(kT)) \Bigg|_{f_i} = \begin{pmatrix} p_1(kT) \\ p_2(kT) \\ \cdot \\ \cdot \\ \cdot \\ p_n(kT) \end{pmatrix} \Bigg|_{f_i} \tag{5}$$

Now, defining an "RTCE weighting" matrix as the row matrix $(\underline{A})$ where:

$$(\underline{A}) = \begin{pmatrix} a_1 & a_2 & \ldots \ldots a_n \end{pmatrix} \tag{6}$$

and an RTCE "figure of merit", Q_a, for the i th channel as:

$$Q_a(kT) \Bigg|_{f_i} = (\underline{A})(\underline{P}(kT)) \Bigg|_{f_i} \tag{7}$$

Therefore

$$Q_a(kT) \Bigg|_{f_i} = [a_1 p_1(kT) + \ldots + a_n p_n(kT)] \Bigg|_{f_i} \tag{8}$$

As will be discussed later in this lecture, in practical RTCE schemes only one of the parameters $p_1, p_2, \ldots p_n$ is normally employed as the basis of measurement; thus the RTCE weighting matrix reduces to:

$$(\underline{A}) = \begin{pmatrix} 0 & 0 & 0 & \ldots & 0 & a_j & 0 & \ldots & 0 & 0 & 0 \end{pmatrix} \tag{9}$$
$$(j \text{ integer } 1 \leqslant j \leqslant n)$$

and the figure of merit is given by:

$$Q_a(kT) \Bigg|_{f_i} = [a_j p_j(kT)] \Bigg|_{f_i} \tag{10}$$

In addition to the use of the sampled values of the set of measurable parameters for RTCE purposes, the various time derivatives of those parameters may also be sampled and used to characterise the behaviour of the channels with time. For example, a column vector of first time derivatives of the n parameters and a corresponding row vector of RTCE weighting coefficients can be defined in a similar manner, ie

$$(\underline{\dot{P}}(kT)) \Bigg|_{f_i} = \begin{pmatrix} \dot{p}_1(kT) \\ \dot{p}_2(kT) \\ \cdot \\ \cdot \\ \dot{p}_n(kT) \end{pmatrix} \Bigg|_{f_i} \tag{11}$$

and

$$(\underline{B}) = (b_1 \; b_2 \; \ldots\ldots \; b_n) \qquad\qquad (12)$$

The corresponding RTCE figure of merit for the i th channel is thus:

$$Q_b(kT)\Big|_{f_i} = (\underline{B})(\underline{\dot{P}}(kT))\Big|_{f_i} \qquad\qquad (13)$$

Again, in practical RTCE systems, the weighting matrix will take the form:

$$(\underline{B}) = (0\;0\;0\;\ldots\; b_j \;\ldots\; 0\;0\;0) \qquad\qquad (14)$$

Other figures of merit employing higher order time derivatives can be introduced if appropriate. The overall channel selection decision will in general depend upon a weighted combination of the various figures of merit which will be termed the "channel preference factor" (CPF), where

$$CPF(kT)\Big|_{f_i} = F\{ Q_a(kT),\; Q_b(kT),\; \ldots\} \Big|_{f_i} \qquad\qquad (15)$$

$$= F\{(\underline{A})(\underline{P}(kT)),\; (\underline{B})(\underline{\dot{P}}(kT)),\; \ldots\} \Big|_{f_i} \qquad\qquad (16)$$

The nature of the particular RTCE algorithm used in any given situation will determine the form of combining function F{ } and of the weighting matrices $(\underline{A})$, $(\underline{B})$, etc. However, in general, their form will be influenced primarily by:

    (a) The communications objectives of the system which the RTCE algorithm is intended to support, eg whether single mode propagation is necessary, whether the maximum possible signal-to-noise ratio (SNR) is required, what electromagnetic compatibilityconstraints have to be satisfied, etc;

    (b) The capabilities and limitations of the available RF and processing equipment.

The overall function of the RTCE system, therefore, is to compute the CPF for each member of the set of m available channels, ie

$$CPF(kT)\Big|_{f_i} \qquad \text{for } 1 \leqslant i \leqslant m \qquad\qquad (17)$$

and then to select the value, or values, of i for which the CPF is a maximum. A further important function of the RTCE process may be to rank the remaining channels of the set in order of preference in order to identify possible stand-by channels.

Therefore, in its simplest form, a generalised RTCE algorithm may be represented diagramatically as shown in Fig. 2 with the four basic inputs of

        - propagation data
        - interference/noise data
        - communication system objectives
        - equipment characteristics.

In a more detailed form, the generalised RTCE algorithm discussed above can be represented by the schematic arrangement illustrated in Fig. 3, for parameters and their first derivatives.

4. SCENARIOS FOR THE APPLICATION OF RTCE

Three practical HF communication scenarios in which RTCE tecniques are applicable will now be discussed (Darnell, 1975):

4.1 Class 1 RTCE: Remote Transmitted Signal Pre-processing

Fig. 4 shows a block diagram of what will be termed a Class 1 RTCE system. It is applicable to the situation in which a remote or mobile terminal wishes to pass traffic to a base station on a random or scheduled basis. An RTCE probing signal, x(t), is transmitted at appropriate intervals over the HF channel from the base station. Apart from spectral occupancy and EMC constraints, there is no limitation upon the form of x(t). Specialised processing equipment at the remote terminal performs an analysis of the received version of the RTCE signal, y(t), and hence fomulates an appropriate path model. In the modelling process, RTCE information from other sources can be incorporated if available, eg via the analysis of other RTCE probing signals from separate sources, data transmitted to the remote terminal via other propagation media such as satellites or VLF, etc. The model is then used to derive control data for pre-processing the remote transmitted signal, r(t), to yield a signal, r'(t), which is subsequently transmitted over the channel and received at the base as r"(t). Ideally

$$r''(t) = r(t - d) \qquad\qquad (18)$$

where d is the effective propagation delay. It is clear that this type of RTCE and processing necessitates an assumption of propagation reciprocity, ie the propagation model for the base-to-remote path is an adequate description of the remote-to-base path also. Normally, this premise is reasonably accurate, provided that allowance is made for any differences in characteristics between base and remote transmitter, receiver and antenna configurations.

4.2 Class II RTCE: Base Transmitted Signal Pre-processing

The basic block diagram of a Class II RTCE system is given as Fig. 5. In essence, the technique employs base station single-site RTCE in order to determine an appropriate pre-processing algorithm for communications transmissions from that base station to remote stations in defined locations. The RTCE probing signal, x(t), is radiated by the base transmitter; energy returned from the channel to the base in the form of a signal y(t) is then employed to formulate an appropriate channel model. The model is then used to derive the pre-processing algorithm which is applied to the base transmission, r(t), to give a signal r'(t). Note that r'(t) and x(t) may be multiplexed if operationally convenient. At the remote terminal, the signal r'(t) is received as r"(t) and again ideally

$$r"(t) \quad = \quad r(t - d) \qquad \qquad (19)$$

where d is the propagation delay. It is also possible for the RTCE modelling process to make use of data from other sources, eg by monitoring transmissions from other stations in the vicinity of the remote terminal(s). In practice, Class II RTCE is applicable to broadcast type systems.

4.3 Class III RTCE: Remote Received Signal Processing

Fig. 6 shows the general format of a Class III RTCE system. As in the case of Class II RTCE, the RTCE probing signal, x(t), and the base station traffic signal, r(t), are multiplexed prior to transmission over the HF channel. At the remote terminal, the received RTCE signal, y(t), is demultiplexed, processed and then employed to form a model of the channel which is subsequently used to control the signal processing strategy to be applied to the distorted version of the traffic signal, r'(t), to produce the corrected traffic estimate, r"(t). Again, the objective is to make this estimate identical with the original traffic signal. Once more, RTCE data from other sources can be incorporated into the model formulation process.

The scenarios outlined above describe open loop situations in which traffic flow is basically unidirectional. In many practical cases, bidirectional traffic flow will be required and thus equipment of say the Class I or Class III types might have to be provided at both terminals. Alternatively, the availability of feedback between the terminals in the form of a low-rate engineering order wire (EOW) would enhance the flexibility of the procedures and allow RTCE data to be transferred between receiver and transmitter. However, for reliability, the EOW itself would also require some form of RTCE. Auxiliary inputs to the RTCE process in the form of data passed via separate communications media, relay from other remote terminals, interpolation or extrapolation using other RTCE probing transmissions, etc, should always be sought to increase reliability.

5. PRACTICAL RTCE SYSTEMS

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:

(a) Signal amplitude;
(b) Signal frequency;
(c) Signal phase (absolute or differential);
(d) Propagation time (absolute or relative);
(e) Noise or interference level;
(f) Channel impulse response function;
(g) Signal-to-noise or signal-to-interference ratio;
(h) Energy distribution within the channel bandwidth;
(i) Received digital data error rate;
(j) Received speech intelligibility level;
(k) Telegraph distortion factor;
(l) Rate of repeat requests in an ARQ system.

Examples of RTCE systems which have been developed to at least a working prototype stage will now be described in the following sections. In general, practical RTCE systems fall into three basic categories:

- Those which operate on any frequency in the HF band, on the assumption that they cause negligible interference to other spectrum users;

- Those which operate only in the frequency channels assigned to the communication systems which they are designed to support;

> - Those which operate within a single assigned channel
>   which may, or may not, be passing communication
>   traffic.

Systems falling in the first of these categories will now be discussed.

6. RTCE SYSTEMS NOT CONSTRAINED TO OPERATE IN ASSIGNED CHANNELS

6.1 Pulse Sounding

The pulse sounding technique was originally developed as an aid to fundamental ionospheric research, with its value as an RTCE tool only being appreciated at a later stage. Pulse sounders require dedicated transmitters and receivers operating on the basis of time and frequency synchronism. A high power sounding transmitter radiates short pulses in a pre-determined time sequence on a large number of specified frequencies covering part, or the whole, of the HF band. The time/frequency schedule is under the control of the system program, which must be identical for both transmitter and receiver. The program timing is controlled by master clocks at transmitter and receiver; these independent clocks can themselves be synchronised by means of an external standard time transmission such as MSF or WWV, with appropriate allowance being made for differences in propagation time to the two sites. Alternatively, the requirement for alignment using an external standard can be eliminated if atomic clocks are available at the both transmitter and receiver.

If the transmitted pulse, x(t), is of short duration, the response of the sounding receiver corresponds to an approximate channel impulse response for each of the m channels on which a transmission is made. The received signal is given by the convolution integral

$$y(t) = \int_{-\infty}^{\infty} h(u) \, x(t - u) \, du \bigg|_{f_i \, 1 \leqslant i \leqslant m} \qquad (20)$$

where $h(u)$ is the unit impulse response function of the channel and u is a time variable. If $x(t)$ is an approximate impulse, then $y(t)$ is evidently proportional to the impulse response function $h(t)$. One way of overcoming very rapid variations in the response is to transmit several pulses on each channel and compute an average response.

The output information from an ionospheric sounder is normally presented in the form of a visual display termed an "ionogram", which is essentially a two-dimensional projection of a raster of impulse responses for the m channels, as illustrated in Fig. 7. Fig. 7(a) shows typical impulse responses, indicating the presence of different degrees of multipath propagation, taken from the complete m-channel array; Fig. 7(b) is a projection of this raster, in the sense indicated, which forms the ionogram display of ionospheric mode structure in the propagation delay (d) − frequency plane. In the RTCE context, the ionogram can then be used to select say a region of single-mode propagation having minimum time dispersion − as shown in Fig. 7(b). Alternatively, if no region of single mode propagation can be identified, a CPF for each channel could take the form:

$$CPF \bigg|_{f_i} = \left[\frac{\text{Energy in strongest mode}}{\text{Total energy in all modes}} \right] \bigg|_{f_i} \qquad (21)$$

The more nearly the CPF approaches unity, the closer propagation would be to single mode. Allowance would also have to be made for the relative total energies in the propagating channels to ensure that an adequate SNR was maintained.

6.2 Modulated Pulse Sounding

An important practical modification to the basic pulse sounding technique described above is to apply digital modulation to each of the transmitted pulses in order to increase the signal processing efficiency of the system. This process provides two important performance enhancements:

> (a) It enables pulse compression coding to be applied (in the
>   same way as in some types of radar) in order to improve the
>   time resolution of the system without having to resort to
>   shorter pulses, and hence increased peak transmitter powers.
>   For a given time resolution and quality of impulse response,
>   it is necessary to use a certain amount of transmitted
>   energy to probe the channel; for unmodulated pulses, this
>   energy must be applied in the form of short-duration, high
>   amplitude pulses whilst, with pulse compression modulation,
>   the energy can be applied at lower amplitude over a much
>   longer interval, but still achieving the required time
>   resolution.
> (b) The RTCE transmission can be encoded with small amounts of
>   data via manipulation of the pulse modulation, eg to
>   describe noise/interference levels in assigned channels at
>   the transmitter site.

Several forms of pulse compression coding have been developed; these include Barker codes (Barker, 1953), Huffman sequences (Coll & Storey, 1964), binary sequences of length > 13 bits with autocorrelation functions (acf's) approximating to an impulse (Mann, 1968) and complementary sequences (Darnell, 1975). With all these forms of modulating signal, the impulse response of a given channel is obtained by computing the input-output crosscorrelation function (ccf). For a linear system with input x(t), output y(t) and unit impulse response function h(t), the input-output ccf, $\phi_{xy}(\tau)$, is given by (Lee, 1960):

$$\phi_{xy}(\tau) \quad = \quad 1/T' \int_{-T'/2}^{+T'/2} x(t)\ y(t + \tau)\ dt \qquad\qquad (22)$$

$$= \quad \int_{-\infty}^{\infty} h(u)\ \phi_{xx}(\tau - u)\ du \qquad\qquad (23)$$

where τ and u are time variables, T' is the correlation interval and $\phi_{xx}(\tau)$ is the input acf. Using expression (23), it can be seen that if the input acf is an approximate impulse, then the ccf $\phi_{xy}(\tau)$ will be approximately proportional to the system impulse response function.

Ionospheric pulse sounding is a widely-used method of RTCE; many individual sounders exist for specific communication paths, but relatively few networks have yet been implemented. Possibly the most ambitious pulse sounding scheme designed to date was the Common User Radio Transmission System (CURTS) (Probst, 1968), in which a network of pulse sounding transmitters was set up giving complete area coverage for all users with compatible sounding receivers. CURTS is an example of a Class I RTCE system.

6.3 Chirp Sounding

It is also possible to employ a fundamentally different technique known as "chirp" sounding to obtain an ionogram display: as its name implies, chirp sounding makes use of a swept-frequency transmission as a channel probing signal (Barry & Fenwick, 1965). The sweep is typically linear with time, but may take other forms. Again, synchronisation between transmitter and receiver is necessary. Fig. 8 illustrates the principle of the technique: in a multipath propagation situation, several weighted versions of the transmitted sweep will be received as shown. If a correctly timed local oscillator sweep is available at the receiving site, this can be mixed with the incoming sweep components to yield the difference frequency components which are then subjected to spectral analysis.

At time t_k, the frequency of the synchronised local oscillator sweep is

$$f_{min} \ + \ \frac{df}{dt}\ t_k \qquad\qquad (24)$$

whilst the corresponding frequencies of the received component sweeps are

$$f_{min} \ + \ \frac{df}{dt}\ (t_k\ -\ \Delta t_1) \qquad\qquad (25)$$

$$f_{min} \ + \ \frac{df}{dt}\ (t_k\ -\ \Delta t_2) \qquad\qquad (26)$$

$$f_{min} \ + \ \frac{df}{dt}\ (t_k\ -\ \Delta t_3) \qquad\qquad (27)$$

After mixing with the local oscillator signal, the frequency components of the difference signal are:

$$\Delta t_1\ \frac{df}{dt} \qquad \Delta t_2\ \frac{df}{dt} \qquad \Delta t_3\ \frac{df}{dt} \qquad\qquad (28)$$

Hence, propagation delays are translated directly into frequency offsets. Assuming that the mixing process is linear, the relative amplitudes of the individual received sweep components will be preserved. Time dispersion due to the distributed nature of the ionospheric refraction process will cause the received sweep components to be broadened away from ideal spectral lines. Therefore, if the mixer output is displayed on a spectrum analyser, a propagation mode profile equivalent to the channel impulse response will result; taking a projection of the spectrum analyser output as a function of local oscillator frequency will again yield an ionogram.

6.4 Modes of Operation for Ionospheric Sounders

The ionospheric sounding systems described in the previous three sections can be operated in oblique incidence, vertical incidence or backscatter modes.

Oblique incidence implies that the sounding transmitter and receiver are geographically separated so that the transmitted energy impinges upon the ionospheric layers obliquely. This form of sounding can thus be used as the basis of a Class I or

Class III RTCE system.

Vertical incidence sounding employs a transmitter and receiver which are co-sited. The sounding transmissions are directed vertically upwards at the ionosphere in order to determine its structure just above the sounding site. Oblique incidence characteristics may be inferred from vertical incidence measurements. This type of single-site RTCE is therefore well suited to Class II scenarios.

Backscatter sounding is similar in character to vertical incidence sounding in that it can be carried out from a single site, or closely spaced transmitter and receiver sites. However, the transmitted energy is now radiated obliquely rather than vertically to enable the ionospheric structure in a desired direction of propagation to be evaluated. The received signal arises from energy which has been refracted by the ionosphere over a path away from the transmitter, reflected from the earth's surface, and then propagates back to the receiver via a similar ionospheric refraction in the reverse sense. The technique is normally only applicable to single hop paths since multiple hops give rise to excessive received signal attenuation. Also, the received scattered energy is at a much lower level than with vertical or oblique incidence propagation, thus necessitating much higher radiated powers and giving poorer definition. Backscatter sounding is clearly applicable to Class II RTCE scenarios.

7. RTCE SYSTEMS CONSTRAINED TO OPERATE IN ASSIGNED CHANNELS

Whereas the RTCE systems described in the previous sections are designed to operate anywhere in the HF band on the assumption that the interference they cause to other users of the spectrum will be negligible, there are other forms of RTCE specifically intended to function only in the channels assigned for use by the communication systems which they are required to support. Since the assumption of negligible interference by ionospheric sounders is questionable, often being critically dependent upon the nature of the transmission being interfered with, the latter class of RTCE systems would appear to have more potential for widespread application in future by virtue of its more efficient spectrum utilisation; also, as discussed in an earlier lecture (Darnell, 1983), such systems would tend to employ the same RF and processing units as used by the communication system, thus resulting in economy of implementation.

7.1 Channel Evaluation and 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 several 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. At the remote receiver alerted by the selective calling code, the base station average noise levels

$$\overline{n(t)}\ \Big|_{f_i} \quad 1 \leqslant i \leqslant m \tag{29}$$

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

$$\overline{n(t)}\ \Big|_{f_j} \quad \text{where j can take any k distinct values} \atop \text{in the range 1 to m and } k \leqslant m \tag{30}$$

The subset of corresponding average received signal levels at the mobile

$$\overline{A(t)}\ \Big|_{f_j} \tag{31}$$

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

$$SNR(t)_{base}\ \Big|_{f_j} = \left[G\ \frac{\overline{A(t)}}{\overline{n(t)}} \right]\ \Big|_{f_j} \tag{32}$$

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 by ASWE in the UK (Wynne, 1979) and the Canadian radio telephone with automatic channel evaluation (RACE) system (Chow et al, 1981). The practical situations in which CHEC-

type systems have been applied correspond to the Class I scenario defined in Section 4.

7.2 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 transmission 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, θ_n, compared with the phase difference measured and stored at the previous sampling instant, θ_{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, θ_t, a "phase error" is counted, ie

$$\left| \theta_n - \theta_{n-1} \right| > \theta_t \qquad (33)$$

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. 9. 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. 10. 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 CPF.

Pilot tone RTCE could be used in a Class I 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. 11, 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.

7.3 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 necessary is somewhat less than that for 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 CPF 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 intellibility - as shown by the empirical model for 1.2 kbits/s digitised speech given in Fig. 12.

A similar, but less precise, relationship exists between analogue speech intelligibility and data error rate. Fig. 13 shows a baseband spectrum in which LINCOMPEX-processed speech (Awcock, 1968) is frequency multiplexed with low-rate

binary FEK telegraphy. Fig. 14 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. A technique termed "pseudo-error" counting has been proposed to overcome this problem (Leon, 1973); here the error rate is artificially amplified by the 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.

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.15. 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.

The RTCE by error counting method is chiefly applicable to Class I scenarios.

8. RTCE SYSTEMS OPERATING WITHIN A SINGLE ASSIGNED CHANNEL

All the RTCE techniques described in the previous sections have been applicable to situations in which the communicator has available for his use a number of assigned channels. In many cases, however, a communicator may wish to examine the state of a particular channel in more detail, eg:

    (a) To assess the state of the channel currently carrying traffic relative to the states of alternative channels; obviously, for a number of reasons, it may not be possible to employ the same RTCE algorithm for evaluating the traffic-carrying channel as for evaluating stand-by channels.
    (b) 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.
    (c) To determine the optimum signal processing procedures to be applied to a traffic transmission within the channel by making use of an appropriate RTCE model (Class III operation).

Various RTCE techniques applicable to this single assigned channel situation will now be described.

8.1 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.

8.2 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 CPF, ie

$$CPF = F\{ \left| \theta_r(t) - \theta_o(t) \right| \} \tag{34}$$

Alternatively, the CPF 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.

8.3 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 protection 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.

8.4 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 7.2, 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. 16 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.

9. 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. Similarly, 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 8.1 could form the basis of interference assessment systems for incorporation into RTCE procedures. Other systems can also be used (Cottrell, 1979) (Barry & Fenwick, 1975).

In some cases, the interference assessment will be explicit; in others, such as the error counting technique described in Section 7.3, the RTCE process evaluates the combined effects of both propagation and interference in a single measurement process.

10. CONCLUSIONS

10.1 General

As was pointed out in a previous lecture (Darnell, 1983), the rationale for the development of RTCE techniques is that significant improvements in the use of the HF propagation medium can only be achieved if a communicator, or HF system controller, using a specific path at a given time has access to real-time data on the relevant path parameters, rather than having to rely on off-line propagation analysis which can be subject to appreciable inaccuracy. In particular, off-line techniques can never provide accurate information on noise and interference levels in a given channel at a given time - although, with further refinement, they may well provide a reasonable statistical model for such interference which can be used in the system design process. The need for RTCE is most pronounced for links involving mobile terminals since the nature, orientation, etc of the paths will change with time thus making off-line analysis more approximate.

To date, there has been considerable experimental work on alternative RTCE techniques, but relatively little has been published quantifying their benefits in relation to systems making use of off-line data. One set of results (Darnell, 1978) indicates that an improvement in circuit availability of the order of 45% can be achieved by simple error counting RTCE in comparison with 2-frequency operation; however, much more performance data is required for other algorithms.

It would seem that dedicated RTCE systems such as ionospheric sounders, which require expensive special-purpose equipment and cause significant spectral congestion, will not find wide application in HF communications. Rather, RTCE procedures which can be integrated into the communications system, will use the same basic equipment and will operate only in assigned channels appear to offer a more logical and economic way forward. This may well place additional requirements upon the equipment specified for future HF communications in terms of control, flexibility, frequency agility. etc.

Bearing in mind the above comments, the techniques which currently appear to offer the greatest promise are:

(a)   Simple error counting (Section 7.3).
(b)   Pilot tone phase error measurement (Sections 7.2 and 8.4).
(c)   Systems based upon the general CHEC principle (Section 7.1).
(d)   The use of auxiliary EDC   processing in encrypted systems (Section 8.4).
(e)   Passive monitoring of noise and interference characteristics (Sections 8.1 and 9).
(f)   In-band RTCE within a   nominal 3 kHz assigned channel (Section 8.1).

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.

10.2 Potential Advantages of RTCE

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 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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FREQUENCY

ENERGY

FREQUENCY
DOMAIN
PROJECTION

PROPAGATION

INTERFERENCE NOISE

OBJECTIVES

EQUIPMENT
CHARACTERISTICS

GENERALISED
RTCE
ALGORITHM

CHANNEL SELECTION
DECISION

ENERGY

TIME

TIME DOMAIN PROJECTION
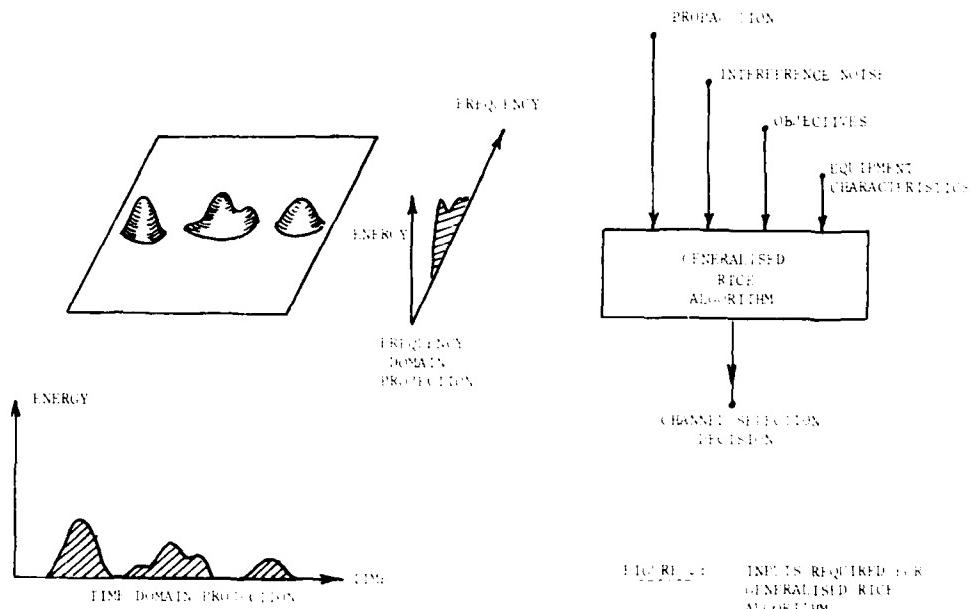
FIGURE 2:   INPUTS REQUIRED FOR
GENERALISED RTCE
ALGORITHM

FIGURE 1:   TYPICAL CHANNEL SCATTERING FUNCTION

PARAMETERS DESCRIBING CHANNEL STATE
(PROPAGATION & NOISE)

SAMPLE IN TIME
& FREQUENCY

EQUIPMENT
CHARACTERISTICS

COMMUNICATION
SYSTEM
OBJECTIVES

$\frac{d}{dt}$ $\frac{d}{dt}$ $\frac{d}{dt}$

a_1 a_2 a_n b_1 b b_n

Σ Σ

$Q_a(kT) \big/_{i}$ $Q_b(kT) \big/_{i}$

F $\{$ $\}$

CPF

CHANNEL
SELECTION
DECISION

VALUE OF i FOR
SELECTED CHANNEL
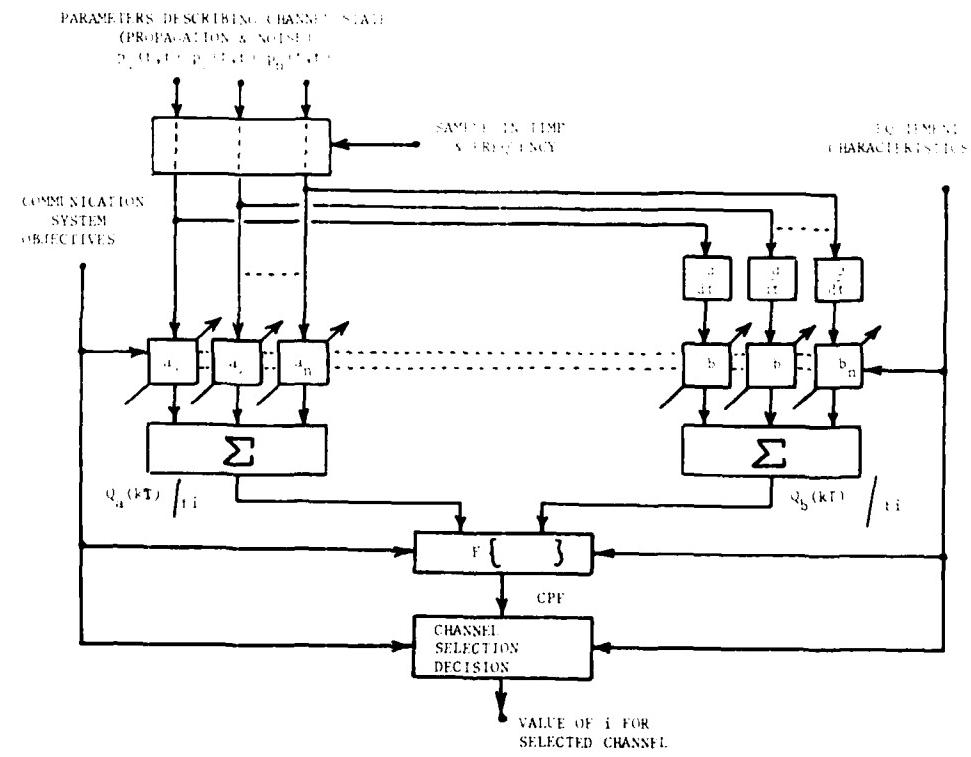
FIGURE 3:   GENERALISED RTCE ALGORITHM
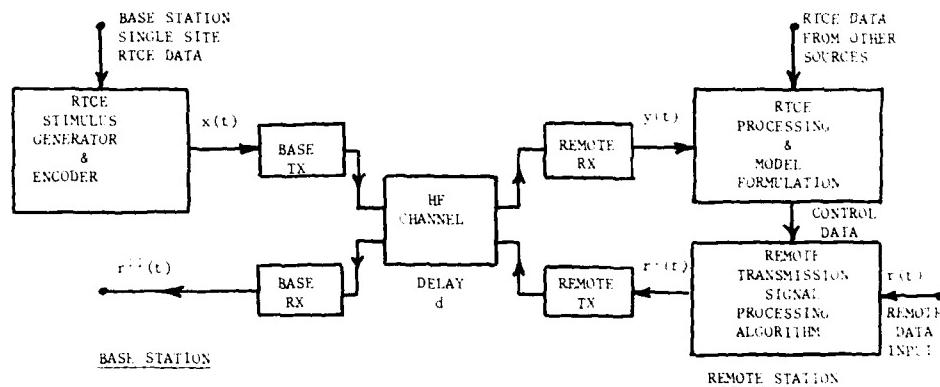
FIGURE 4: CLASS I RTCE WITH REMOTE TRANSMITTED SIGNAL PRE-PROCESSING
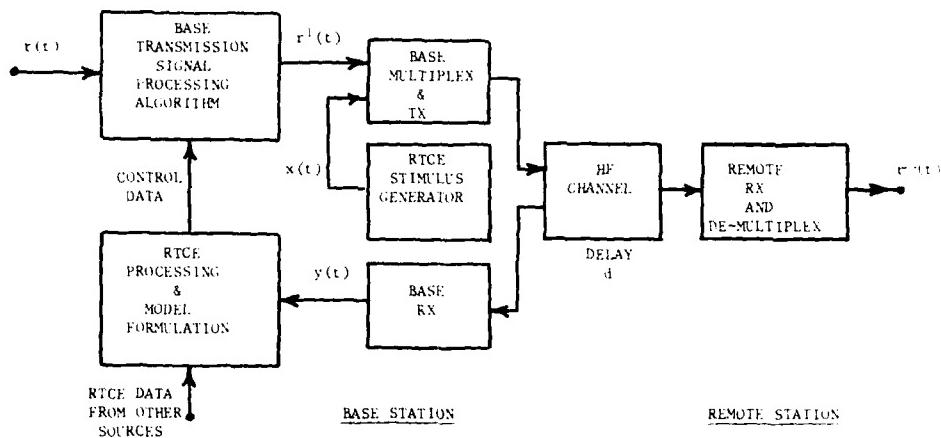
FIGURE 5: CLASS II RTCE WITH BASE TRANSMITTED SIGNAL PRE-PROCESSING
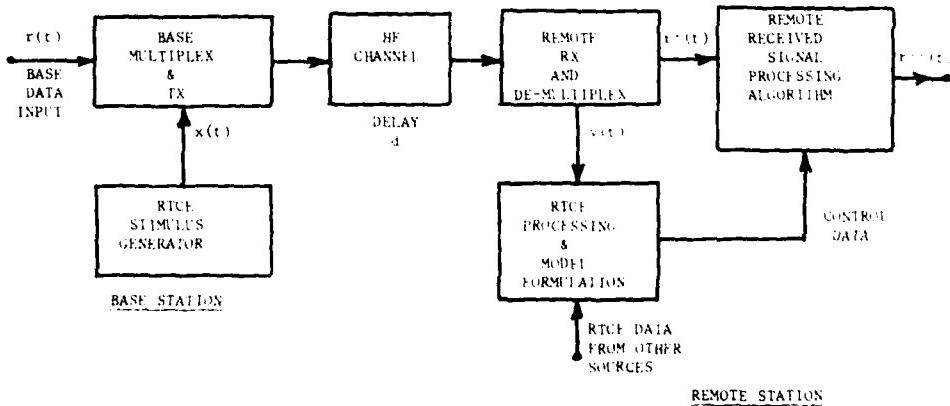
FIGURE 6: CLASS III RTCE WITH REMOTE RECEIVED SIGNAL PROCESSING
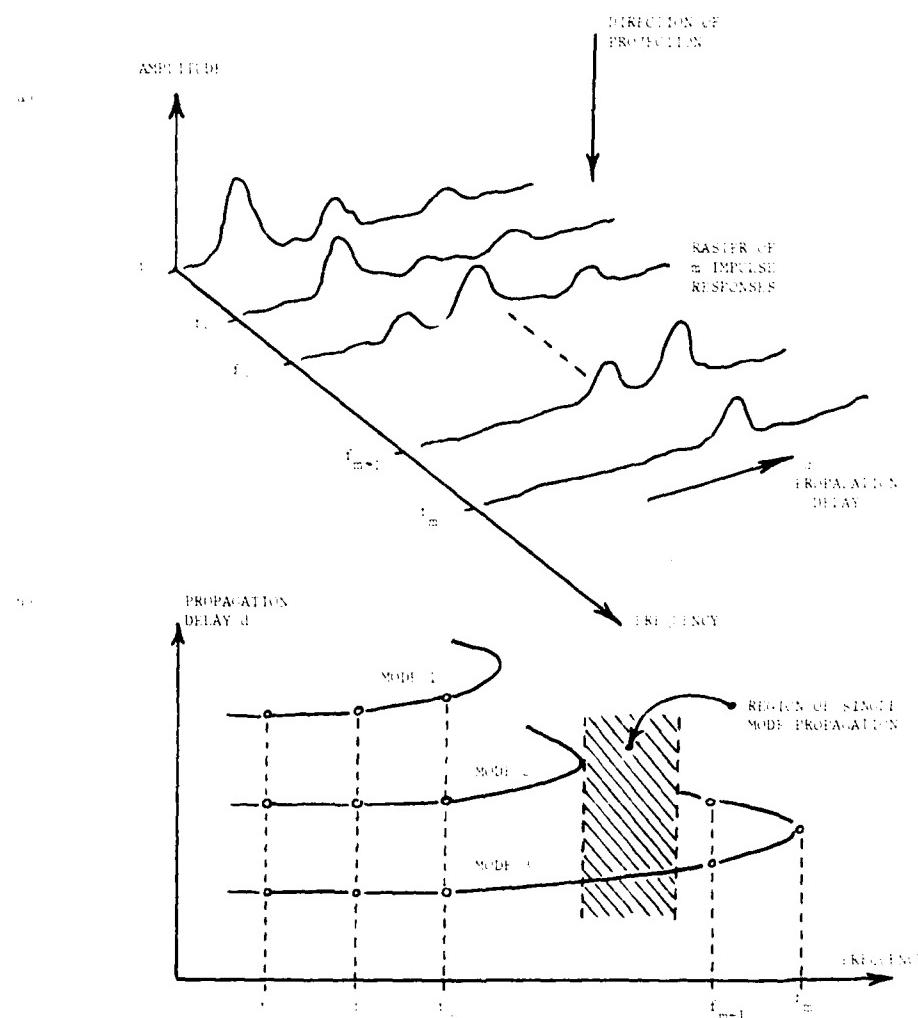
FIGURE 7: DERIVATION OF IONOGRAM FROM m-CHANNEL IMPULSE RESPONSE RASTER
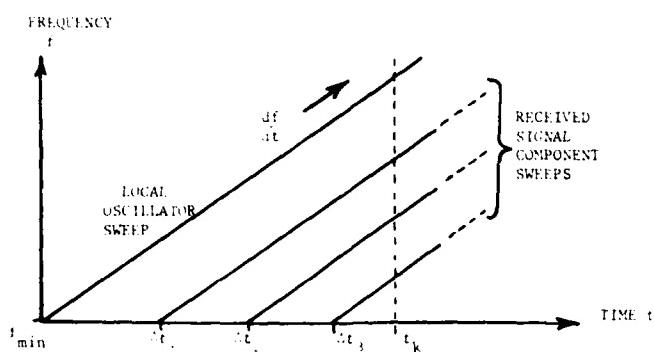

FIGURE 8: PRINCIPLE OF CHIRP SOUNDING

AMPLITUDE
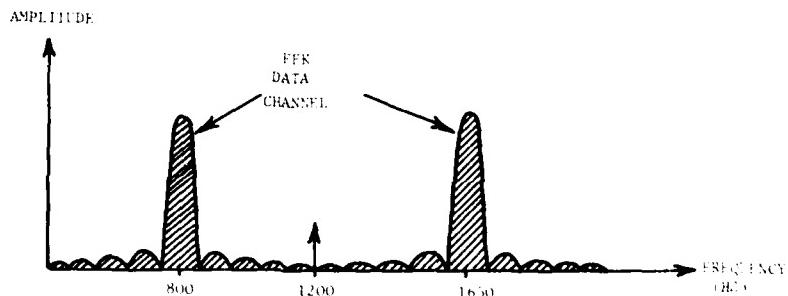
FIGURE 9: BASEBAND SPECTRUM OF FFK DATA CHANNEL WITH PILOT TONE INSERTED
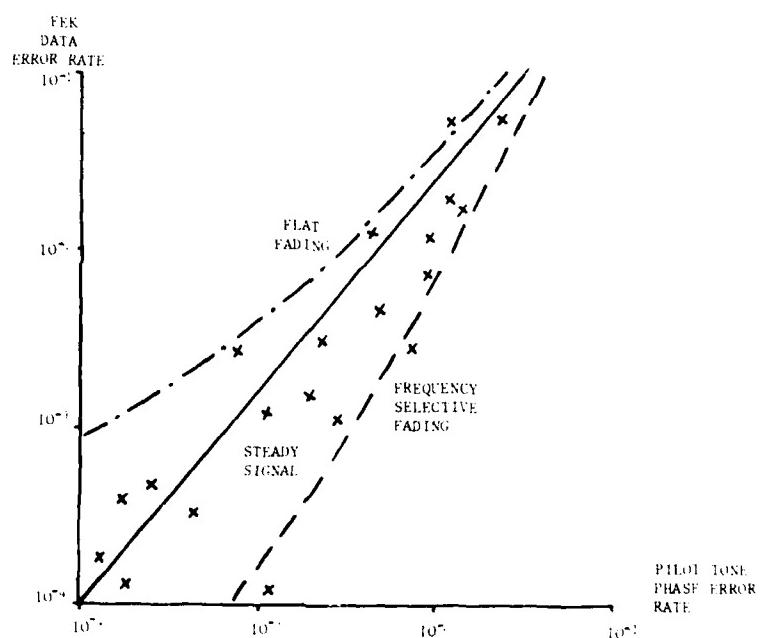
FIGURE 10: THEORETICAL AND EMPIRICAL RELATIONSHIPS BETWEEN PILOT TONE
PHASE ERROR RATE AND FFK DATA ERROR RATE
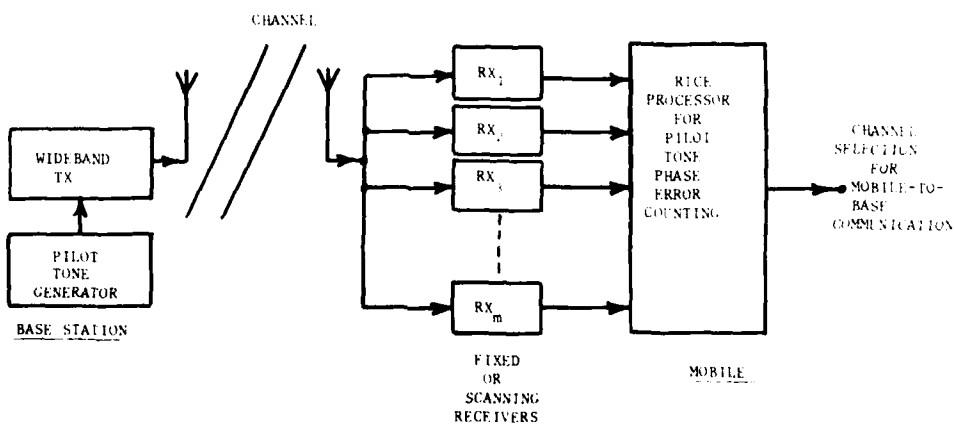
FIGURE 11: CLASS I APPLICATION OF PILOT TONE RICE

SPEECH
INTELLIGIBILITY
(%)
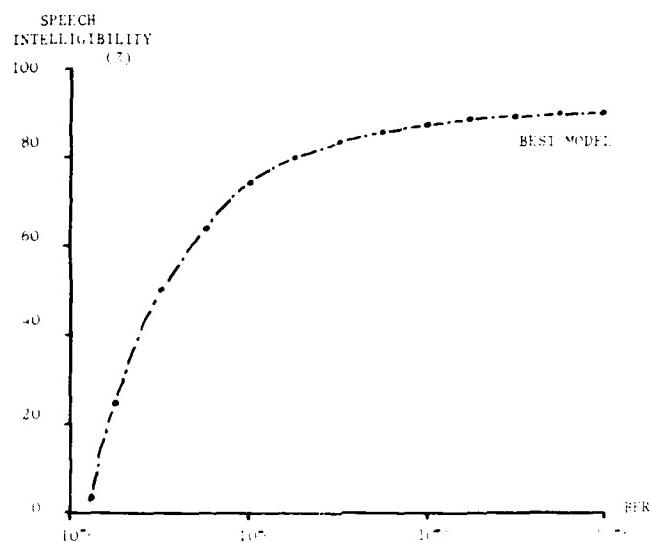
FIGURE 12: EMPIRICAL RELATIONSHIP BETWEEN DIGITIZED SPEECH INTELLIGIBILITY
AND DATA ERROR RATE (1200 BITS S)
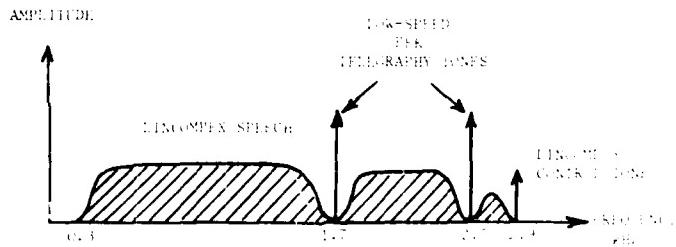
FIGURE 13: COMPOSITE BASEBAND SPECTRUM OF LINCOMPEX-PROCESSED SPEECH
& FFK TELEGRAPHY
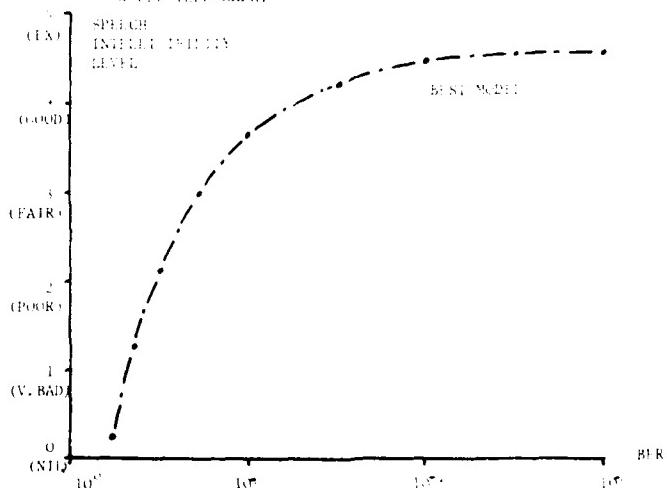
FIGURE 14: EMPIRICAL RELATIONSHIP BETWEEN LINCOMPEX-PROCESSED SPEECH
INTELLIGIBILITY & FFK TELEGRAPHY BER
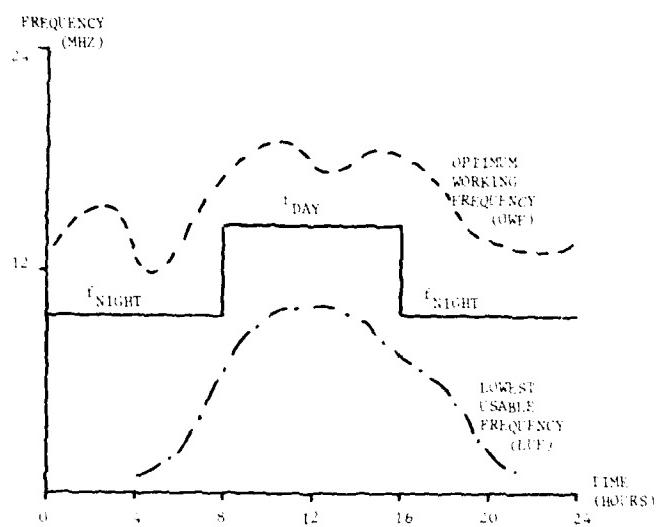
FREQUENCY
(MHZ)

FIGURE 15: EXAMPLE OF 2-FREQUENCY OPERATION
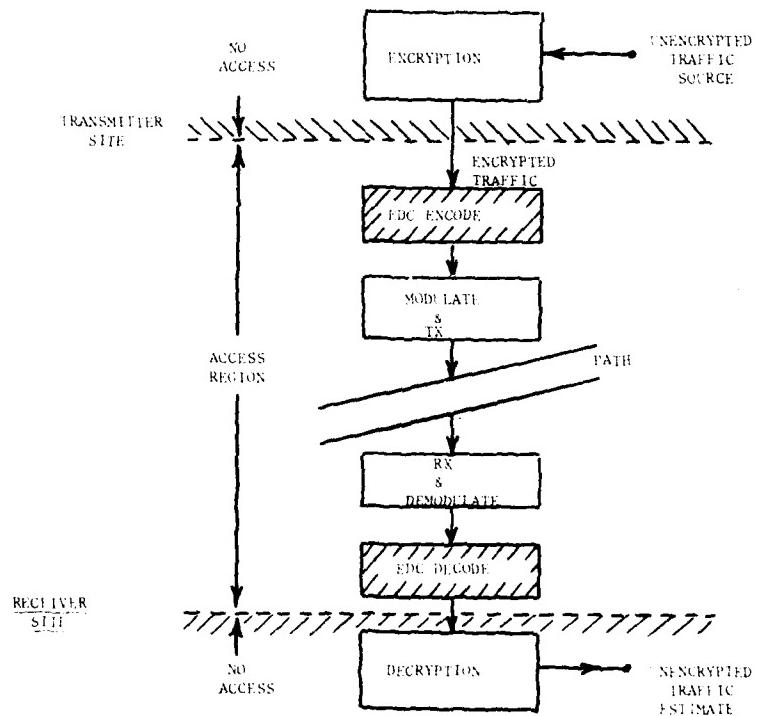
FIGURE 16: THE USE OF AN AUXILIARY EDC SYSTEM FOR RTCE WITH ENCRYPTED TRAFFIC

INTRODUCTION TO CODING FOR HF COMMUNICATIONS

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ABSTRACT

The increasing demand for reliable, low error-rate, high-speed digital data transmission at HF has created the need for the adoption of coding schemes. As it is well appreciated by the practicing communicator, a problem in high-speed data transmission is the occurrence of errors. Codes provide an effective approach for the reduction of the error rate. 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 immediately 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 preceeding them, but in other blocks as well. For the various coding schemes reviewed in the lecture, several numerical examples are given , to help in the quantitative appraisal of the merits of a 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 widening 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 enters the picture, as an approach to optimize communications on a given channel (in our case, ionospheric HF channels) rather than a way to achieve secrecy in military communications.

In order to better illustrate the point above, we must backtrack a few 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 every 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 'ave 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 every 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 $n = 5 \times 7 = 35$) 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 even in each line and in each column. At the end, we add a symbol at the lower right corner,to make even the number of 1s contained in the last line. The new block is as follows :

```
0001011 | 1
0100001 | 0
0010011 | 1
0010011 | 1
0000111 | 1
0101101 | 0
```

If during the transmission process an error occurrs, this can be corrected, provided that there is only one of them. Correction is done by checking each line and each column for even parity (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 in order to provide the code word with error-correcting capability.

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 $2 \cdot 10^{-4}$, $4 \cdot 10^{-4}$, $8 \cdot 10^{-4}$, $1.4 \cdot 10^{-3}$ and $2.4 \cdot 10^{-3}$. 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. These 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 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 $q_c = 8 \ 10^{-4}$, if we signal at the rate in the modem of 3600 bits/sec. Numerically, the equation above yields:

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

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 $q_c = 2.4 \ 10^{-3}$. 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 :

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

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 retransmits 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 for message retransmission, 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 (if 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 retransmission. An error now occurs only if all three bits of the triplet are wrong. We have therefore for P_e :

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

In fact, because the modem now works at 3600 bits/sec, we know that $q_c = 8 \ 10^{-4}$. 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 preceeding the check bits. In convolutional codes, check bits are continously interleaved with information bits, not only in the block immediately preceeding 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
We will now illustrate 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 kxn , and has the form:

$$G = \left[I_k \mid P \right] \quad kxn$$

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

$$G = \begin{bmatrix} 1 & 0 & 0 & 0 & 1 & 1 \\ 0 & 1 & 0 & 1 & 1 & 1 \\ 0 & 0 & 1 & 1 & 1 & 0 \end{bmatrix}$$

and we want to find all code vectors of this code. The message block size for this code 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 :

$$H = \left[P^T \mid I_{n-k} \right] \quad (n-k)xn$$

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 another example of codeword generation. We want to encode an 11 bit data sequence into a (15,11) code word. This code is fully specified by the related 11-by-15 generator matrix. This could be the matrix here below:

$$G = \left[I_k \mid P \right] = \begin{bmatrix} 10000000000 & 1001 \\ 01000000000 & 1101 \\ 00100000000 & 1111 \\ 00010000000 & 1110 \\ 00001000000 & 0111 \\ 00000100000 & 1010 \\ 00000010000 & 0101 \\ 00000001000 & 1011 \\ 00000000100 & 1100 \\ 00000000010 & 0110 \\ 00000000001 & 0011 \end{bmatrix}$$

In this matrix, the 11x11 part is just the identity matrix of order 11. The second part is totally arbitrary, as already indicated. Encoding the data sequence D = [100010011010] gives : C=DG = [100001001010 0011] . Note that the first 11 bits of C are identical to D and that the last 4 bits of C are the sums of the bits of D specified by the last four columns of G. For example, the first parity bit is the sum of the first, second, third, fourth, sixth, eight, ninth bits of D, that is 1+0+0+0+0+1+0 =0. In an equivalent way, C can be thought of as the sum of the rows of G which correspond to ones in P. Thus, in the example, C is the sum of the first, fifth, eight, and tenth rows of G.

As defined, a code word in a systematic linear code is an n-tuple with the property that the n k subsets of the k information bits of the word specified by the rightmost n-k colums of the generator matrix add to the corresponding parity check bit.

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 | 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

$$G = \begin{bmatrix} 1 & 0 & 0 & 0 & | & 1 & 1 & 1 \\ 0 & 1 & 0 & 0 & | & 1 & 1 & 0 \\ 0 & 0 & 1 & 0 & | & 1 & 0 & 1 \\ 0 & 0 & 0 & 1 & | & 0 & 1 & 1 \end{bmatrix}$$

$$\underbrace{\qquad\qquad}_{I_4} \quad \underbrace{\quad}_{P}$$

The parity check matrix H for this code is:

$$H = \begin{bmatrix} 1 & 1 & 1 & 0 & | & 1 & 0 & 0 \\ 1 & 1 & 0 & 1 & | & 0 & 1 & 0 \\ 1 & 0 & 1 & 1 & | & 0 & 0 & 1 \end{bmatrix}$$

$$\underbrace{\qquad}_{P^T} \quad \underbrace{\quad}_{I_3}$$

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 0 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 ith bit of C would lead to a syndrome vector that would be identical to the ith 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 ith received bit if S matches with the ith 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 the weigth of a code and its minimum distance. The weight of a code is defined as the number of non-zero components in the codeword. The distance between two code vectors is the number of the components in which they differ, while the minimum distance 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. 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

TABLE I

Codewords' Weight

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

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 :

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

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. Hamming Codes

From the equation just written, it follows that linear block codes capable of correcting single errors must have $d_{min} = 3$. Such codes are easy to construct. Each row in H^T has (n-k) entries, and each entry can be either a 0 or a 1. We have therefore 2^{n-k} distinct rows of (n-k) entries, out of which we can select $2^{n-k} - 1$ distinct rows of H^T (the row of 0's is the only one that we cannot use). Since the matrix H^T has n rows, for all of them to be distinct, we need: $2^{n-k} - 1 \geq n$. In other words, the number of parity bits in this (n,k) code satisfies the inequality $(n-k) \geq \log_2 (n+1)$. So, given a message size k , we can determine the minimum size n for the codewords from $n \geq k + \log_2 (n + 1)$, where n has to be an integer.

Assume that we have to design a linear block code with a minimum distance of three and a message block size of eight bits. We proceed this way : from the inequality above ,we have $n \geq 8 + \log_2(n+1)$. The smallest value of n that satisfies this condition is n = 12. Thus we need a (12,8) block code. The transpose of the parity check matrix H will have a size 12 by 4. The first 8 rows are arbitrarily chosen, with the restrictions that no row is identically zero, and all rows are distinct. A possible choice for H^T is:

$$H^T = \left[\begin{array}{c} P \\ \hline I_{n-k} \end{array}\right] = \left[\begin{array}{c} 1100 \\ 0110 \\ 0011 \\ 1001 \\ 1010 \\ 0101 \\ 1110 \\ 0111 \\ \hline 1000 \\ 0100 \\ 0010 \\ 0001 \end{array}\right]$$

The generator matrix is:

$$G = \left[\begin{array}{c|c} I_k & P \end{array}\right] = \left[\begin{array}{cc} 10000000 & 1100 \\ 01000000 & 0110 \\ 00100000 & 0011 \\ 00010000 & 1001 \\ 00001000 & 1010 \\ 00000100 & 0101 \\ 00000010 & 1110 \\ 00000001 & 0111 \end{array}\right]$$

The receiver forms the syndrome $S = RH^T$ and accepts the received code vector if $S=0$. Single errors are always corrected; double errors can be detected but cannot be corrected; multiple errors will result ,in general, in incorrect decoding. The efficiency of block codes with minimum distance three (Hamming Codes) improves as the message block size is increased. We can verify that for k=64, a minimum distance three code with efficiency 0.90 does exist, and that for k=256, a code with efficiency 0.97 can be found.

We have seen in the above that the decoding operation consists in finding a codeword C that is the closest to the received word R. This requires storing 2^k codewords and comparing R with each of them. Since each codeword is of length n, the storage required contains $n2^k$ bits. Even for modest size of k and n, the storage requirement is excessive, and the processing requirement also becomes soon impractical. There is ,however, a way of reducing considerably both these requirements, as we will see next, in Section 4.

4. Table lookup and standard array

We will consider now a decoding scheme (table lookout) that requires the storage of 2^{n-k} syndrome vectors of length n-k bits and the 2^{n-k} n-bit error patterns corresponding to these syndromes. Thus, the storage required will be $2^{n-k} \times (2n - k)$ bits. For high-efficiency codes, $2n-k \approx n$; hence, the storage required will be of the order of $n2^{n-k}$ bits, a much smaller capacity than $n2^k$ bits required if the table look-out approach is not used. Even with this reduction, though, the decoding scheme of block codes may be impractical. For instance, for a (200,175) code, the storage requirement is of 7.6 Gigabits!!

The table look-out approach works in the following way. Suppose that a (n,k) linear code is used for error correcting purposes. Let C_1, C_2, ..., C_{2^k} be the code vectors of C. If R is the received vector, it can be any one of the 2^k n-tuples that are valid codewords. The decoder can perform the task of associating R with one of the n-tuples, by partitioning the set of 2^n n-tuples as shown in Table II. The code vectors

TABLE II
Example of standard array
for a (n,k) linear block
code

C_1	C_2	C_3 .... C_{2^k}
E_2	C_2+E_2	$C_3 + E_2$ ....$C_{2^k} +E_2$
E_3	C_2+E_3	$C_3 + E_3$ .... $C_{2^k} +E_3$
.	.	. .
.	.	. .
$E_{2^{n-k}}$	$C_2+E_{2^{n-k}}$	$C_3+E_{2^{n-k}}$ .....$C_{2^k} + E_{2^{n-k}}$

C_1, C_2,...C_{2^k} are placed in the first row, with the code vector of all zeros appearing in the leftmost position. The first element in the second row E_2 is any one of $(2^n - 2^k)$ n-tuples not appearing in the first row. Once E_2 is chosen, the second row is completed by adding E_2 to the codewords, as shown in Table II. Once that the second row has been completed, an unused n-tuple E_3 is chosen to begin the third row and the sum of $C_i + E_3$ (with $i=1,2, ...2^k$) are placed in the third row. The process is continued until all the 2^n n-tuples are used. The resulting array is called the standard array for the code and it consists of 2^k columns that are disjoint. Each column has 2^{n-k} n-tuples with the topmost n-tuple as a code vector. The jth column is the partition T_j that will be used for decoding. The rows of the standard array are called co-sets and the first element in each row is called a co-set leader. The standard array has the property that each element is distinct and hence the columns are disjoint; furthermore, if the error pattern coincides with a co-set leader, the received error is correctly decoded. If it does not, then an incorrect decoding will result. Thus the co-set leaders are called correctable error patterns.

In order to minimize the probability of incorrect decoding, the 2^{n-k} co-set leaders are chosen to be the error patterns that are most likely to occur for a given channel. If E_i and E_j are two error patterns with weights W_i and W_j, then for a channel in which only random errors are occurring, E_i is more likely to occur than E_j, if $W_j > W_i$. Therefore, when constructing a standard array, the co-set leader should be chosen as the vector with minimum weight from the remaining available vectors. A standard array has an important property that leads to a simpler decoding process : all the 2^k n-tuples of a co-set have the same syndrome and the syndromes of different co-sets are different. The one-to-one correspondence between a co-set leader (correctable error pattern) and a syndrome leads to the following procedure for decoding: a) compute the syndrome RH^T for the received vector R. Let $RH^T = S$.; b) locate the co-set leader E_i that has a syndrome $E_i H^T = S$. Then, E_i is assumed to be the error pattern caused by the noisy channel; c) the code vector C is obtained from R by $C = R + E_i$. Since the most probable error patterns have been chosen as the co-set leaders, this scheme will correct the 2^{n-k} most likely error patterns introduced by the channel.

C. CYCLIC CODES (*)

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

$$C = [C_{n-1}, \quad C_{n-2}, \quad \cdots\cdots \quad C_0]$$

is a code word, so are

$$[C_{n-2}, \quad C_{n-3}, \cdots\cdots\cdots C_{n-1}]$$
$$\vdots \quad\quad \vdots \quad\quad\quad \vdots$$
$$[C_0, \quad C_{n-1} \cdots\cdots\cdots C_1]$$

Example for n=3
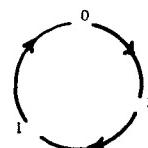
This formulation of cyclic codes suggests treating the elements of each code word as coefficient of a polynomial of degree n-1. With this convention, the codeword C can be represented with the polynomial

$$C(x) = C_{n-1} x^{n-1} + C_{n-2} x^{n-2} + \cdots\cdots + C_1 x + C_0.$$

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 $x^i C(x)$ 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 $x^i C(x)$-modulo-(x^n+1) are, in fact,

$$[C_{n-i-1}, \quad C_{n-i-2}, \cdots\cdots C_1, C_0, C_{n-1} \cdots C_{n-i}]$$

Several interesting properties of the generator matrix of a cyclic code can be deduced from the above definition. We will illustrate them by via of an example, considering again the (15,11) code introduced in section B.1. This code is a cyclic code and its generator matrix, written in polynomial form, is as follows:

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

where $g(x) = x^{4}+x+1$. Encoding a k-bit data block by multiplying it by the generator matrix G is equivalent to the following polynomial operation.The polynomial representation of the information block, denoted by d(x) has a degree less than k; therefore $x^{n-k}d(x)$ has a degree less than n. Also

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

where q(x) has a degree less than k and r(x) has a degree less than n-k, which is the degree of g(x). Thus, the polynomial $x^{n-k}d(x) + r(x)$ is divisible by g(x) and is a codeword in the code generated by g(x). This word consists of unaltered k-bit information block followed by n-k linear combinations of the information bits. It must be identical to the word formed by multiplying the k-place vector D by the matrix G since, as we have seen, there is only one code polynomial of degree n-k in a cyclic (n,k) code. The syndrome associated with a given n-tuple can be calculated in a similar manner. Let $e(x) = e_d(x) + e_p(x)$ be the polynomial re-

presentation of such n-tuple. The syndrome is obtained by encoding the information section of the n-tuple and by adding the resulting check bits to the corresponding bits of the parity section of the n-tuple. That is, the syndrome s(x) is given by $s(x) = e_p(x) + r(x)$, where r(x) is the remainder obtained by dividing

$e_d(x)$ by g(x). Since $e_p(x)$ has a degree less than g(x), this is identical to the remainder obtained by dividing $e(x) = e_d(x) + e_p(x)$ by g(x). Since g(x) divides every code word c(x), the syndrome is identical to that of e(x)+ c(x). Encoding and syndrome calculation can be mechanized in a remarkably simple manner.

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 Nxn digits, where N is the size of the shift register, and n is the number of the commutator segments.

3. The decoding operation

It helps,in understanding the decoding operation,to make use of the code tree shown in Figure 3, which applies to the convolutional encoder shown in Figure 2. The starting point of the code tree is at left and

it corresponds to the situation before the occurrence of the ith message bit d_i. We assume that the message bits d_{i-1} and d_{i-2} are zero. The paths shown in the code tree are generated by using the rule that we shall diverge upward from a node of the tree when the input bit is 0. The starting node (A) corresponds to $d_{i-1}=$ 0 and $d_{i-2}=0$. Then, if $d_i=0$, we move upward from the initial node and the coder output is 000. Nodes 1,2, 3 and 4 can be considered as starting nodes for d_{i+2}, with $d_i d_{i+1} = 00$, 01, 10, and 11 respectively. For any given starting node, the first message bit influences the code blocks generated from the starting node and the two succeeding nodes. Thus each message bit will have an effect on nine code digits and there are eight distinct 9-bit code blocks associated with each starting node.

The code tree can be used as follows in a decoding operation called the exhaustive tree-search method of decoding. In the absence of noise, the codewords will be received as transmitted. In this case, it is simple matter to reconstruct the transmitted bit sequence. We simply follow the codeword through the code tree n bits at a time (n=3 for the example considered here). The transmitted message is then reconstructed from the path taken through the tree. The presence of noise introduces tranmsission errors; in this case the following procedure can be used for reconstructing the transmitted codeword. Consider the ith message digit d_i that has an influence on nN bits in the codeword. For the example shown in Figures 2, 3 and 4, N=3 and n=3, so that d_i has an effect on nine code digits. Hence, in order to deduce d_i, there is no point in examining the codeword beyond the nine-digit block that has been influenced by d_i. On the other hand, we would lose some of the benefits of the code, if we would use less than nine bits of the received codeword. If we assume 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. We compare the 9-bit received code block we are examining with the eight valid code blocks, and discover the valid code block closest to the received code block (closest in Hamming distance). If this valid code block corresponds to an upward path from the starting node on the code tree, then d_i is decoded as 0; otherwise, it is decoded as 1. After d_i is decoded, we use the decoded values of d_{i-1} and d_i to define a starting node for decoding d_{i+1}. This procedure is repeated until the entire message sequence is decoded.

The exhaustive tree-search method of decoding convolutional codes, briefly illustrated above, becomes impractical as N becomes large, since the decoder has to examine 2^N branch sections of the code tree. Another decoding scheme avoids this lengthy process : it is called the Sequential Decoding Scheme. In sequential decoding, at the arrival of a n-bit code block, the encoder compares these bits with the code blocks associated with the two branches diverging from the starting node. The encoder follows an upward or downward path (hence it decodes a message bit as 0 or 1, respectively) in the code tree, depending on which of the code blocks exhibit the fewest discrepancies with the received bits.

If a received code block contains transmission errors, then the decoder might make an error and start out on a wrong path in the code tree. In such a case, the entire continuation of the path taken by the encoder will be in error. If the decoder keeps a running record of the total number of discrepancies between the received code bits and the code bits encountered along its path, then the likelihood is great that, after having made the wrong turn at some node, the total number of errors will grow more rapidly than in the case that the decoder follows the correct path. The decoder can be programmed to respond to such situations by retracing its path to the node at which an apparent error has been made, and then taking an alternate branch out of that node. In this way, the decoder will eventually find a path through N nodes. When such a path is found, the decoder decides about the first message bit. Similarly, the second message bit is then determined on the basis of the path searched out by the decoder, again N branches long.

The decoder begins retracing when the number of accumulated errors exceeds a threshold, as shown in Figure 4. Since every branch in the codetree 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 $(P_e)(n)(j)$ even when the correct path is being followed. The number of bit errors accumulated will oscillate about E(j) (see Figure 4), if the encoder 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 (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.

4. Recapitulation of the properties of convolutional codes

Although the theory of convolutional codes is not as well developed as that of block codes, several practical convolutional encoders/decoders have been built, for use in various applications. Convolutional codes have many of the basic properties of block codes : a) they can be encoded using simple shift registers and modulo-2 adders; b) while decoding is difficult, several classes of codes have been found for which the amount of equipment required for decoding is not excessive; c) practical convolutional codes exist that are capable of correcting random and burst types of errors. Their advantages over the block codes are several: 1. since they operate on smaller blocks of data, the decoding delay is small; 2. smaller amount of storage hardware is required; 3. loss of synchronization is not as serious a problem as with block codes. In summary, convolutional codes, though not as well developed and more difficult to analyze than block codes, are competitive with block codes in many applications.

E. GETTING CLOSER TO THE ENGINEERING ASPECTS OF CODING [*]

We will illustrate in this section the mechanization of some of the coding schemes discussed in the course of the Lecture. We will specialize our treatment to the encoding of cyclic codes, touching base also with syndrome calculation, error detection, and error correction. Some mathematics is still here, but equipment block diagrams make their appearance.

We recapitulate that encoding consists of appending a set of parity checks to the data block before transmission; that the syndrome is simply the modulo-2 sum of these checks after transmission and the check bits calculated on the received data block at the decoder; that error detection consists of determining whether or not the syndrome is the all-zero (n-k)-tuple (if it is not, errors have occurred); that these errors can usually be corrected by adding to the received word the error pattern associated with the syndrome; that this association of error pattern with syndrome is essentially the decoding operation.

As an example of encoding mechanization, let's consider again the equation that was introduced in *Section C:*

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

where $g(x)$ denotes the generator polynomial of the code and $x^{n-k}d(x) + r(x)$ is the n-bit cyclic codeword into which a k-bit data block $d(x)$ has to be encoded. Since the data bits are transmitted without alteration, encoding consists simply of determining $r(x)$. The necessary division can be performed with the circuit in Figure 5. In the Figure, the meaning of the symbols is as follows : 1) ▢ denotes a single binary shift register stage ; 2) ⊕ denotes an EXCLUSIVE -OR circuit or modulo-2 adder; 3) (f_i) simply denotes a connection, or the lack of one, depending on whether f_i is 1 or 0, respectively.

Encoding is accomplished as follows: the data polynomial $d(x)$ is shifted into the feedback shift register, as shown in Figure 5. After n-k shifts, the register contains the n-k high-order terms of $x^{n-k}d(x)$. After the last data bit has been fed into the register, the feedback connection is broken, by moving the switch from position D (data) to position P (parity), and the content of the register is shifted to the channel. That this circuit actually divides $x^{n-k}d(x)$ by $g(x)$ to produce the (n-k)-bit remainder $r(x)$, can be seen by considering the following example. Let's use again the (15,11) code already introduced in this Lecture. The generator polynomial of this code is $g(x) = x^4 + x + 1$. Suppose that the data polynomial $d(x) = x^{10}+ x^6 + x^3 + x$ is to be encoded. The remainder $r(x)$ can be obtained by dividing $x^4 d(x)$ by $g(x)$. Thus:

Let's consider now the encoder for this code, shown on Figure 6. The first four shifts merely serve to load the first four terms of the dividend into the register. Table III shows the contents of the register as the encoding process is performed. Also shown are the dividends in the above division which correspond to the register contents at various stages of the encoding process.

TABLE III
Comparison of Register Contents & Dividends

Shift number	Register contents	Bits in data source		Corresponding dividends		
0	1000	1001010	x^{14}	x^{10}	x^7	x^4
1	0010	001010				
2	0100	01010				
3	1000	1010	x^{11}	x^7	x^4	
4	0010	010				
5	0100	10				
6	1001	0	x^8	x^5		
7	0001					
8	0010					
9	0100					
10	1000		x^4			
11	0011		$x + 1$			

After the first four shifts, the four high-order bits of $x^4 d(x)$ occupy the register. On the next shift, a 1, corresponding to the quotient term x^{10}, is emitted. Thus 1s are added into the stages following the feedback connections, and the register contents then correspond to the new dividend, obtained by adding $x^{10} g(x)$ to the first dividend. The key point is that when the register output is 1, the dividend is modified exactly as in the division process; likewise when it is 0, nothing is done in either case.

Since the first n-k shifts serve only to load the register, the encoder of Figure 5 (a) can be modified into the configuration of Figure 5 (b). In the latter figure, those load shifts are not necessary. The feedback information, which is simply the sum of the high-order bit of the dividend and the various past multiples of $g(x)$, is the same in either case. The contents of the register differ until after the nth shift of Figure 5 (a) (the kth shift of Figure 5 (b)). At that time, the dividend has been shifted completely through the encoder, and thus both registers contain $r(x)$.

Because it requires n-k fewer shifts, the circuit of Figure 5 (b) is usually preferable to that of Figure 5 (a) for encoding serial data. Encoding is performed here by shifting the data sequence simultaneously into the register and onto the line. After the last data bit has been shifted out, the switch is thrown from D to P, and the n-k parity bits are shifted out. This leaves the register empty and ready to accept the first data bit of the next word, immediately.

Syndrome calculation differs from encoding only in that the received parity bits must be added to the checks calculated on the received data. The circuit of Figure 5 (a) performs this addition automatically, if the received parity bits are shifted in immediately after the data. The circuit in Figure 6 yields the same result : that is, the output is the syndrome.

Concerning the error detection, this function can be implemented simply by adding a flip-flop to the output of the syndrome generator. If one or more 1s appear in the syndrome, the flip-flop sets and an error has bben detected; otherwise, the received n-tuple is a code word. Error correction is simply an association of a syndrome with an error pattern. This association can be visualized conveniently in terms of the standard array. A given code has exactly 2^{n-k} distinct co-sets, including the code itself. Only one error pattern in each co-set is correctable. Thus, by choosing the most probable error pattern in each co-set as the co-set leader, the probability of erroneous decoding is minimized.

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 retransmission (ARQ). Forward error correction has waited considerably before gaining even partial acceptance, essentially 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 signalling 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 condiderable 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, hand-in-hand with the progress in microprocessor technology, and with the ever greater easiness by which complex circuitry is mechanized.

It is hoped that this introductory lecture did strike a balance in pointing out both the potential usefulness of codes in HF communications, and the complexity of the required circuitry. 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 confidence coding problems of the HF communications practice.

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Terminal A

Terminal B

| Source | Source Encoder | Channel Encoder | Modulator | Channel | Demodula- tor | Channel Decoder | Source Decoder | User |

Noise
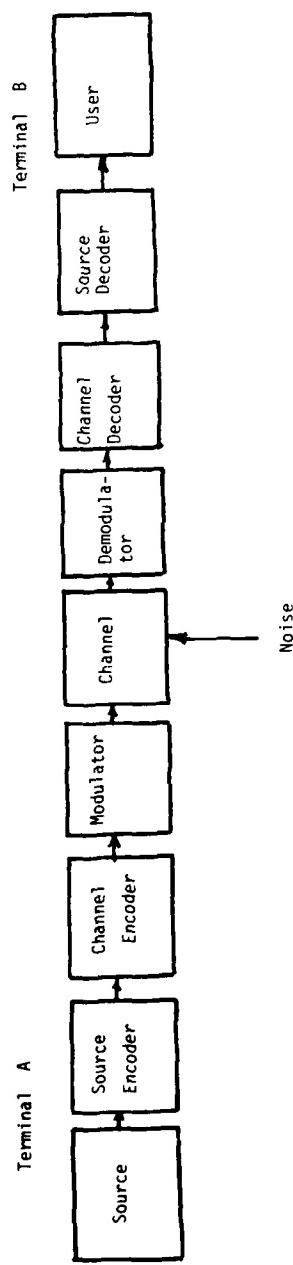
Fig.1   Block diagram of a communications link that uses coded transmission.
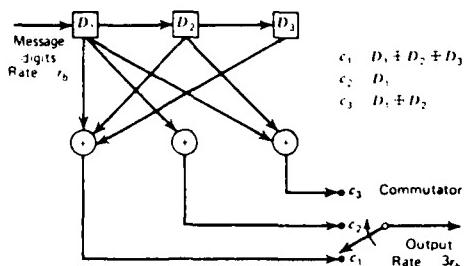
Fig.2 An example of convolutional encoder, with N = 3, n = 3, k = 1.
(from Shanmugan (1979), [4])
(by permission of the Publisher John Wiley & Sons)
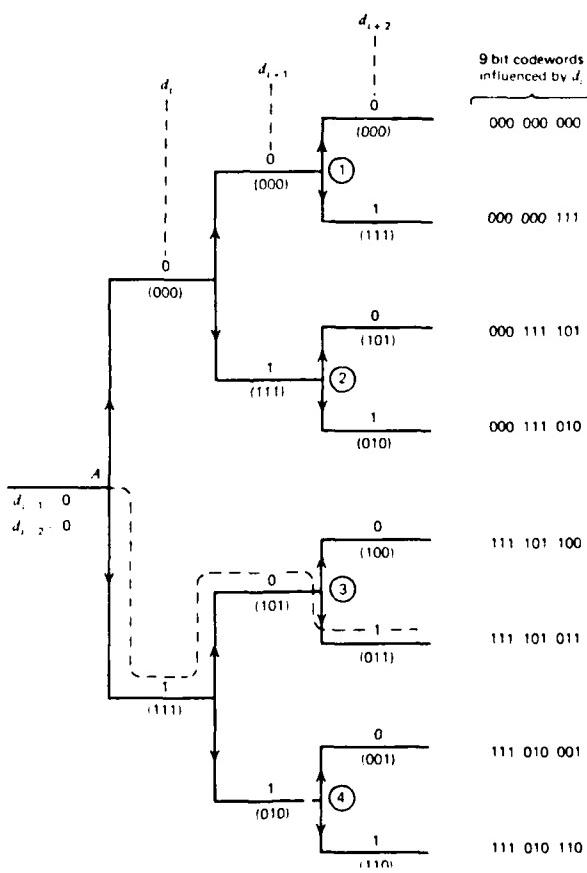
Fig.3 Code tree for the convolutional encoder of Figure 2
(from Shanmugan (1979), [4])
(by permission of the Publisher John Wiley & Sons)
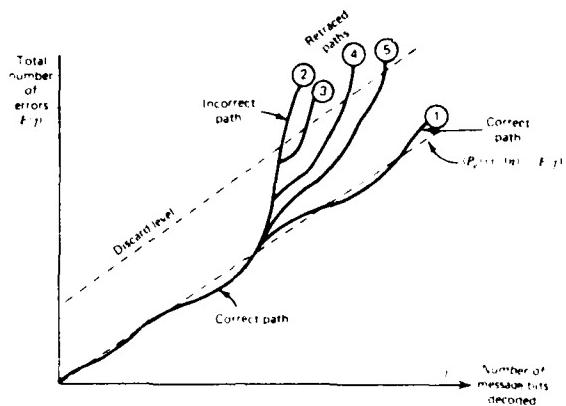
Fig.4   Threshold setting in sequential decoding (P_e = probability that a received bit is in error)
(from Shanmugan (1979), [4]) (by permission of the Publisher John Wiley & Sons)

(a)

(b)
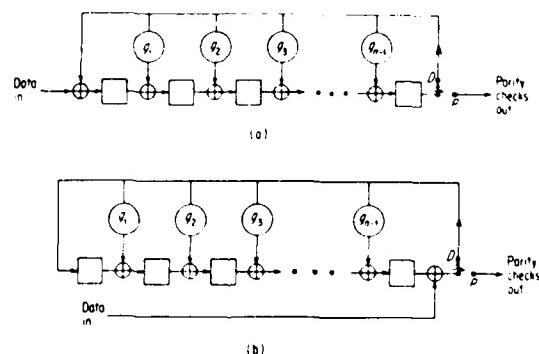
Fig.5   Two (n-k) stage encoding circuits for the cyclic code generated by $g(x) = x^n + g_{n-1} x^{n-1} + \ldots$
$g_2 x^2 + g_1 x + 1$. (from Lucky et. al., (1968), [5]) (by permission of the Publisher McGraw-Hill Book Co.)

(a)

(b)
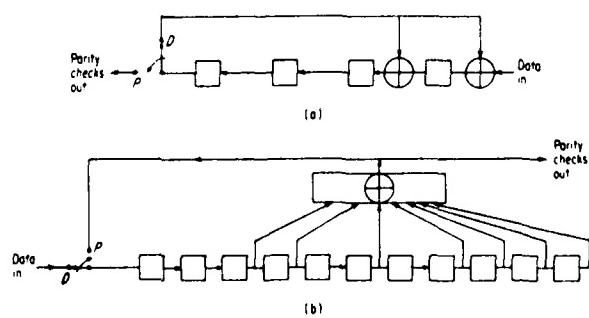
Fig.6   Encoders for the (15,11) code. (a) An n-k = 4-stage encoder (b) An k = 11-stage encoder
(from Lucky et. al., (1968), [5]) (by permission of the Publisher McGraw-Hill Book Co.)

MODERN HF COMMUNICATIONS, MODULATION AND CODING

by

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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

$$\sigma_t = \frac{1}{2*t_d} \ Hz$$

is a measure of the fading rate or bandwidth of the random channel. The quantity σ_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

$$\sigma_f = \frac{1}{2\pi f_d} \text{ seconds}$$

defines the extent of the multipath delay. The quantity σ_f is often referred to as the multipath delay spread as it is a measure of the width of the received process in the time domain when a single impulse function is transmitted through the channel.

The range of values for these spread factors for HF communication are

$$\sigma_t \sim 0.1 - 1.0 \text{ Hz}$$

$$\sigma_f \sim 0.5 - 5 \text{ milliseconds}$$

where the symbol $\sim$ denotes "on the order of".

The spreads can be defined precisely as moments of spectra in a channel model [1] which assumes Wide Sense Stationarity (WSS) in the time variable and Uncorrelated Scattering (US) in a multipath delay variable. This WSSUS model and the assumption of Gaussian statistics for $H(f,t)$ provide a statistical description in terms of a single two-dimensional correlation function of the random process $H(f,t)$.

This characterization has been quite useful and accurate for a variety of radio link applications. However the stationarity and Gaussian assumptions are not necessary for the utilization of adaptive signal processing techniques on these channels. What is necessary is first that sufficient time exist to "learn" the channel characteristics before they change and second, that decorrelated portions of the frequency band be excited such that a diversity effect can be realized. These conditions are reflected in the following two relationships in terms of the previously defined channel factors, the data rate R, and the bandwidth B:

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

$$B(Hz) \gtrsim f_d(Hz) \quad \text{Diversity Constraint.}$$

where the symbol $\gtrsim$ denotes "on the order of or greater than".

The learning constraint insures that sufficient signal-to-noise ratio (SNR) exists for reliable communication at rate R over the channel. Clearly if $R \sim \sigma_t$, the channel would change before significant energy for measurement purposes could be collected. When $R \gg \sigma_t$, the received data symbols can be viewed as the result of a channel sounding signal and appropriate processing can generate estimates of the channel character during that particular stationary epoch. The signal processing techniques in an adaptive receiver do not necessarily need to measure the channel directly in the optimization of the receiver but the requirements on learning are approximately the same. If only information symbols are used in the sounding signal, the learning mode is referred to as Decision-Directed. When digital symbols known to both the transmitter and receiver are employed, the learning mode is called Reference-Directed. An important advantage of digital systems is that in many adaptive communications applications, adaptation of the receiver with no wasted power for sounding signals can be accomplished using the decision-directed mode. This is possible in digital systems because of the finite number of parameters or levels in the transmitted source symbols and the high likelihood that receiver decisions are correct.

Diversity in fading applications is used to provide redundant communications channels so that when some of the channels fade, communication will still be possible over the others that are not in a fade. Some of the forms for diversity employed in HF systems are space using multiple antennas, polarization, frequency, and time. These diversity techniques are sometimes called explicit diversity because of their externally visible nature. An alternate form of diversity is termed implicit diversity because the channel itself provides redundancy. In order to capitalize on this implicit diversity for added protection, proper receiver techniques have to be employed to correctly assess and combine the redundant information. The potential for implicit frequency diversity arises because different parts of the frequency band fade independently. Thus, while one section of the band may be in a deep fade the remainder can be used for reliable communications. However, if the transmitted bandwidth B is small compared to the frequency decorrelation interval f_d, the entire band will fade and no implicit diversity can result. Thus, the second constraint $B \gtrsim f_d$ must be met if an implicity diversity gain is to be realized. In diversity systems a little decorrelation between alternate signal paths can provide significant diversity gain. Thus it is not necessary for $B \gg f_d$ in order to realize implicit frequency diversity gain although the implicit diversity gain clearly increases with the ratio B/f_d. Note that the condition $R \ll B \gtrsim f_d$ does not preclude the use of implicit diversity because a bandwidth expansion technique can be used in the modulation process to spread the transmitted information over the available bandwidth B. We shall distinguish between these low data rate and high data rate conditions because the appropriate receiver structures take on somewhat different forms.

The implicit diversity effect described here results from decorrelation in the frequency domain in a slow fading $(R \gg \sigma_t)$ application. This implicit frequency diversity can in some circumstances be supplemented by an implicit time diversity ef-

fect which results from decorrelation in the time domain. In fast fading applications ($R \geq \sigma_t$) redundant symbols in a coding scheme can be used to provide time diversity provided the code word spans more than one fade epoch. In the slow fading application typical of HF communication this condition of spanning the fade epoch can be realized by interleaving the code words to provide large time gaps between successive symbols in a particular code word. The interleaving process requires the introduction of signal delay longer than the time decorrelation separation t_d. In many practical applications which require transmission of digitized speech, the required time delay is unsatisfactorily long for two-way speech communication. For these reasons there is more emphasis on implicit frequency diversity techniques in practical systems. The receiver structures to be discussed next are applicable to situations where the implicit frequency diversity applies. In a subsequent section we discuss error correction coding techniques which combined with interleaving exploit the implicit time diversity of the channel.

3.   ADAPTIVE RECEIVER STRUCTURES

We consider a general modulation system which accepts binary input data at a rate R and converts each group of $\log_2 M$ bits into a sample value which can take on one of M values. This M-ary discrete sample $\{a_k\}$ is to be communicated over the channel using a modulation technique which forms a one-to-one correspondence between the sample a_k and the waveform structure of a transmitted pulse. Independent modulation of quadrature carrier signals (i.e., $\sin 2\pi f_0 t$ and $\cos 2\pi f_0 t$, f_0 = carrier frequency) is included in this class. An important example with optimum detection properties is Quadrature Phase Shift Keying (QPSK) which transmits the sample set $\{a_k = \pm 1 \pm j\}$ by changing the sign of quadrature carrier pulses in accordance with the sign of the real and imaginary parts of the source sequence $\{a_k\}$.

3.1   Receivers for Channels with Negligible Intersymbol Interference

If each sample a_k can be one of M possible waveforms (M=4 for QPSK), the transmitted data rate is

$$R = (\log_2(M))/T$$

where $1/T$ is the transmitted symbol rate.

Many HF channel applications are signal-to-noise ratio limited rather than bandwidth limited. In order to maximize signal detectability, only a few waveform choices are usually employed in these applications. When the symbol period T is much greater than the total width of the multipath dispersion of the channel, only a small portion of adjacent symbols interfere with the detection of a particular symbol. For the slow fading application, the diversity constraint requires that the signal bandwidth B be on the order of or larger than the frequency decorrelation interval f_d. Conditions for negligible intersymbol interference (ISI) and adaptive processing to obtain implicit diversity are then

$$T \gg 2\pi\sigma_t = \frac{1}{f_d}$$

$$B \geq f_d .$$

When the number of waveform choices M is small, these conditions imply a low data rate system relative to the available bandwidth, i.e., a bandwidth expansion system. The low data rate condition for implicit frequency diversity can be expressed in terms of the data rate and bandwidth as

$$R \ll B \log_2(M)$$

In the absence of intersymbol interference, it is well known [2] that the optimum detection scheme contains a noise filter and a filter matched to the received pulse shape. The optimum noise filter has a transfer function equal to the reciprocal of the noise power spectrum K(f). When the additive noise at the receiver input is white, i.e., its spectrum is flat over the frequency band of interest, the noise filter can be omitted in the optimum receiver. The optimum receiver for a fixed channel transfer function H(f) then contains a cascade filter with component transfer functions

$$R(f) = \frac{1}{K(f)} H^*(f)$$

$$\text{Noise} \quad \text{Matched}$$
$$\text{Filter} \quad \text{Filter}$$

where the * denotes complex conjugation. In practical applications, signal delay must be introduced in order to make these filters realizable.

In general K and H change with time and the adaptive receiver must track variations. The tapped-delay-line (TDL) filter is an important filter structure for such channel tracking applications. The TDL filter shown in Figure 1 consists of a tapped delay line with signal multiplications by the tap weight w_i for each tap. For a band-

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 σ_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

$$R(b/s) \gg \sigma_t \ (Hz)$$

introduced earlier. If the signal rate R from which adaptation is being accomplished is not much greater than the channel rate of change σ_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 σ_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

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

or in terms of the multipath spread

$$T(sec) \leq 2\pi\sigma_f (sec) \quad .$$

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 2σ 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	
Harris (HRS) Modem:	1 dB	SERIAL TONE
USC-10:	7 dB	
ACO-6:	7 dB	PARALLEL TONE
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.

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

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

$$R_0 = 1 - \log_2(1 + Z) \qquad (2)$$

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,

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

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

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

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

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

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

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

For uncoded coherent phase shift keying (CPSK) operating under the same Rayleigh fading conditions, one has for large E_b/N_0,

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

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 non-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
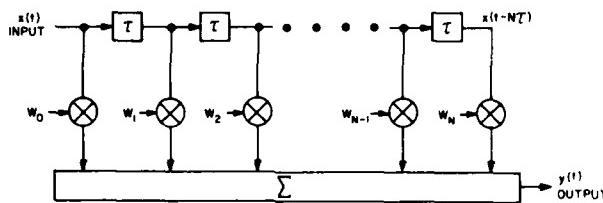
FIGURE 1   TAPPED-DELAY LINE (TDL) FILTER
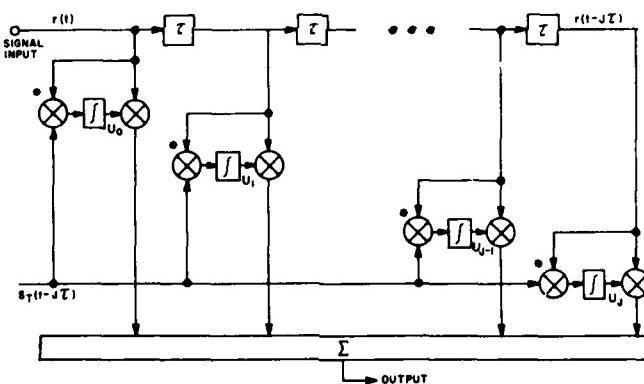
FIGURE 2   RAKE FILTER, COHERENT SYSTEM
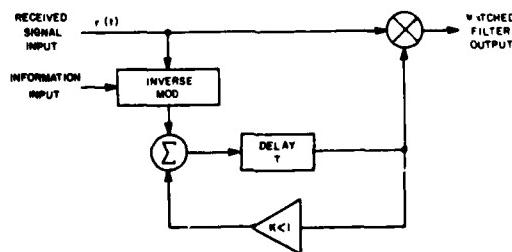
FIGURE 3   CORRELATION FILTER FOR PAM SYSTEM, COHERENT SYSTEM
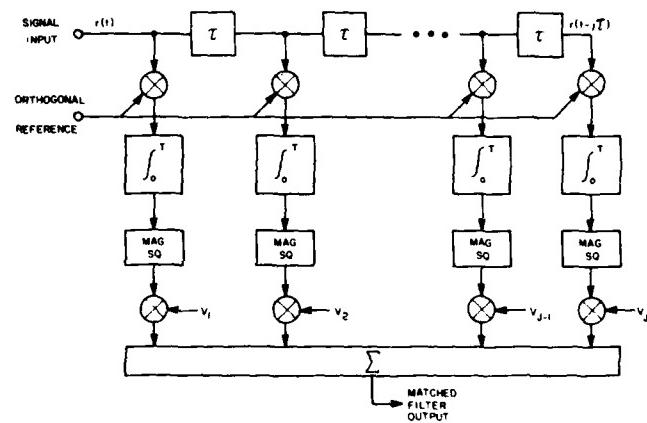
FIGURE 4   ORTHOGONAL SIGNALING MATCHED FILTER, INCOHERENT DETECTION
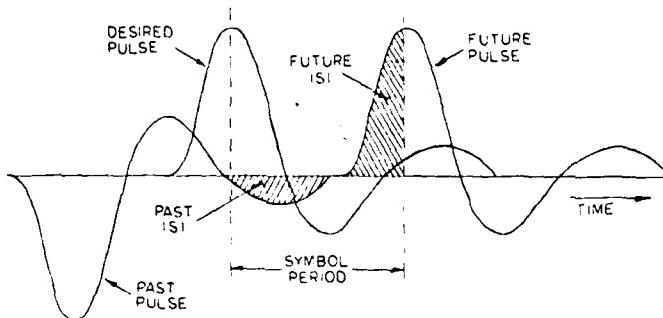
FIGURE 5   RECEIVED PULSE SEQUENCY AFTER CHANNEL FILTERING
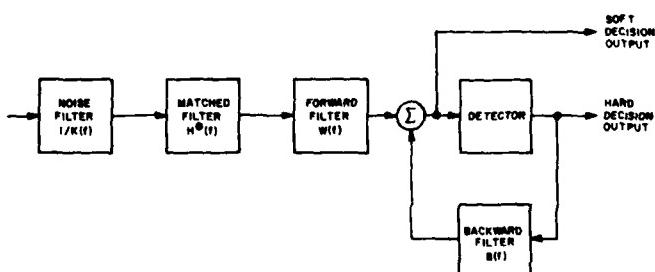
FIGURE 6   DECISION-FEEDBACK EQUALIZER RECEIVER

FIGURE 7   MLSE RECEIVER

FLAT FADING

USC-10, ACQ-6, MX-190 1 0 HZ
USC-10, ACQ-6, MX-190 0 2 HZ

AVERAGE BIT ERROR PROBABILITY

1 0 HZ
0 2 HZ
S/G
S/G
HRS

SIG 2900 b/s
HRS
ACQ-6
USC-10   2400 b/s
MX-190

E_{pb}/N_0 (dB)

FIGURE 8   FLAT FADING

DUAL FADING PATHS
1 MSEC DELAY

AVERAGE BIT ERROR PROBABILITY

1 0 HZ
USC-10
ACQ-6
USC-10/ACQ-6 0 2 HZ
2 0 HZ
S/G
HRS
MX-190
1 0 HZ
1 0 HZ
S/G
HRS
0 2 HZ

SIG 2900 b/s
HRS
ACQ-6
USC-10   2400 b/s
MX-190

E_{pb}/N_0 (dB)
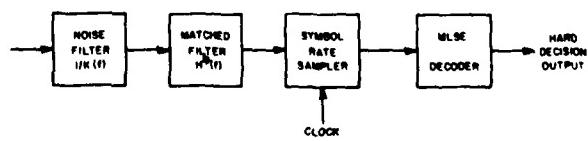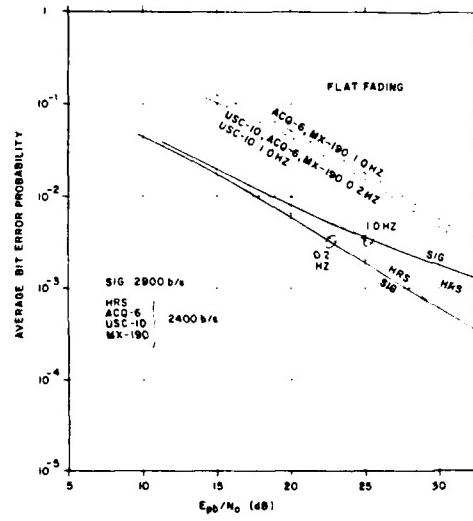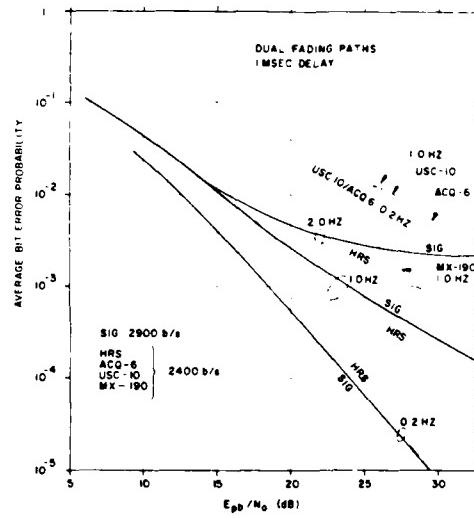
FIGURE 9   WATTERSON TEST CHANNEL

R_0

bits/CHANNEL USAGE

NON FADING

$R_0 = 1 - LOG_2 (1 + e^{-rE_b/N_0})$

RAYLEIGH FADING, IDEAL INTERLEAVING

$R_0 = 1 - LOG_2 (1 + (1 + rE_b/N_0)^{-1})$

rE_b/N_0 (dB)
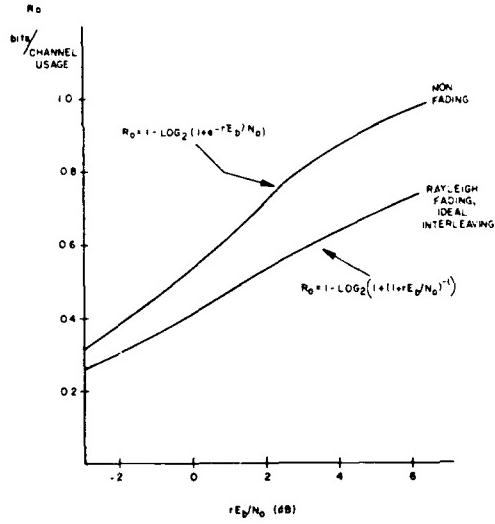
FIGURE 10   LOSS IN DETECTION
EFFICIENCY DUE TO FADING
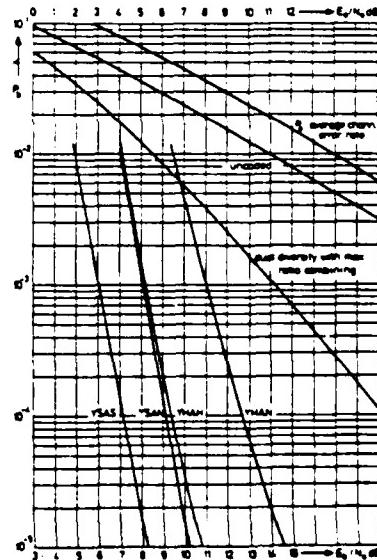
FIGURE 11 BIT ERROR PROBABILITY P_b
(UPPER BOUND) AS A FUNCTION OF
AVERAGE E_b/N_0.

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

YHAH (y hard, a hard (BINARY QUANTI-
                        ZATION)

FROM HAGENAUER [21]

EQUIPMENT: ANTENNA SYSTEMS
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CANADA

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 communication circuit or service area. Sources for obtaining or computing radiation pattern informantion are outlined. Comparisons are presented 'etween 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 service 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. Souces 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 angle 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 inorder to select one that meets the particular needs. Several reports provide graphical information (Dept. of the Army, 1950; Thomas and DuCharme 197'' on the radiation pattern as a function of elevation angle at the centre of the main lobe. In these reports, patterns are presented for antennas typically used for HF communications. 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 tothe power radiated from the antenna. The antenna gain does not account for impedance mismatch losses. Vertical radiation patterns are shown for

a) horizontal half-wave dipole
b) rhombics, types A, B, C, D, and E
c) half rhombic
d) inverted L
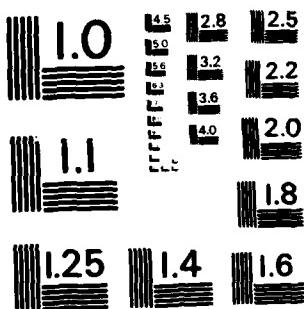
MICROCOPY RESOLUTION TEST CHART

NATIONAL BUREAU OF STANDARDS – 1963 – A

e) sloping V
f) horizontal log periodic
g) vertical antenna

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). 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.

The type of ground generally has a limited effect on the main lobes of the pattern for horizontally polarized antenna. However for short vertically polarized antenna, the type of ground can significantly alter the efficiency of the antenna and its power gain. The effect of ground on the antenna radiation 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 diploe	rhombic	rhombic
inverted L	half wave dipole	log periodic
vertical	vertical	beverage
long wire	inverted L	
delta		

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.

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 V	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 communication conditions expected from 1981 to 1985. The Sloping V 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. 4. 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 (Petrie, 1978; Petrie 1979). 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 measureemnt 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. 5,6,and 7. 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 prdictable 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 be incorporated into the desired communication signal. The 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 prot .tion 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 more 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 level from a one meter length rod is generally greater than the amplifier 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.

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WINNIPEG-BAKER LAKE

Fig. 1

Elevation Angle (degrees)

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

POWER GAIN RELATIVE TO ISOTROPIC (dB)

Fig. 2
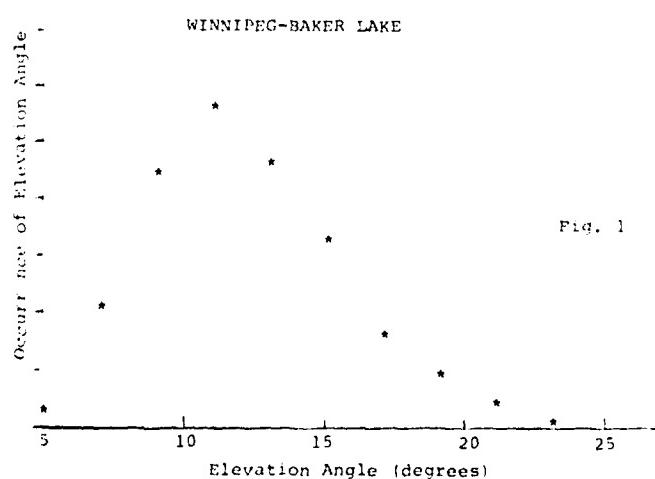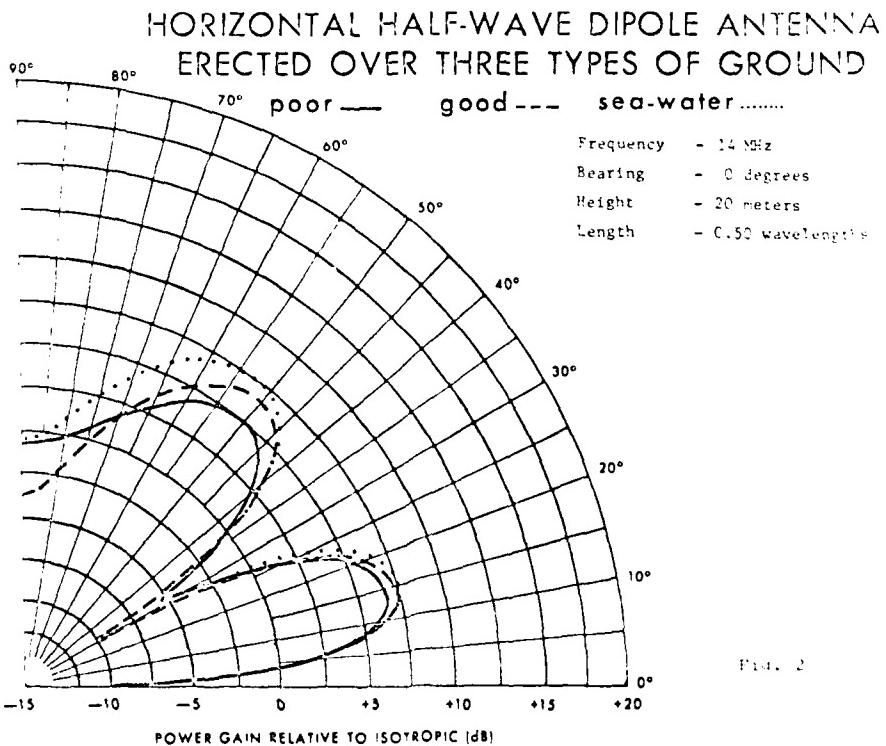
VERTICAL ANTENNA ERECTED
OVER THREE TYPES OF GROUND

poor ____ good ____ sea—water ....

Frequency - 6 MHz
Length - 0.25 wavelengths

POWER GAIN RELATIVE TO ISOTROPIC (dB)

Fig. 3

Fig.4

Occurrence of Reliability

Antenna #1

Antenna #2

Reliability (%)

LOG PERIODIC
FREQUENCY 5.1 MHZ
ELEVATION ANGLE 4.5°

Fig.5
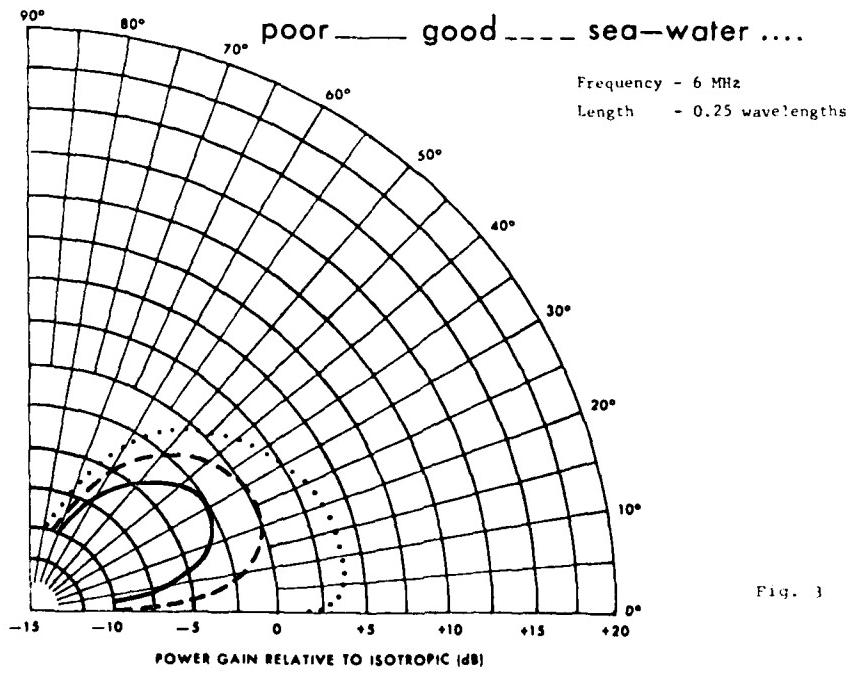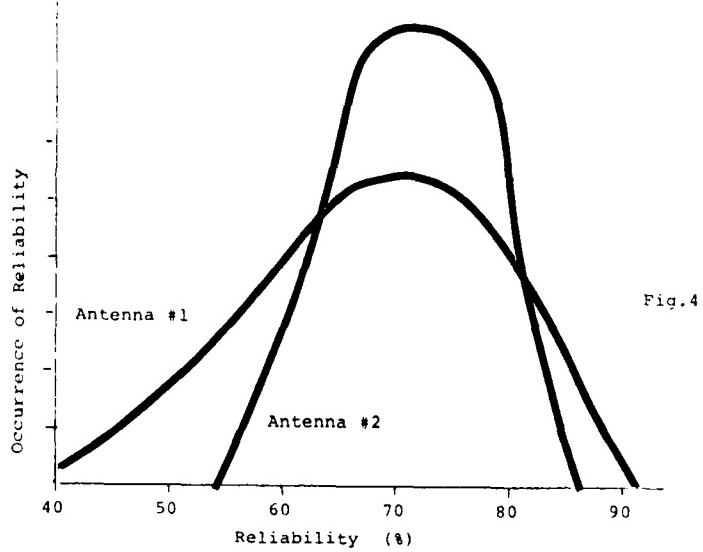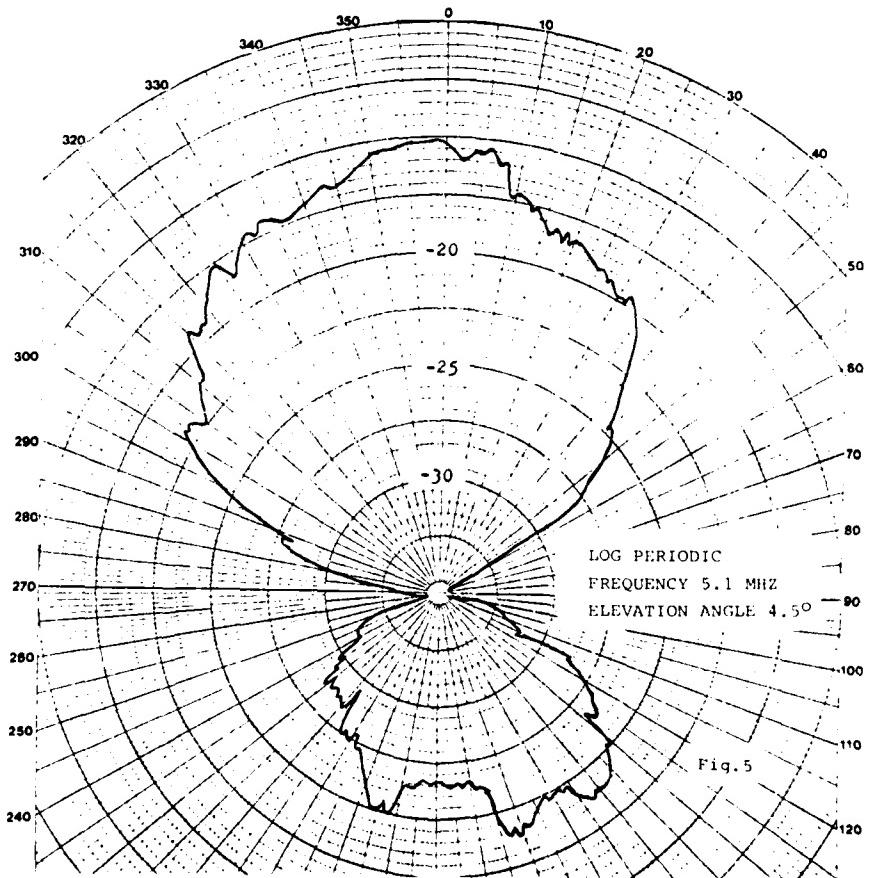
Antenna Radiation Pattern
Elevation Plane

DELTA
FREQUENCY 5.1 MHz

Fig. 6

POWER GAIN RELATIVE TO ISOTROPIC (dB)

LOG PERIODIC
FREQUENCY 5.1 MHz
ELEVATION ANGLE 9.1°
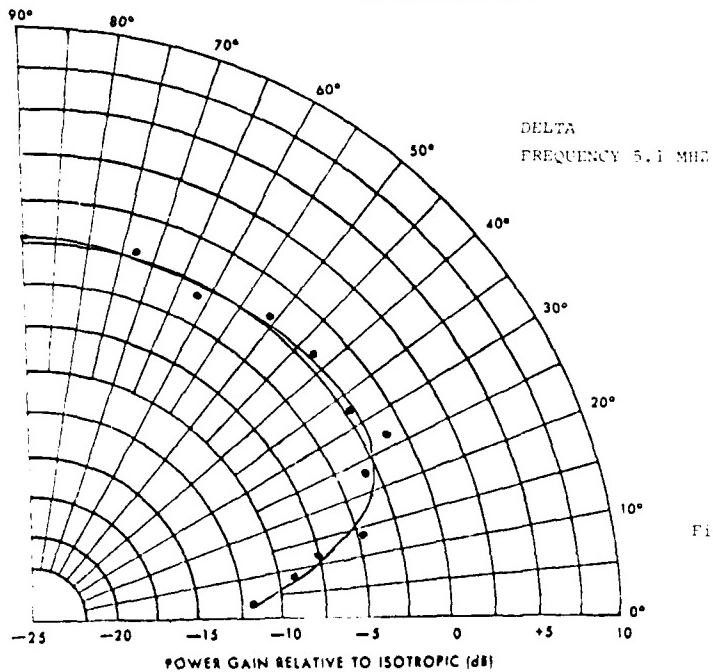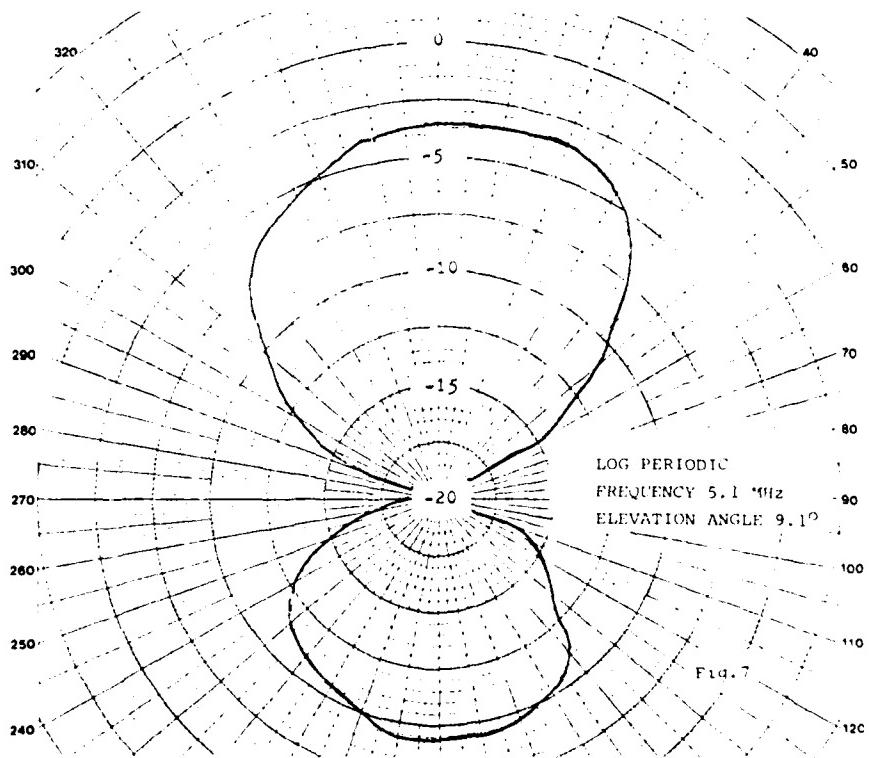
Fig. 7

HF EQUIPMENT: RECEIVERS, TRANSMITTERS, SYNTHESIZERS, AND PERIPHERALS

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SUMMARY

Although long distance communications via satellites has dominated the last two decades of radio equipment development, high frequency (HF) radio equipment is experiencing a high technology renaissance.

Satellite systems now transmit quality low data rate communications and navigation aids to mobile users, but the low cost and survivability attributes of HF radio are again being recognized. Emerging new systems automate network operations while adapting to propagation conditions. However, neither new replacement radios nor new systems furnish the potential capability of state-of-the-art components. This paper summarizes the advances in equipment, systems, and components.

INTRODUCTION

Before the advent of satellite platforms, sophisticated 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 which do 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 include microprocessor and synthesizer elements which have led to new circuit architectures. The first of the following three sections describes: Receivers; Transceivers and Antenna Couplers; 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 diffraction, aircraft, etc. allow low probability propagation to service 500 stations.

These systems are representative of what may be accomplished to satisfy specific needs. The equipment developed for these systems is also unique. More general capabilities may result from the use of the 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; Wideband Antenna Systems; Processors; and Real-Time Clocks.

A network control facility commonly comprises individual components of superior quality because space is not constrained and cost is minimized through reduced labor. However, compromises among performance, size, and cost considerations are usually necessary for the large numbers of mobile or transportable stations.

LOGISTICAL REPLACEMENTS

Evolving equipment designed to meet continuing needs has yielded significant improvement of receivers, transceivers, and input/output devices. Simplified operations through electronic control is a dominant impetus. The use of many highly trained manual operators 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 losing the performance available from the best manual equipment.

Receivers

The best predecessor surveillance receivers use ganged, narrowband, mechanical-tuning radio frequency (RF) amplifiers and local oscillators. New receivers tune by local oscillator synthesizers under digital control and do not use narrowband RF preselection because electronic preselection introduced so much nonlinearity as to cause an increased noise floor composed of high-order intermodulation (IM) products. Images are suppressed by having a first intermediate frequency (IF) in the very high frequency (VHF) band. The architecture illustrated in Figure 1 has obtained almost universal acceptance.

Competitors concentrate on dynamic range improvements and on digital control and display. The dominant performance improvement has been in decreasing front-end mixer nonlinearity which is often specified by the third-order intercept point (a single dBm level obtained from measuring the $2f - f$ intermodulation product as illustrated in Figure 2). Greater than 30 dBm is obtainable by commercial receivers exhibiting 14 dB noise figures. This improvement in linear mixer technology has exposed nonlinearities in the first IF filter (which is often a narrowband crystal lattice network). Synthesizer phase noise is now the dominant deterrent to discriminating against a strong adjacent signal, as illustrated in Figure 3. The strong signal acts as a local oscillator with which the phase noise in the receiver local oscillator "reciprocates" by acting as the signal; hence the term "reciprocal mixing." The local oscillator synthesizers are a competitive economical compromise involving the dichotomy between tuning speed and phase noise, but standard specifications or measurements do not exist nor would they be supported by system analysis. The impact of synthesizers will be greater than the advantage of digital frequency tuning which can be accomplished through a digital frequency counter. Synthesizers can provide arbitrarily quick frequency selection.

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 push-button 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 I lists typical receiver performance parameters. Note that digital modem compatibility now requires a specification on overall time delay as a function of frequency within single sideband (SSB) channels.

Transceivers and Antenna Couplers

New transceivers are following the lead of digitally controlled receivers. Transceivers continue to be designed primarily with SSB modulation for providing a voice bandwidth channel from which the interface to digital data peripherals is specified. Fast automatic antenna tuning is becoming available. The manifestations of these developments follow.

Remote controlled transceivers solve several problems. Aircraft cockpit space is minimal, as is indicated by the relative size of the control box of the HF transceiver components shown in Figure 4. Note the large size of the antenna coupler which has been designed to aircraft control surface constraints. Manpacks need to be controlled by the operator without removing the pack. Tactical ground vehicles are vulnerable to attack so that remoting control through fiber optics or wire is provided. The convenience of computer terminal control through a standard telephone interface is generally acknowledged by many users 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, but the implementation into a system implies systems network concepts which are generally lacking except, apparently, for several European "slow-frequency-hopping" applications.

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 ground wave, 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 which switch vacuum relays to prelearned positions. However, rotary mechanisms taking seconds to tune are common. In both cases, the reliability of mechanical components (exacerbated by high vibration at some aircraft antenna locations) causes radio maintenance problems.

Since human speech communications does not depend on phase or time delay characteristics, SSB radio and telephone service practice has been to emphasize sharp amplitude response filters which pass 300 Hz to 3000 Hz frequencies and strongly attenuate frequencies outside of this region. Phase characteristics which are 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 when receiving serial data 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 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 could be serious. Transistor oscillator circuits have less microphonics but relatively more flicker noise and often use synthesizer designs which accentuate some 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. The link 11 modem requires higher performance than most other data systems.

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 which puts performance requirements on the transceiver.

Handheld Data Terminals

Handheld terminals are substitutes for teletype equipment, computer terminals, or trained continuous wave (CW) operators. They allow off-line composition and editing before transmission at teletype or burst rates. Simple frequency shift keying (FSK) or phase shift keying (PSK) modulations are typical. Demodulation, storage, and display complete their capabilities. Some convert from alphanumerics to Morse code and subsequent Morse code demodulation. One generates a map from a plasma display grid. Most use a miniature membrane-type teletype keyboard but one uses tri-valued alphanumeric keys and a shift controlled by the fingers of the holding hand. Figure 6 illustrates representative military and commercial models.

The development of appropriate devices will continue towards some configurational consensus. Speculation suggests diverse possibilities. The Sholes (QWERTY) keyboard was designed (in 1878) specifically to slow typists and avoid mechanical key jamming. The efficient Dvorak keyboard may be more appropriate for miniature keyboards. Scrolling display keyboards, which are contingent on digram, trigram, or operational probabilities and on shift keys, seem likely candidates to reduce the number of keys.

Data systems requiring fast synchronization and burst transmission systems require fast automatic level control (ALC) and automatic gain control (AGC) in addition to linear phase (constant delay or low ringing) voice channel filtering.

Secure Voice

The United States has worked towards secure voice communications systems for 20 years. Earlier programs included 2400 bits per second (b/s) 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.

HF modems for high speed data such as 2400 b/s 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 of peak clipping provides the best overall performance. Nevertheless, new exploratory serial modem developments may provide performance improvements over the parallel modems.

The voice data modem need not maintain better than a 1/1000 bit error rate. Furthermore, some data are more important than other data. The HF 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 which must be especially robust or reception will not begin. Preambles comprise synchronization signaling sequences which may require 1-2 second for transmission. The synchronization time may be impacted by radio ALC and AGC time constants and by automatic antenna-tuning algorithms in some transceivers designed for push-to-talk voice communications.

Analog-to-digital conversion of voice for HF now strives for natural voice transmission within a 2400 b s data rate. Both modem and analog-to-digital 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 the modem will permit superior performance in the threshold performance area.

NEW SYSTEMS

The systems discussed below describe networks which may replace or supplement present 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 frequency agility has been limited by antennas, antenna switching, and antenna tuning. As the 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 propagation paths and modes.

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 of the Admiralty Surface Weapons Establishment is echoed by an attempt by the US Navy to radically lower all self-generated components of the noise floor. Although this symbiosis is indicative of a probable new HF networking capability, ICS-3 primarily enhances ground wave and sky wave transmissions through rapid propagation selection and data handling.

Figure 7 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 handles 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 ICS-3 related programs spawn new hardware of revolutionary capability. That equipment, in turn, should enable networks capable of controlling link emissions so as to minimize self-interference and notice by other spectrum users.

SELCAL, Scanning, and Channel Evaluation

Large SSB voice networks that service aircraft seek to automate 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 along with a scanning system which 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 such as: signal level, signal-to-noise ratio, digital error detection, sounder signal parameters (timing and time delay continuity), and memory of recent assessments. The requirement is to recognize a desired signal and to rate it among alternatives which may be in different bands or from alternative paths or modes. Signal fading and noise variations preclude accurate single-channel measurements in less than several minutes; quick measurements require using several frequencies simultaneously to obtain frequency diversity. Therefore, a memory table is often the final arbiter of channel selection. One scheme gives greater weight to recent measurements.

No HF networks have been automated using scanning although 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 8 shows a proposed scanning scheme which requires 0.5 second per frequency.

A prognosis for the use of scanning for adapting to ionospheric conditions is that the advantages of automatic alternative routing via the best frequency and path will be utilized in limited networks but timed systems will provide access within large networks.

Implementation of SELCAL and scanning techniques can take place through the remote electrical capabilities of new replacement radios. Add-on funding for aircraft terminals may be anticipated for interim advantages if the ground network control stations can be reconfigured economically. Ultimately, programs such as ICS-3 answer more needs.

Packet Radio

In contrast to SELCAL and scanning schemes, packet radio is a complex computer data exchange relayed among many line-of-sight (LOS) radio nodes. An "L-band" system was chosen for its frequency assignment availability and in consideration of exchanging 1000-bit data packages in microseconds. Table 2 lists hardware for this cooperative network development, which is already influencing the international HF community.

Packet messages carry their own addresses to allow the message to seek a path through several nodes. The dominant network concern is to avoid overload disasters. Message loops are eliminated by protocols which keep track of transition through "tiers." Overload is minimized by "transfer point" routing and by slowing the pace

under stress. Unknown routes may be ascertained by "flooding" requests for routing aid. Present implementations use relatively few nodes and repeaters located at highly visible locations. In fact, test network sizes are more appropriate for HF. The reliance on alternative parallel routes to the next tier will be of interest to automatic HF network designers.

Figure 9 shows a packet radio comprising antenna, RF module, and processor module. The processors would be adequate for the message handling function in an automatic HF radio.

SNOTEL

SNOTEL is a meteor-burst communications system which relays snow cover data from over 500 remote mountain data collection sites in the Western United States to a central processing facility in Oregon. It is notable for the size of the network (see Figure 10), the transient and diverse propagation path modes, the small remote terminal, and the speed of installation. Furthermore, it provides an argument against the arbitrary limiting of HF equipment to 30 MHz. HF alone does not provide the adaptivity desired for communicating during disturbed conditions. 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 maximum usable frequency (MUF), sporadic E, meteors, diffraction, ground wave, and even reflections off aircraft.

All the above propagation modes are exercised by meteor burst communications systems, but the dominant design consideration has been the transmission of a packet during a meteor trail life of less than a second. Meteor availability is a seasonal and diurnal variable which complements HF propagation. The SNOTEL remote terminal generates 300 W to send 2000 b s PSK transmission in 0.1 second. The low duty cycle allows a solar-cell remote terminal power system. Figure 11 illustrates the portable test terminal.

The network interrogation protocol makes use of the limited propagation footprint afforded by meteors and protects against continuous propagation afforded by ground wave and some diffraction paths approaching 100 miles. Although the system experiences difficulties with auroral conditions, it successfully interrogates 90% 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 which uses many propagation modes in the relatively unused HF VHF border frequencies. SNOTEL demonstrates that a few network control stations have a strong probability of communicating with a mobile terminal within a few minutes.

NEW COMPONENT TECHNOLOGY

New system concepts have promoted component improvements which 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 colocated equipment would approach atmospheric noise. Some of the reports from development programs are now appearing.

Wideband HF antennas are not new, but innovations are under investigation. Some ways to obtain a wideband capability are described in this section.

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 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 colocated with transmitters is that of receiver front-end protection. Current shipboard installations often include pads in the receiver antenna leads to minimize the problem. In other cases, inefficient receive antennas (such as short active whips) are used. These approaches assume the presence of an excess of wanted signal-to-noise ratio; this assumption is inconsistent with scenarios involving ground wave propagation and excessive ionospheric absorption resulting from nuclear blasts or from naturally occurring phenomena.

The newest diode mixers allow systems to obtain +33 dBm input intercept points but suffer 12 dB noise figures. Diode mixer loss contributes directly to the noise figure along with various circuit insertion losses such as those from the first IF filter.

Two circuit approaches to achieving low noise figure 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 up-convertor mixers have been employed in HF systems which exhibit 6 dB noise figures and 20 dBm intercept points. Amplifiers are characterized in Figure 12 by noise figure, and 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. Filtering can help decrease intermodulation products from interference sufficiently far from the signal frequency. High-power, agile, programmable filters have been under research for ultra high frequency (UHF) transmitters, but not for HF receivers.

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 systems. Hence, they are used in both transmitter and receiver 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 operating in an environment of noisy, high-level, off-frequency interference. This latter problem occurs even if transmitter signals are pure tones because of reciprocal mixing whereby off-tune synthesizer noise signals mix with off-tune receiver input signals to produce the receiver IF frequency (see Figure 3).

Two basic kinds of synthesizer have evolved. One is the indirect synthesizer which 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 which 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 essentially equal noise-floor performance is now available from commercially available units of both types (phase noise floor in the order of 135 to 140 dB below the desired output). However, the direct synthesizer is still 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 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 colocated transmitters.

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 paragraphs generally applies to these stand-alone units. Noise performance of built-in synthesizers tends to be poorer than for laboratory type equipment (see Figure 13). Synthesized communications receivers and transceivers currently available use phase-lock-loop systems which exhibit several millisecond tuning times rather than the several microsecond tuning times that are available through the use of larger and more expensive direct frequency synthesizers. Furthermore, their performance is frequently poorly documented. New competitive designs are underway which make use of novel, high-speed, large scale integration (LSI) count-down circuits which aid performance through the improved circuit architecture that they permit.

Standards are required for specifying all aspects of spectral purity and settling time both for stand-alone and built-in synthesizers.

Power Amplifiers

Most power amplifiers (PAs) 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 electromechanical in nature, and are therefore relatively slow (on the order of 10 to 30 seconds to tune across the HF band) and can be relatively unreliable.

Alternatively, wideband amplifiers have been produced, with early models using distributed transmission line technology. These suffer from relatively high harmonic, IMP, and spurious outputs, as well as from wideband noise. However, since they eliminate the function of PA 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 is to use wideband technology at the lower power levels and switched half-octave output filters. The latter have the characteristics that for any given frequency within a half octave, all harmonics lie outside the selected half-octave pass band. This approach permits implementation of an 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 transfer time of the vacuum relays required for switching (about 25 ms). Reducing this switching time through the use of diodes has been demonstrated at UHF but not at HF. Additionally, there is the question of spurious products resulting from the nonlinear diode characteristics.

With the development of RF power transistors, a new linear RF power amplifier architecture has emerged whereby the outputs of a number of relatively low-power modules are combined using 90° hybrid couplers to produce a high-power unit, typically several hundred watts. 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 output models 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 of 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.

PA units for use at HF are currently designed for linear operation amplification of SSB signals. Such operation permits easy implementation of output emission level control through manipulation of low-level input signals, minimizes output filtering requirements, and minimizes interactive generation of spurious signals by multiple colocated PA units (for example, in ships). The serious problem of colocated equipment in the ICS-3 system has prompted recent activity to produce better linearity (and lower noise) through the use of feed-forward techniques. Figure 14 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, 80 dB suppression of two-tone, third-order products has been obtained for 500 W peak effective power (PEP) at 5 MHz. Performance is poorer by 20 dB at 30 MHz. Unfortunately efficiency is low (less than 10%); hence mobile equipment performance would suffer.

In addition to discrete spurious emissions, wideband noise can present a problem, particularly for systems which require colocation of transmitters and receivers where wideband noise from the PA interferes directly with wanted low level signals at the receiver.

Modern PAs do not suffer damage from output mismatch. Circuitry can now be provided to sense the mismatch and turn 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.

The use of linear class-A RF power amplifier circuit operation 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, 7 dB of improvement for portable power-limited transmitters and jammers is available through efficient transmitters.

Linear 1 kW PA units, available commercially, are capable of frequency-agile operations over the entire HF band. These users provide acceptable harmonic suppression over reasonably low-level intermodulation products, mismatch protection, and utilize the principles of graceful degradation of performance and on-line PA module replacement.

Wideband Antenna Systems

In proposed wideband adaptive HF systems, calibrated antenna gain and elimination of mechanical tuning are more important than efficiency.

Poorly calibrated conventional antennas require larger power margins by increasing system uncertainties. Accurate antenna calibration is expensive and magnifies the bookkeeping problem with attendant increases in the amount of memory and computation required to implement real-time minimum power adaptations. A fully adaptive HF system generally operates on low power and does not require efficient antenna structures, while well-behaved antennas can enhance system performance and the prospects of quiet operation. Using identical and nearly colocated transmit and receive antennas simplifies the bookkeeping problem in systems that assume propagation path reciprocity.

The large inverted cone is an international antenna used for sounding and monitoring that will undoubtedly be used for many more years. The fact that such antennas are not necessarily well-behaved is important in path loss measurements and in systems employing power control.

Figure 15 describes an airborne pattern testing configuration used for inverted-cone antenna calibration. The antenna was exercised by a transmitting antenna towed from a helicopter flying systematic concentric maneuvers about the inverted cone. A small monopole on the ground was used for comparison. Results of the tests are shown in figures 16 and 17.

The inverted cone exhibits good loading properties and accommodates very high transmitter power. However, the actual gain characteristics of the popular inverted cone supported from six poles diverged considerably from expectations and specifications. Figure 16 shows the vertical antenna gain in decibels (over isotropic) as a function of frequency and elevation angle. The largest gains were several decibels lower than expected. Also, notice the fairly wide variations between equal-gain contours at a constant frequency or constant elevation angle. Two nulls on the order of 15 or 20 dB are apparent. Fortunately, both these nulls are outside the usual operating region, which roughly includes frequency and elevation angle pairs below the straight line. Even in this region the antenna gain can vary by several decibels, especially at the lower elevation angles. Such data would have to be stored for each antenna employed in HF systems measuring path loss.

Variations in the horizontally-polarized inverted-cone antenna gain were truly dramatic, as depicted in Figure 17. According to the antenna specifications, there should have been essentially no horizontal gain and less than 1 dB variation with azimuthal angle in vertically-polarized patterns. Even the vertical gain had several decibel discrepancies at different azimuths. Again, such data must be part of any accurate path loss measurements.

Other broadband antennas are illustrated in Figure 18. Recent approaches include tuned-trap antennas which intentionally introduce resistive loading. Figure 19 shows the performance of a tuned-trap antenna developed for ICS-3. A voltage standing wave ratio (VSWR) less than 3.5:1 over the frequency range of 6 to 30 MHz is obtained at efficiencies near 40%.

A simple, electrically-short monopole inductively loaded in the middle exhibits improved overall efficiency if it is resistively loaded at the base. Figure 20 shows the performance improvement of an antenna 3 m high. Combining antenna loading and a fat monopole produces the Smith chart performance shown in Figure 21. Considerable freedom is available towards developing improved field antennas having acceptable loading properties.

Frequency-agile antenna couplers may not be necessary for frequency hopped emissions. The bandwidth of couplers may be approximately 1 30 of the center frequency. Hopping within the bandwidth is appropriate for HF ionospheric propagation. Resistive antenna loading can improve the bandwidths available. Alternatively, several antennas can be used.

For most systems, reception requires greater agility than transmission. Receiving antennas can be made broadband through impedence matching and active elements. There are now many commercial and military versions of the short electrical antenna coupled by transformer and or amplifier to a transmission line.

Processors

Digital control systems provide the nervous system for modern HF system designs. They orchestrate automatic adaptivity, timing, frequency hopping, channel evaluation, radiated-power level control, information flow control, packet and other protocols, and other control functions required in a modern HF adaptive system. Specialized computers have been provided in those few adaptive systems so far investigated.

A new general class of equipment known as microcomputers or instrument controllers has recently emerged. One of their functions is to control and or monitor other instruments. They have been used for control of receivers and communications systems but may not be suitable for the more sophisticated systems. However, the principle of using a central controller is now well established and understood. It appears that BASIC is the software language and the RS-232C and IEEE 488 bus are the external interfaces selected by the controller design community. BASIC language interpreters are slow; therefore, machine language subroutines are necessary for fast functions such as timing and data transfer. Fast digital operations are best accomplished by using the internal bus structure of these computers. Interactive functions are accomplished by well-known higher-level languages.

Since instrument controllers and communications system controllers have important differences (especially regarding real-time functions), the need to consider suitable architectures in these areas is clear. There are virtually no operational military systems which use fully automatic central controllers. Experimental systems are usually based on 16-bit military or commercial minicomputers.

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 printed-circuit (PC) card accessory for, instrument controllers, but their accuracies and resolutions are still mostly unacceptable for 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 exhibits no drift 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 22.

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 1 ms 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 b s synchronized data system which imposes a bit timing accuracy of about 40 μs 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.

CONCLUSION

Radio equipment is electronically controllable by small computers. New communications systems which depend on electronic agility are in experimental stages. Analog radio circuits are being developed for these computer controlled systems. The interfaces between computer and analog radio circuits are as yet ill defined but a new architecture is evolving which provides increased agility and adaptivity in HF VHF systems.

Table 1.   Typical Receiver Performance

Sensitivity	0.5 μv for 10 dB (S + N) N (BW = 2.4 kHz)
Linearity	+20 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 2.4 kHz channel
Tuning Speed	Less than 25 ms
Extraneous Responses	80 dB suppression of images, IF frequencies, etc.
Spurs	Not more than ten 3 dB above receiver noise floor None greater than 10 dB above receiver noise floor
Bandpass Time Delay	Less than 0.5 ms variation between 600 Hz and 2700 Hz
Control	All remoted via RS-232C, IEEE 488 or BCD parallel
Built-in-Test	Function level

Table 2.   Packet Radio Evolution

	Experimental Packet Radio (EPR)	Improved Packet Radio (IPR)	Value Eng Packet Radio (VPR)	Low Cost Packet Radio (LPR)
Development	1974–1977	1976–1979	1980–1982	1981–1983
Number	28	27	18	2
Net Timing	Asynchronous	Asynchronous	Asynchronous	Asynchronous
PN Code	Fixed	Fixed	Fixed	DES Driven
EDAC	CRC	CRC	CRC	CRC FEC
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: 10 W, Bandwidth: 20 MHz
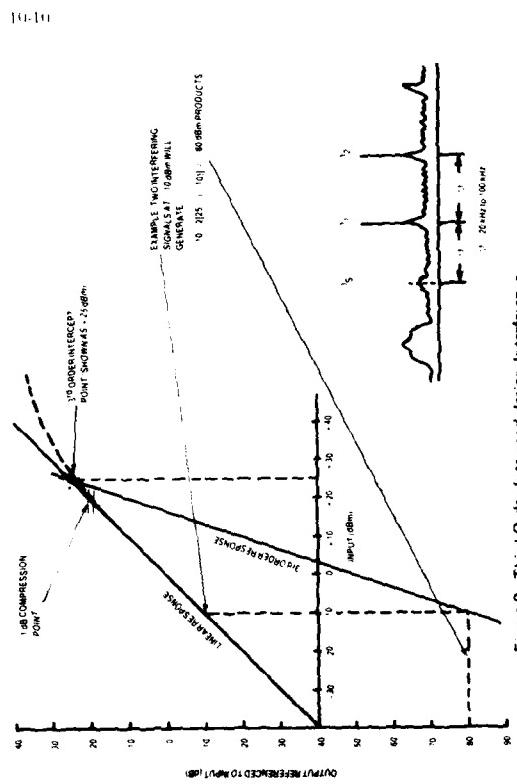
Figure 2 Third-Order Intermodulation Interference

Figure 4. Airborne Transceiver Components

Figure 1. Block Diagram of a Contemporary Receiver

Figure 3 Reciprocal Mixing of Interference and Local Oscillator

Figure 6. Small Input/Output Devices

Figure 8. Calling Sequence for N Channel Scanning

Figure 5. IF Filter Response

Figure 7. Narrowband and Wideband Transmitter Systems
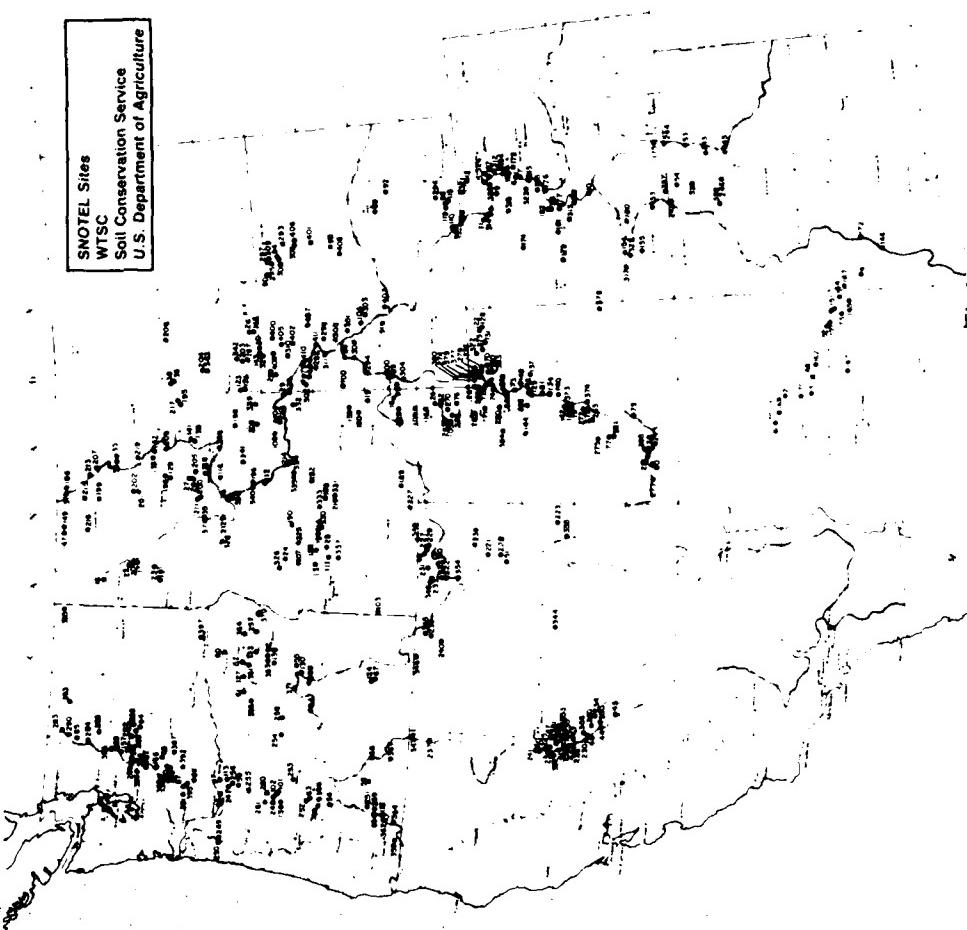
Figure 10. Western United States SNOTEL Sites
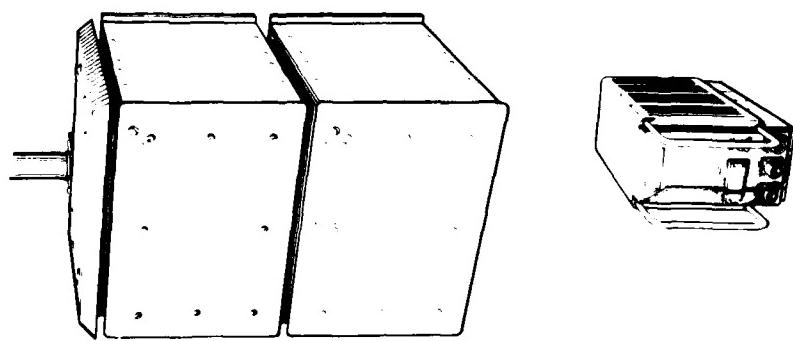
Figure 9. Packet Radios

Figure 12. RF Amplifier Noise Figure vs. Third-Order Intercept

Figure 14. Feedforward Amplifier Linearization

Figure 11. Portable Meteor Burst Terminals

Figure 13. Phase Noise of Representative Synthesizers
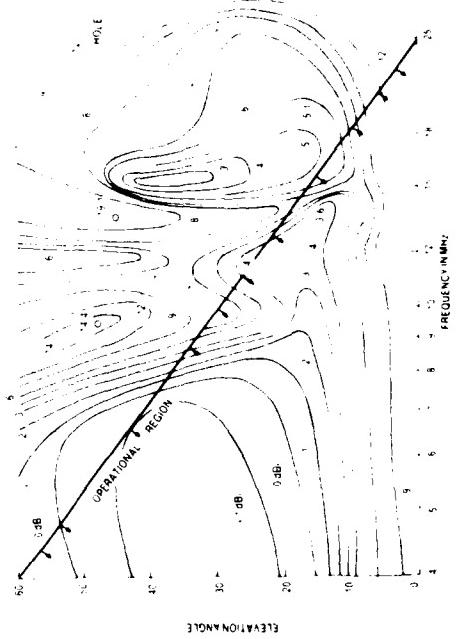
Figure 16. Inverted Cone Antenna Gain (Vertical Polarization)

Figure 18. Wideband Antennas

$$ERP(0°) \cos^2\psi = ERP(\psi) P_{(\psi)}$$
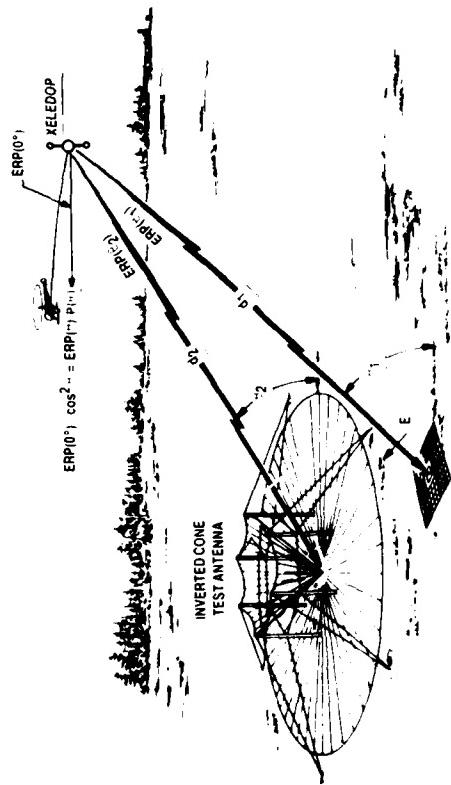
Figure 15. Antenna Testing
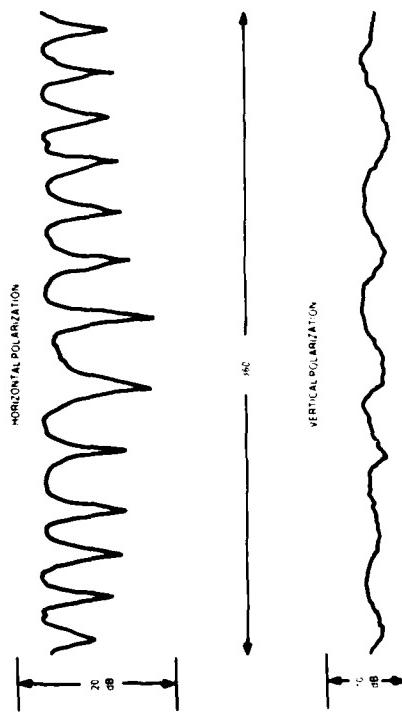
Figure 17. Inverted Cone Azimuthal Gain

Figure 20. Efficiency Gain vs. Ground Resistance for Loaded Whip

Figure 22. Typical Short-Term Instability Curves

Figure 19. Transmission-Line Loaded Monopole

Figure 21. Impedance for HF Field Monopole

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 con-
ditions, 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 mini-
mizing the effects of ionospheric disturbances, such as path and time diversity, the use of early war-
nings, 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 com-
panion 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. Fre-
quency 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 inter-
ference 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 F-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 problems 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

$$n = \mu(N_e, \nu) - i\chi(N_e, \nu)$$

which both depend upon the electron density N_e and on the collision frequency ν 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 ν 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 ν 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 e) in Table 1) are strongest and most likely in latitudes above 40°-50°.

3.3 Winter anomaly in ionospheric absorption

At middle latitudes (35°-60°), 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°. 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 = $\dfrac{\text{Oblique path absorption (dB)}}{\text{Riometer absorption (dB at 30 MHz)}}$
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 (450 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 R to A).

IONOSPHERIC CONDITIONS SYSTEM	QUIET I(0-0.1 dB)	MODERATE II(0.2-2 dB)	DISTURBED 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 absence 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	
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	Probable 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 patterns
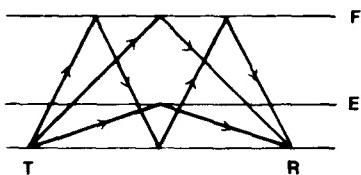
Figure 1   Schematic representation of possible multipath propagation
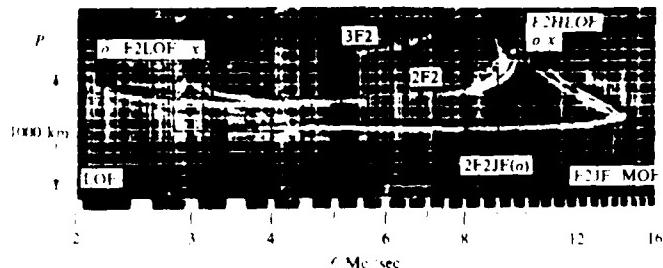
Figure 2   Example of ... propagation with multipath ... Muller 1964 .

t = 8.1 MHz, MUF = 8.2 MHz
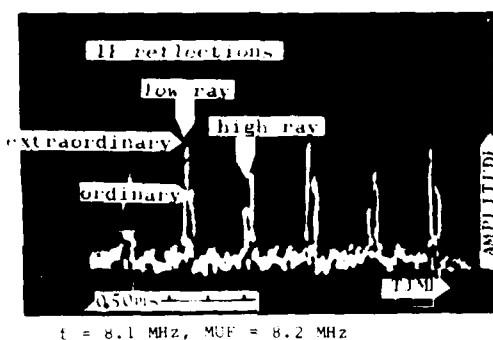
Figure 3   Multipath echoes from spread spectrum transmissions at 8.1 MHz (Kain 1981).
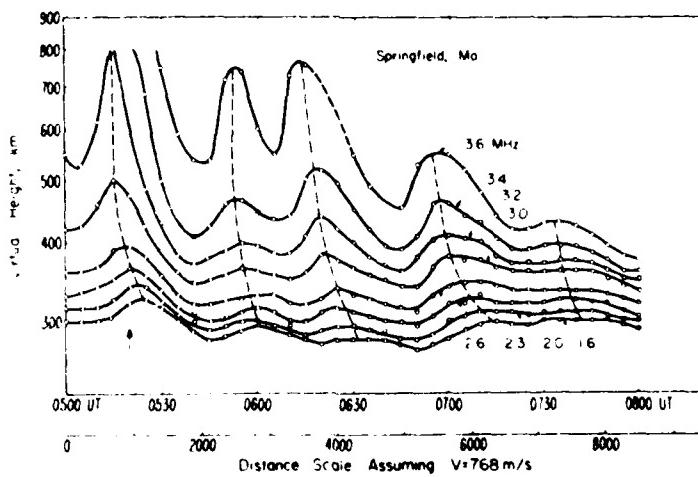
Figure 4   Virtual height variations observed in the F-layer during a large scale travelling ionospheric disturbance (Georges 1967).
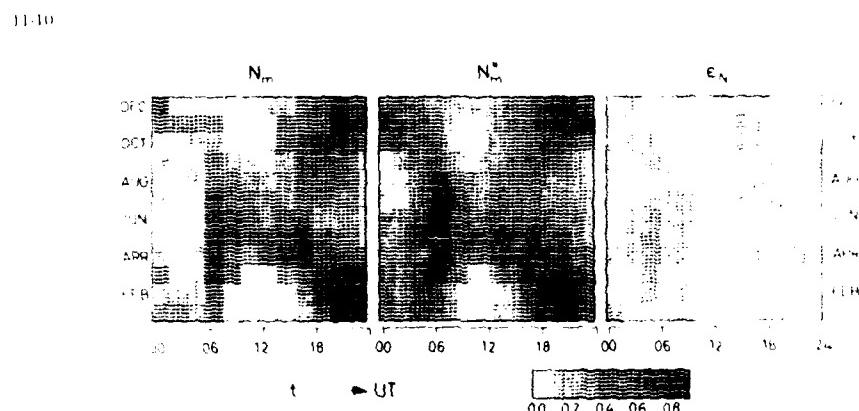
Figure 5   Mean occurrence pattern N_m of TIDs near equator (mean taken over 28 months). The gray scale gives the occurrence frequency between 0 and 1. N_m^* is the filtered occurrence pattern, taking the fourier series up to the third harmonic where E_N is the residual value (Röttger 1978).
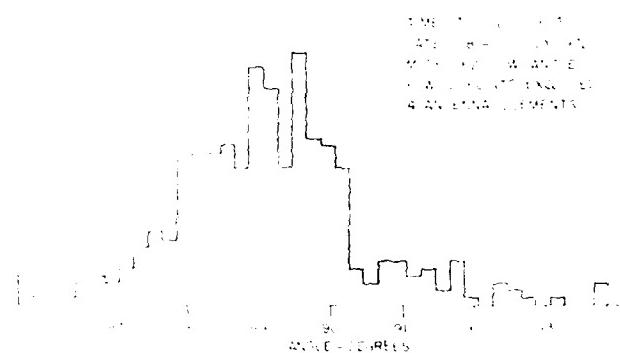
Figure 6   Measured distribution of direction of arrival angle. The angle is here measured relative to the antenna axis. The great circle path to the transmitter has an angle of 89.43° (Pike 1976).
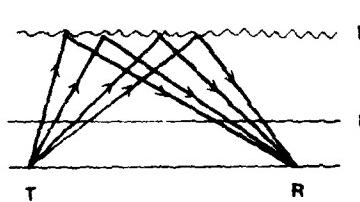
Figure 7   Illustrating reflection from an irregular ionosphere
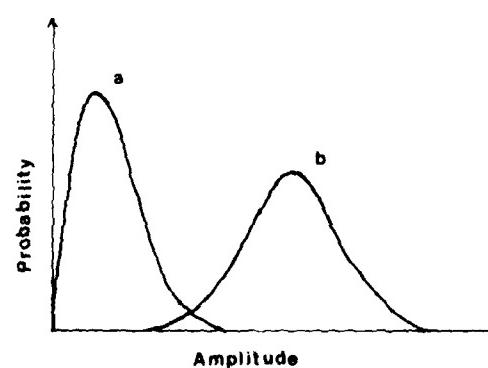
Figure 8   Schematic probability distribution of signal amplitude a) Rayleigh distribution from a random screen b) Pice distribution in the presence of a steady (specular) component (Davies 1969).
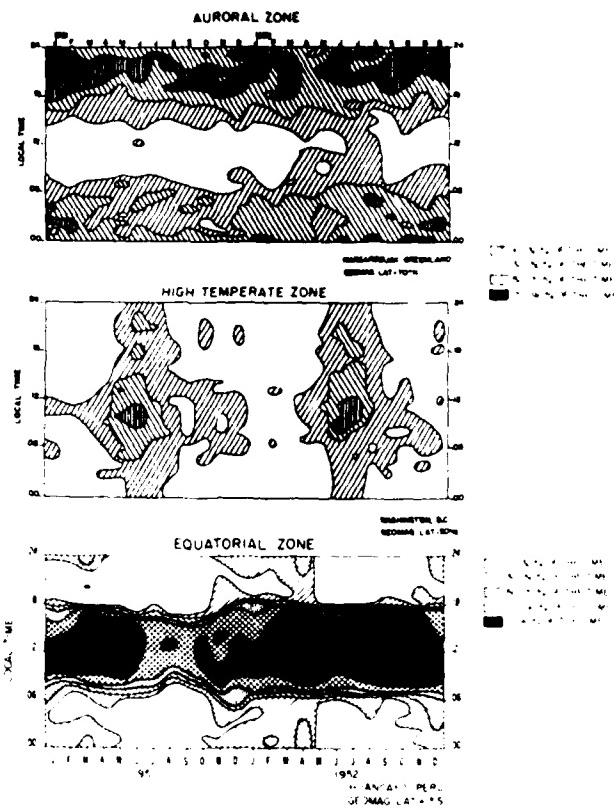
Figure 9   Fraction of time f$_{o}$E$_s$ > 5 MHz (Smith 1957).
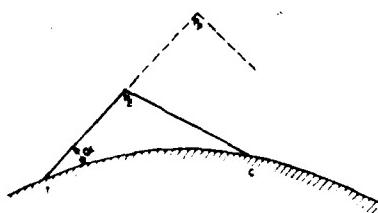
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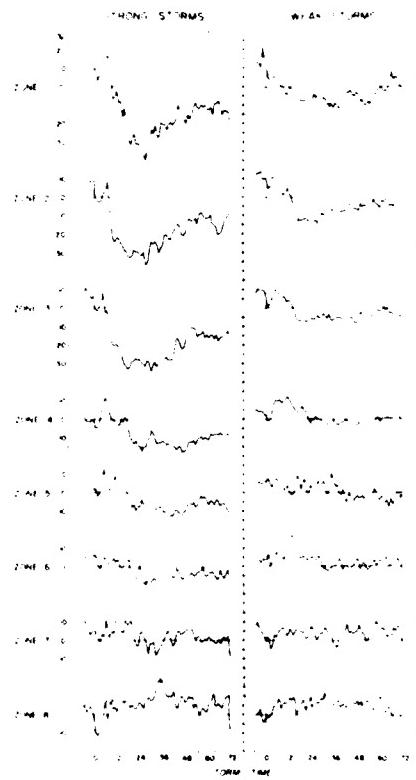
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            ordinate is the approximate percentage deviation from the quiet day behaviour versus storm
            time. The zone number is in the following shown in parenthesis between the applicable
            geomagnetic latitudes 60° (1) 55° (2) 50° (3) 45° (4) 40° (5) 30° (6) 20° (7) 10° (8) -10°
            (Matsushita 1959).
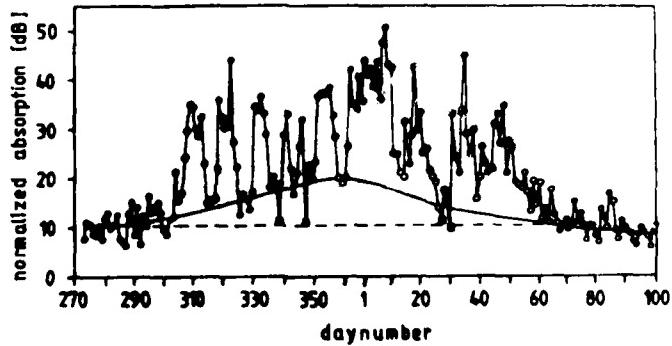
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            Full squares show "summerlike" days (Schwentek 1971).
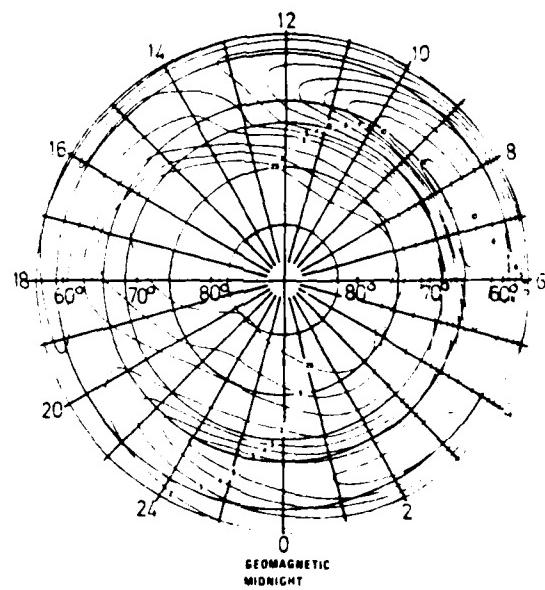
Figure 13   Percentage of the time that auroral radio wave absorption of 1.0 dB or more occurred at
30 MHz. The data are plotted as a function of geomagnetic latitude and mean geomagnetic
time. (Hartz et al 1963).
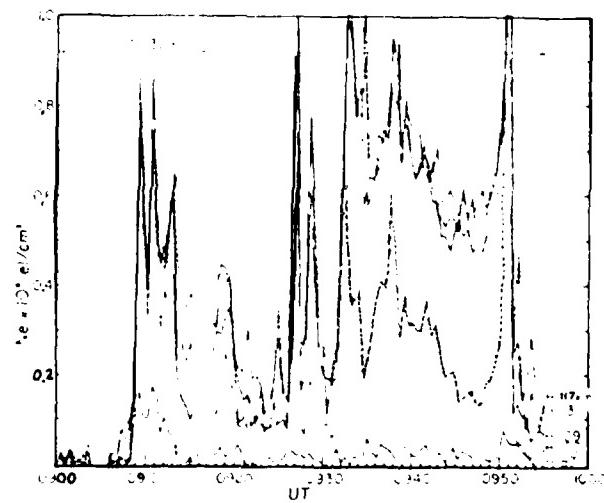
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with the Chatanika radar (Hunsucker 1975).
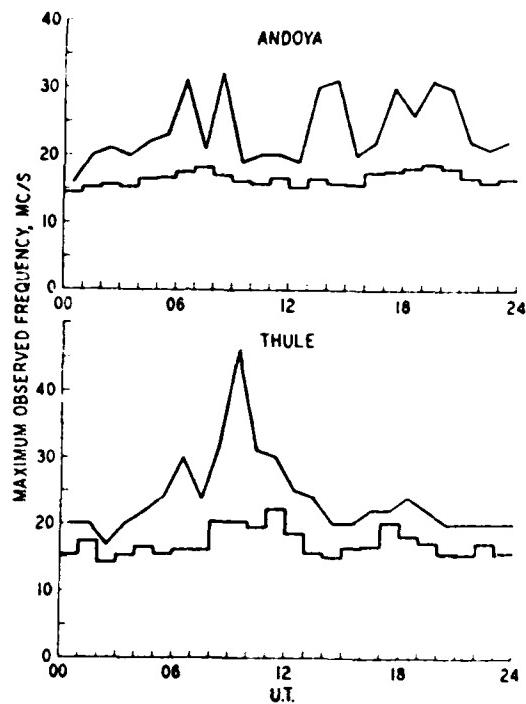
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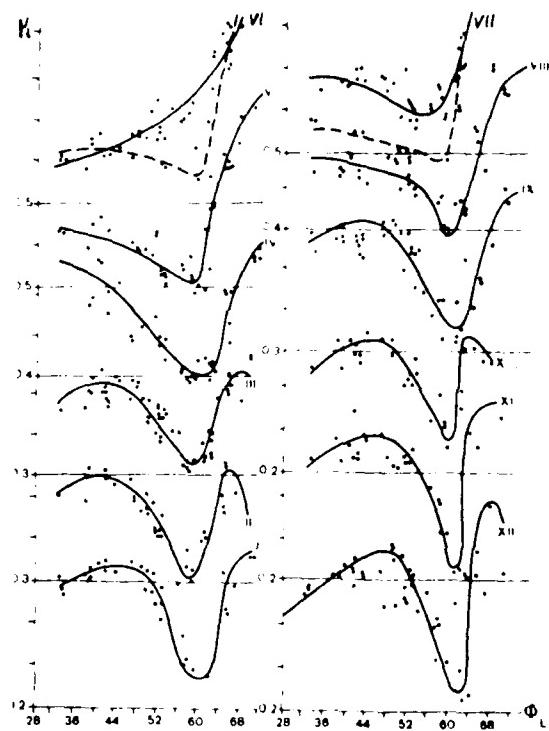
Figure 16   Latitude dependence of the normalized-to-noon values of the critical frequency for the
              F2-layer at 02 local time for the months of 1964. Besprozvannaya et al 1979.
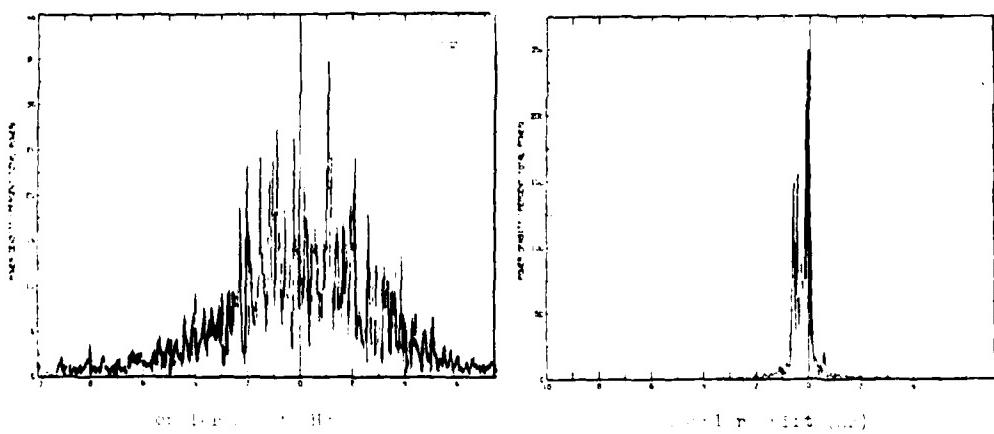
Figure 17   Typical power spectra of transmissions from Thule (high latitude) and from Fort Monmouth
              (mid-latitude) to Palo Alto (Hunsucker and Bates 1969.
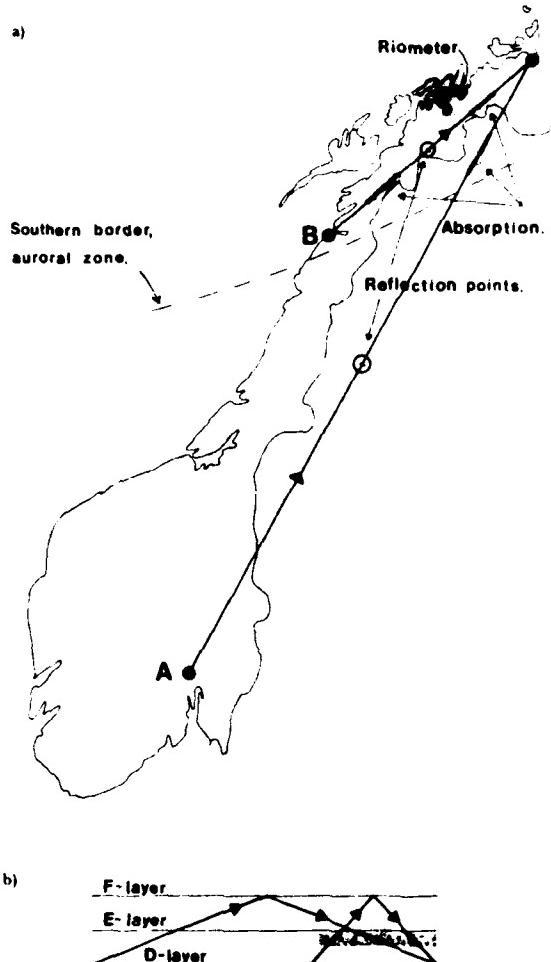
Figure 18   a)   The two transmission paths used in the tests.

b)   Schematic representation of the wave paths through the disturbed auroral D-region
(Thrane 1979).
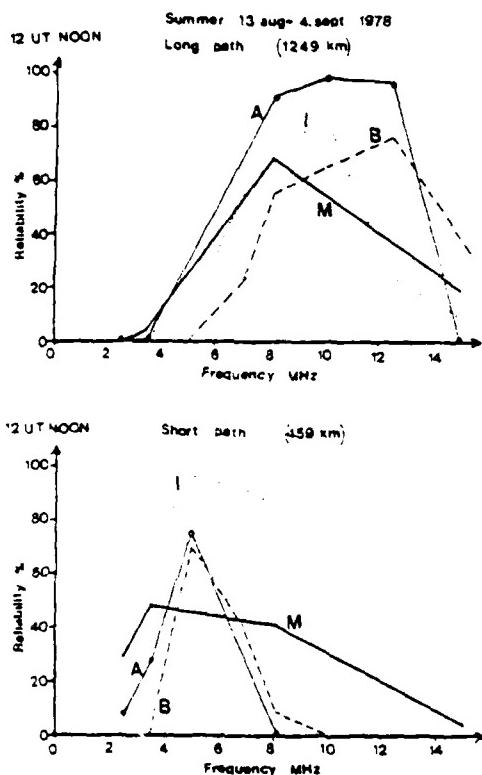
Figure 19   Measured and predicted reliabilities for the two paths for summer noon

        Code: A   Applab III
               B   Bluedeck
               I   IONCAP
               M   Measurements

        (Thrane and Bradley 1981).

HF PROPAGATION MEASUREMENTS FOR REAL-TIME
CHANNEL EVALUATION (RTCE) SYSTEMS

by

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ABSTRACT

Recent adavances in microprocessor technology, in frequency-agile HF equipment and in the understand-
ing of the propagation medium make it feasible and practical the adoption of adaptivity approaches in HF
communications. Signal-to-Noise ratio, multipath spread and Doppler spread are the basic parameters to
which the link has to adapt itself. A necessary prerequisite for adaptivity is Real Time Channel Evalu-
ation (RTCE). Measurements of these parameters 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 emphasize the measurement of
signal amplitude and noise levels (inclusive of interference) at several frequencies in specific HF paths
of interest, with a few instances of inclusion of multipath spread measurement. There is little doubt that,
at the end of the R&D phase, RTCE and link adaptivity will enter the practice of modernized HF communi-
cations.

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 the 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 operating as a complete terminal for two-way digital
communications). During the pauses of the emissions, measurements of noise and interference could be per-
formed 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 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 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 communicating. At each sounding cycle, information about the fre-
quency 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 cod-
ed 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 investi-
gation 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 appropri-
ate as the basic sounding/probing peridiocity. Adaptive approaches to the communication problem are requir-
ed 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 measure ment 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 (1980 [1] ; 1981 [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 trans-auroral path, in the band 3 to 30 MHz. These authors advocated a tightly integrated sounding/probing/communi cations 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 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 constraints, 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 forth coming conditions of the path. Early knowledge of PCA events, of magnetic storms 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 electr_ magnetic 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 succesfully implemented in correcting for ionospheric-induced errors in high-performance radars (Katz et al., 1978 [4]). Figure 1(taken from Aarons and Grossi, 1982 [1]) 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 (1980 [1] ; and 1981 [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 (1980 [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 preceeded by a "tone" of notification and takes place even while communications go on, 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	5 MHz to 30 MHz	5 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 seconds	one every 480 seconds
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
Pulses 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

(*) 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 666 Hz.

(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).

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 maesurement of multi path spread and Doppler spread can be achieved with a variety of methods, either based on the direct measure ment 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_s, B_d, B_m and the function $S(\cdot , \cdot)$, called the path's scattering function. In actual practice, it is sufficient to simplify the scattering function to a group of N gaussoids, with $B_s = 0$, $B_d = 0$, 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 f the path is

$$S(\cdot , \cdot) = \cdot (\cdot) \cdot (\cdot)$$

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 infiniresimal 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. Figure 8, taken from Green ([5], 1968), is useful in visualizing 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:

$$S(\cdot , \cdot) = \sum_{i=1}^{N} P_i \, (2 \cdot B \, L_i)^{-1} \exp \left\{ -\frac{1}{2} \left[\frac{(\cdot - \cdot_i)^2}{L_i^2} + (\frac{\nu}{B})^2 \right] \right\}$$

In this formula, the parameter N represents the number of paths in the structure, $\cdot_i$ 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 ith 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. Maesurement 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 (1980, [7]) give an account of the methods usable in probing the channel and illustra te possible hardware implementations. 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 two 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 $Q(\cdot)$ and the Doppler power density spectrum $P(\cdot)$ (Bello, 1963 [8]; 1965 [9]).

2. Multipath spread measurement

A signal waveform that appears an obvious choice for estimating the Delay Power Spectrum $Q(\cdot)$ is a short pulse of time duration $\cdot$, narrow in width when compared with the characteristic variations in $Q(\cdot)$. Consider the signal

$$z(t) = p(t)$$

$$p(t) = \begin{cases} 0 , & |t| > \cdot /2 \\ 1 , & |t| \leq \cdot /2 \end{cases}$$

$\cdot \leq$ individual
path width (mode width)

An estimate can be obtained by square-law detection of the observed process, and by subtracting from it the noise bias. Thus, for the response to one of the pulses in the train, we have :

$$\hat{Q}(t) = \left| r(t) \right|^2 - \overline{\left| n(t) \right|^2} = \left| \sqrt{A} \int g(\cdot) p(t - \cdot) d\cdot + n(t) \right|^2 - \cdot_n$$

where

$r(t) = \sqrt{A} W(t) + n(t)$, received signal intensity

$\left| n(t) \right|$ rms noise level

$\cdot_n$ noise variance

A average power level of received carrier

$g(\cdot)$ time-invariant response of the channel to an impulse $\cdot$ seconds earlier

$w(t) = \int_{-\infty}^{+\infty} g(t,\cdot) z(t - \cdot) d\cdot$, channel output

From the equation for $Q(t)$ written above, we can determine the range of delay values wherein $g(t,\cdot)$ as a function of $\cdot$,is negligibly small. Thus, if

$$g(t, \cdot) \approx 0 , \quad \text{for } \cdot < \cdot_{min} \text{ and for } \cdot > \cdot_{max}$$

then, the wanted delay spread is given by $\cdot_{max} - \cdot_{min}$.

$Q(t)$ will be a filtered version of the true function, if sufficient averaging is carried out. The measurement resolution is determined by the signal properties, so that the pulse width must be narrower than the mode widths of $Q(\cdot)$, to give an adequate estimate. However, even in the best possible conditions, the noise contribution to the estimate of the variance is significant. This conclusion applies directly to the measurement of the peaks of $Q(\cdot)$, and some allowance must be made for the accurate measurement of the low-level details of $Q(\cdot)$. Therefore, the single pulse approach appears to be unadvisable, because of its requirement of high Signal-to-Noise ratios.

An alternative estimation scheme can be utilized, if there is the possibility of carrying out coherent processing at the receiving terminal. Let $z(t)$ be a pseudo-random sequence of period T, which is used to modulate the carrier (bit length equal to $\cdot$). When the sequence is periodic, it can be assumed to have the

following property :

$$\int_0^T z(t) \, z^*(t - \tau) \, dt = \xi(\tau)$$

with

$$\xi(\tau) = \begin{cases} \xi = 0 \quad , \quad \tau = 0 \\ \dfrac{\xi(0)}{\sqrt{N}} \quad , \quad |\tau| \geq \tau \end{cases}$$

where

$$N = \text{sequence length} = T/\tau = 2^n - 1$$

and where $\xi(0)$ is the energy in one period. When we are, then, interested in transmitting only a single period before switching to a different channel, we have :

$$z_0(t) = \begin{cases} 0 \quad , \quad 0 \leq t \leq T \\ z(t) \quad , \quad \text{otherwise.} \end{cases}$$

The autocorrelation for this aperiodic $z_0(t)$ sequence has been shown to have properties similar to those of a periodic sequence, i.e. :

$$\xi_0(\tau) = \int_{-\infty}^{+\infty} z_0(t) \, z_0(t - \tau) \, dt$$
$$\approx \begin{cases} \xi_0(0) \quad , \quad \tau = 0 \\ \dfrac{\xi_0(0)}{\sqrt{N}} \quad , \quad |\tau| \geq \tau \end{cases}$$

and the estimate is of the form

$$\widehat{Q}(\tau) = \left| \int_{-\infty}^{+\infty} r(t) \, z_0^*(t - \tau) \, dt \right|^2 - b$$

where b is the bias term. The integral in this equation can be mechanized by multiplying the original sequence (which can be generated using phase-reversal keying, thus implying that $z(t)$ is real) against the in-phase and quadrature components of the received waveform. This is followed by square-law (or possibly, linear) detection and by an integrate-and-dump procedure, using a low pass filter with time constant greater than T. To implement the scheme for a set of τ values, requires parallel processing and possibly the use of a tapped delay line. The number of taps depends on the product of the signal bandwidth x the delay range (τ) to be observed. Figures 9 through 13 indicate a possible mechanization approach for the transmitter and for the receiver. Not included in the block diagrams are the logic units required to switch through the set of frequencies at which the measurements must be carried out.

3. Doppler spread measurement

The Doppler spread of the channel is best observed using a CW source. However, the fading rate in HF ionospheric channels is relatively small ($B_{tot} \cong 1$ Hz) so that an efficient measurement of the Doppler spread requires the use of a sequential sampling approach, in cases, such as ours, that a group of channels must be monitored. Individual channels have to be probed periodically at intervals T_0 ($T_0 \gtrsim 1/B_{tot}$) to give useful estimates. By probing at a faster rate than required ($T_0 < 1/B_{tot}$), information is obtained which is redundant in terms of the Delay Power Spectrum estimate, but which may still provide useful data for Doppler spectrum estimation.

Let's consider the sequence of outputs $\{f(nT_0)\}$ obtained when the equation

$$\widehat{Q}(\cdot) = \left| \int_{-\infty}^{+\infty} r(t) \, z_0^*(t - \tau) \, dt \right|^2 - b$$

is implemented. For a particular τ, we have :

$$f(nT_0) = \widehat{Q}_{nT_0}(\tau) \qquad \text{(estimate at } nT_0).$$

Then, it can be shown that, if the sidelobe interference is neglected, we have :

$$f(nT_0) = \left| g(nT_0, \tau) \right|^2.$$

Thus, any spectral information obtained from this sequence of outputs will relate to the shape of the scattering function for this particular delay τ . If the Signal/Noise data are used, the complete spectrum is involved, and from the previous equation, we have:

$$f\ (nT_o)\ =\ \left|A\ w(nT_o)\right|^2\ =\ A\left|\int g\ (\ nT_o,\ \cdot\ )\ dt\ \right|^2$$

Because of the inherently higher Signal-to-Noise ratio, it appears that the latter sequence offers the best alternative for spectrum analysis. In addition, the total scan time T_o required to probe all channels once, is significantly less than for the wideband coded sequence, thus providing higher sampling rate and lower danger of aliasing effects.

Before discussing spectral measurement schemes usable in our case, it should be noted that the phase instability of the oscillators used at each terminal, if not kept under acceptable levels, imposes the use of an incoherent Doppler measurement approach. If phase coherence can be assumed for the link, then the in-phase and quadrature components generated at each T_o seconds could be used to obtain the spectral properties directly.

Concerning then the estimate of the Envelope Correlation Function, we note that for a complex Gaussian process, the envelope correlation function is closely related to the process correlation function. If f(t) is complex Gaussian, and

$$\overline{f(t)\ f^*(u)}\ =\ R_f\ (\ t\ -\ u\ )$$

then :

$$\overline{\left|f\ (t)\right|^2\left|f(u)\right|^2}\ =\ R_f^2\ (0)\ +\ \left|R_f\ (\ t\ -\ u\ )\right|^2.$$

Thus, if the average power is already known, an estimate of $\left|R_f\right|^2$ follows from an estimate of the correlation function for $\left|f\ (t)\right|^2$. If we let $\widehat{R}_{|f|}^2\ (t\ -\ u\ )$ to be the estimate of $\overline{\left|f(\ t\ )\right|^2\left|f\ (\ u\ )\right|^2}$, then we have :

$$\left|\widehat{R}_f\ (t\ -\ u\ )\right|^2\ =\ \widehat{R}_{|f|}^2\ (\ t\ -\ u\ )\ -\ R_f^2\ (0).$$

Note that , without imposing further constraints, the estimate may be negative, particularly for insufficient averaging.

Concerning the implementation of this scheme, the sequence $f\ (nT_o)$ is first divided into blocks, the length of which determines the resolution of the spectral estimate. For example, a Doppler resolution of 0.2 Hz requires 5 seconds of data. Each block of data, 5 seconds long, is used to provide a single estimate of $\left|R_f\ (\tau)\right|^2$, or its Fourier transform. Thus, in this example, 12 individual estimates are performed and averaged in 60 seconds. The sequence of samples in each block can be used to find either the correlation function or the spectrum, directly. If a spectral analysis of the samples is performed, the form of the equation above suggests that the resulting spectrum will contain an impulsive term at the origin, and generally have a width of about twice the width of the spectrum of f(t). Note that the sampling rate $1/T_o$ must be consistent with the maximum Doppler spread expected, in order to avoid aliasing effects.

Several techniques for real-time estimation of spectrum parameters are illustrated by Bello (1965,[9]). These techniques make it possible to perform the measurement of the center frequency and of the rms bandwidth of a narrow-band process. The center frequency is the centroid of the power spectrum of the process, while the rms bandwidth is twice the radius of the power spectrum. The same technique may be applied to $\{f\ (nT_o)\}$ so that the width and the location of the spectrum may be determined. We briefly summarize here under the measurement technique and the effect of additive noise for a complex low-pass process g (t). The results will then be specialized to the real envelope process. Let

$$\begin{cases} g(t)\ =\ x\ (t)\ +\ j\ y\ (t) \\[4pt] <\dot{x}>\ =\ \dfrac{dx}{dt} \\[4pt] <\dot{y}>\ =\ \dfrac{dy}{dt} \end{cases}$$

with the triangular brackets denoting time averages. From the analytical conclusions of Bello (1965, [9]), we can readily see that the centroid of the power spectrum of g (t) is given by :

$$c\ =\ \frac{1}{2\pi}\ \frac{<x\dot{y}\ -\ y\dot{x}>}{<x^2\ +\ y^2>}\ =\ \frac{\int v\ S(v)\ dv}{\int S\ (v)\ dv}$$

where S (v) is the spectrum of g(t). The rms bandwidth is given by :

$$D\ =\ \frac{1}{\pi}\ \sqrt{\frac{<\dot{x}^2\ +\ \dot{y}^2>}{<x^2\ +\ y^2>}\ -\ c^2}\ =\ 2\ \sqrt{\frac{\int (v\ -c\ )^2\ S\ (v\ )\ dv}{\int S\ (v)\ dv}}$$

The influence of noise is readily taken into account by noting that the complex additive noise $N(t)$ is a complex Gaussian process like $g(t)$, so that the measurement will produce the centroid and the rms band-width of the power spectrum of $g(t) + N(t)$, namely:

$$\widetilde{C} = \frac{\int \nu \left[S(\nu) + P_N(\nu) \right] d\nu}{\int \left[S(\nu) + P_N(\nu) \right] d\nu}$$

$$\widetilde{D} = \sqrt{\frac{\int (\nu - C)^2 \left[S(\nu) + P_N(\nu) \right] d\nu}{\int \left[S(\nu) + P_N(\nu) \right] d\nu}}$$

where $P_N(\nu)$ is the power spectrum of $N(t)$. The in-phase and quadrature components (x , y) can be determined by multiplying the received carrier by both a local carrier and a 90^0 shifted local carrier at the same frequency, and then extracting the low-frequency components . Strictly speaking, D is independent of mean Doppler shift, and thus precise knowledge of the received carrier frequency is not necessary. How-ever, as the local carrier frequency departs from the received carrier frequency, the extracted x(t) and y(t) increase in bandwidth, thus requiring filters with larger bandwidth, that let more noise pass through. Thus, from the point of view of maximizing the Signal-to-Noise ratio, it is desirable to keep the local carrier frequency as near as possible to the received signal frequency.

We consider now a simpler technique for the measurement of Doppler spread. It uses only the envelope or, more generally, any well-behaved non-linear function of the envelope of the received carrier. For this technique to be strictly correct, it is necessary to assume that the transmission of a carrier results in the reception of a narrow-band Gaussian process. However, a slight departure from this condition would not affect significantly the measured parameter. It was demonstrated by Bello (1965, [9]), that, if e(t) is some non-linear function of the envelope $x^2 + y^2$ of the received carrier, the rms Doppler spread is given by

$$D = \frac{1}{\iota} \sqrt{\frac{<\left[\dot{e}(t)\right]^2>}{<\left[e(t)\right]^2>}}$$

where ι is a constant dependent upon the properties of the non-linear device used. Thus, if we describe this non-linear device by the function $k(.)$, we have :

$$e(t) = k(x^2 + y^2)$$

and it is shown by Bello (1965, [9]) that

$$\iota = \sqrt{\frac{2 \int_0^\infty r e^{-r} \left[dk(r)/dr\right]^2 dr}{\int_0^\infty e^{-r} k^2(r) dr}}$$

In the case of a linear envelope detector, $\iota = 1/\sqrt{2}$ and for a square-law detector, $\iota = 1$. The formation of the derivatives of the envelope from sampled data requires particular care (Bello, 1965, [9]).

4. Summary and recapitulation of channel evaluation techniques

a) Measurement of Signal and Noise

This can be best be achieved by using an "off-on" keyed signal, with the noise measurement preced-ing the signal measurement at any given frequency. Choosing for the noise measurement integration time the values indicated in Table I, the estimation of σ_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 measure-ment accuracy for signal level estimation is limited by the channel fluctuations, i.e. :

$$\frac{Var(\widehat{A})}{A^2} = 1/N_{eq}$$

where N_{eq} is the number of independent observations available in one sounding scan times(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 $B_{tot} \approx 1$ Hz, we have that $N_{eq} = 300$. The noise level estimate has a ratio

$$\frac{Var(\widehat{\sigma}_n)}{\sigma_n^2} = \frac{1}{TW} = 7.86 \times 10^{-4}$$

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, in page 9-3, 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

$$p(t) = \begin{cases} 0 & , \text{ otherwise} \\ 1 & , 0 \leq t \leq \tau \end{cases}$$

2. Coded sequence (phase reversal keying)

$$z_o(t) = \sum_{k=1}^{N} a_k \; p(t - k\tau)$$

where

$$a_k = \pm 1$$
$$N = 2^m - 1$$
$$T = N\tau = \text{total duration}$$

The coded sequence gives the same performance as a single pulse for equal energy; i.e., if $P_1 \tau = P_2 T$ (neglecting degradation due to sidelobe structure of sequence autocorrelation function).

Concerning the measurement of Q (τ) 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 :

$$\frac{\text{standard dev. } (\hat{Q} (\tau))}{\text{mean } \hat{Q} (\tau)} = \frac{1}{\sqrt{N_{eq}}}$$

In presence of additive noise, performance depends upon the shape of the true Q ( ) being measured. Let's consider an idealized function consisting of M equal rectangular modes:

$$Q (\tau) = \begin{cases} \dfrac{1}{ML_o} & ; \text{ during mode} \quad (L_o \text{ is the duration of each mode}) \\ 0 & ; \text{ elsewhere} \end{cases}$$

Then, at the mode centers, Q (τ) can be estimated with the formula:

$$\frac{\text{standard dev. } (\hat{Q} (\tau))}{\text{mean } (\hat{Q} (\tau))} = 1 + \frac{\sigma_n}{E \, Q (\tau)}$$

where

σ_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 (τ) 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 exclusi vely 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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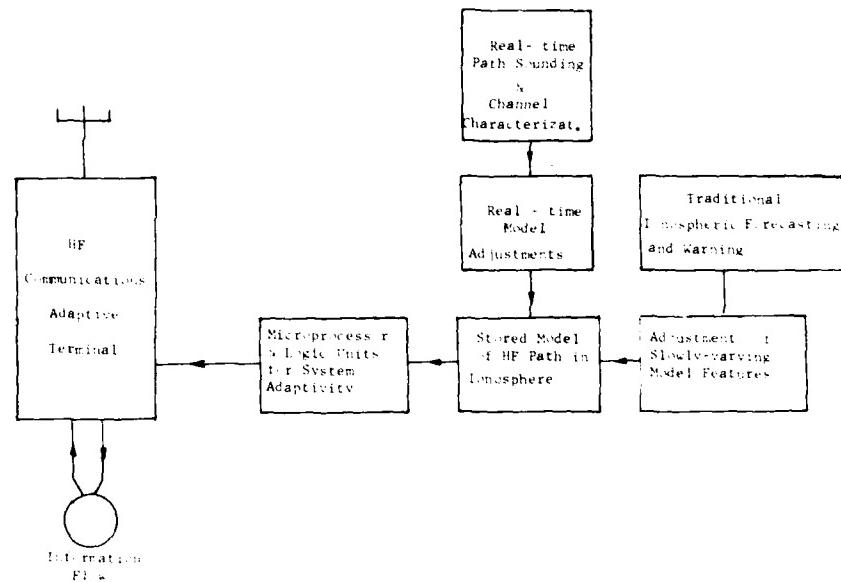
Fig 1   Block diagram of an adaptivity approach that uses traditional HF ionospheric
          forecasting and warning, in addition to real-time path sounding and channel probing
   (taken from Aarons and Grossi, 1982 [3])
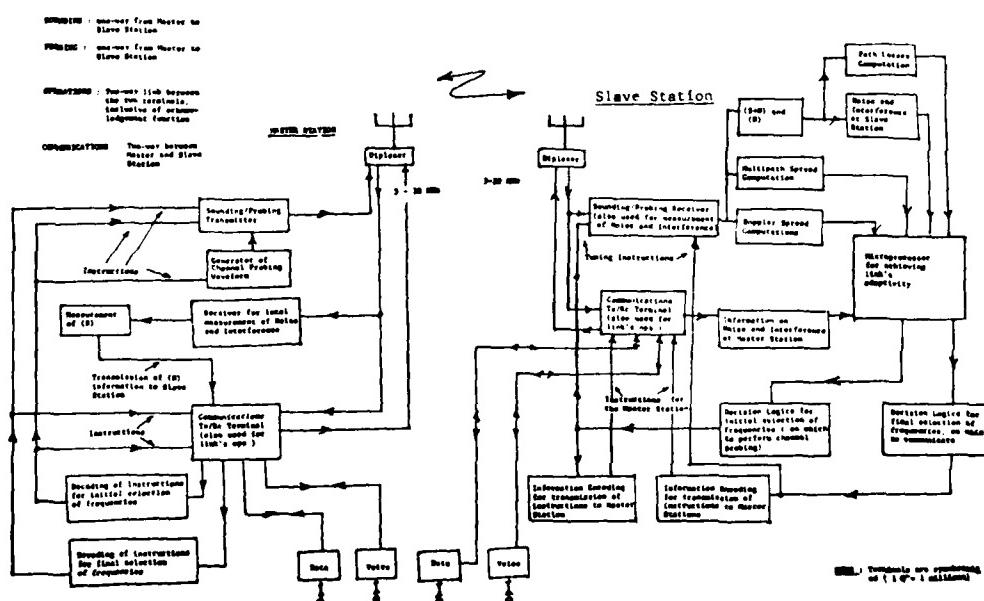
Fig.2   Simplified block diagram of the two terminals of an adaptive link at HF
         (taken from Gupta and Grossi, 1980, [11])
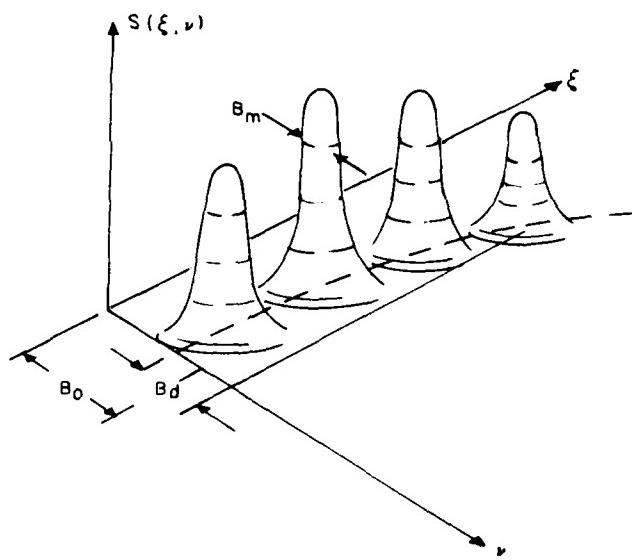
Fig 3   Multimodal scattering function S (ξ, ν)
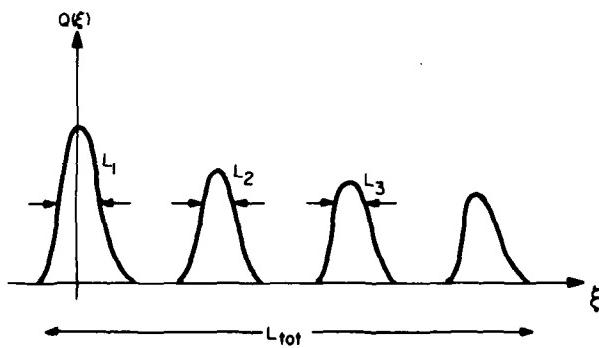
(≈ 100 μ sec – 1.6 ms)

Fig.4   Multimodal delay power spectrum Q (ξ)
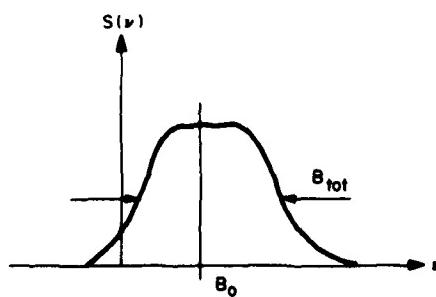
Fig.5   Single mode Doppler power spectrum S (ν)
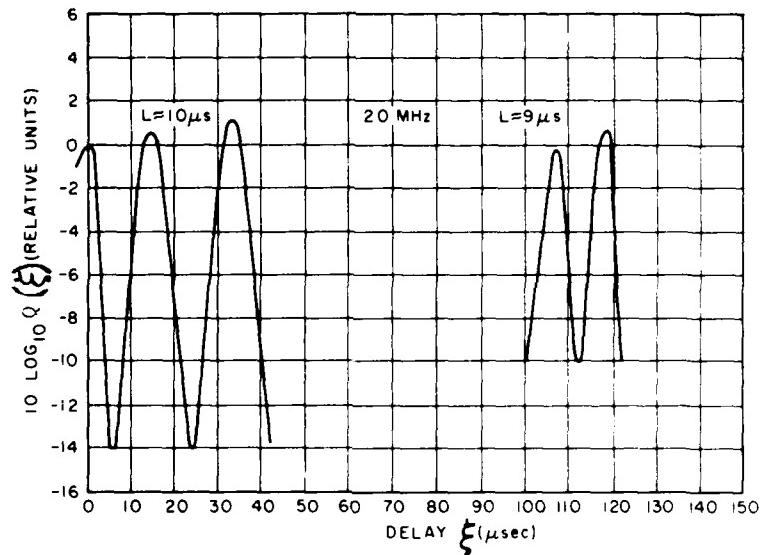
Fig 6   Multipath structure in a one-hop ionospheric path in mid-latitude
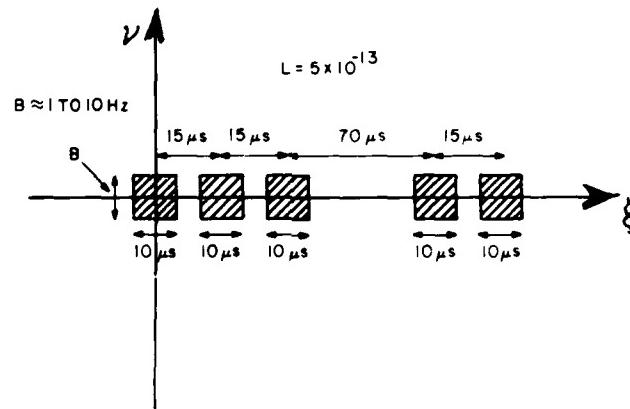
Fig. 7   Cross section in the plane (ξ, ν) of the scattering function of the path in Figure 6

σ(τ,f)

TARGET SCATTERING FUNCTION

∫dτ

σ(f)

ECHO POWER SPECTRUM

∫df

σ(τ)

POWER IMPULSE RESPONSE

FOURIER TRANSFORM

FOURIER TRANSFORM

FOURIER TRANSFORM

R(Δτ)

1/B

ECHO
CORRELATION FUNCTION

Δf = 0

R(Δf, Δτ)

Δτ = 0

R(Δf)

1/L

SPACED-FREQUENCY
CORRELATION FUNCTION

1/L

1/B

Δf

TWO-FREQUENCY CORRELATION FUNCTION
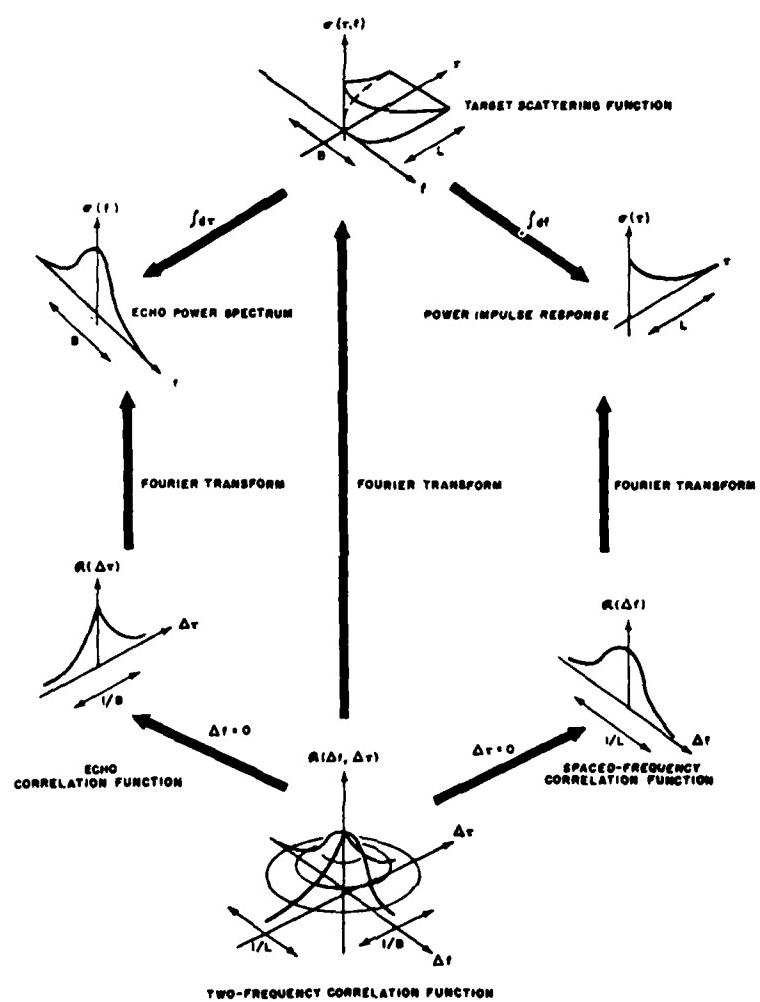
Fig.8   Summary of the interrelationships of the target (channel) scattering function
and other descriptive functions (taken from Green, 1968, [5])
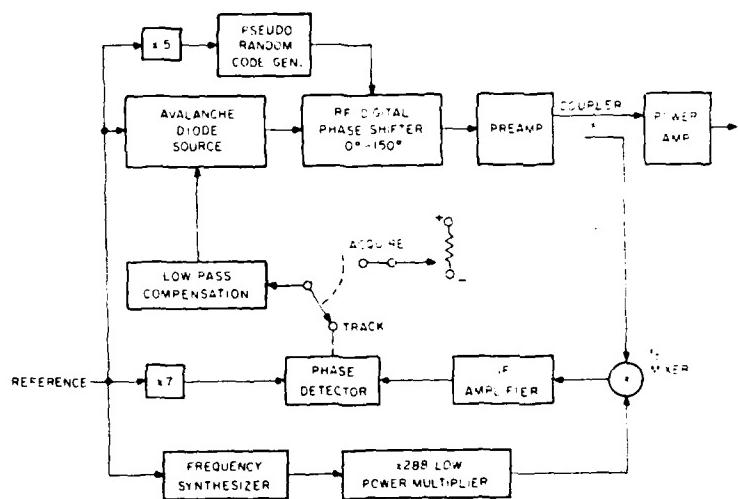
Fig.9   Block diagram of transmitter
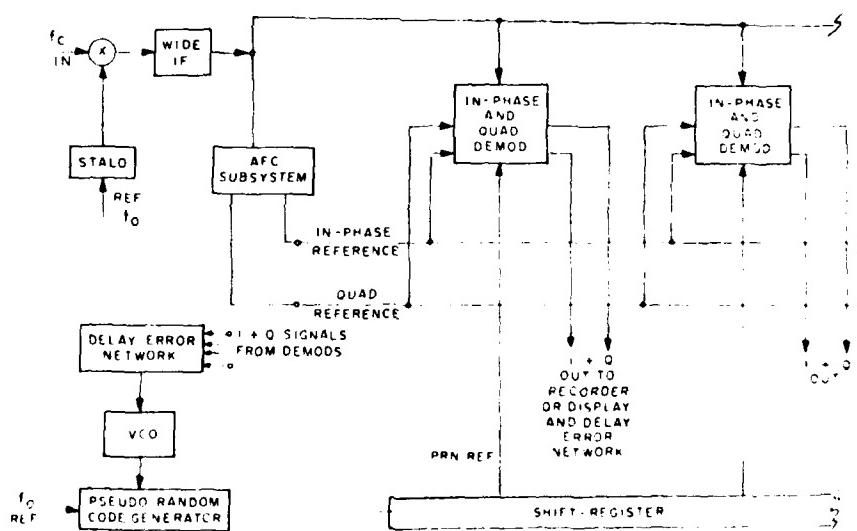
Fig.10   Block diagram of receiver
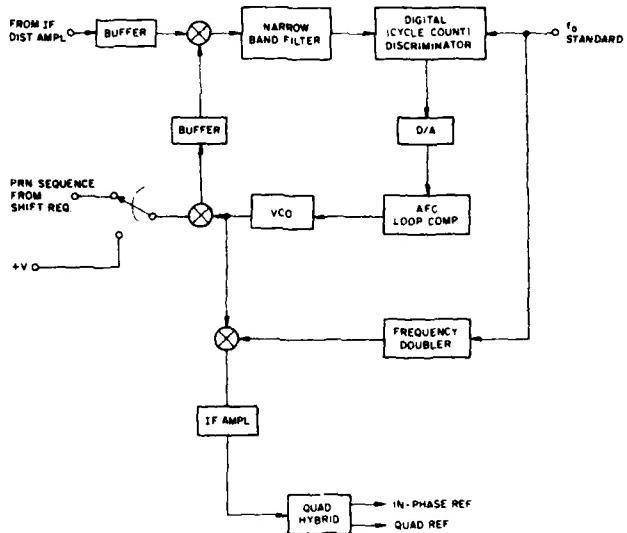
Fig.11   Block diagram of Automatic Frequency Control (AFC)
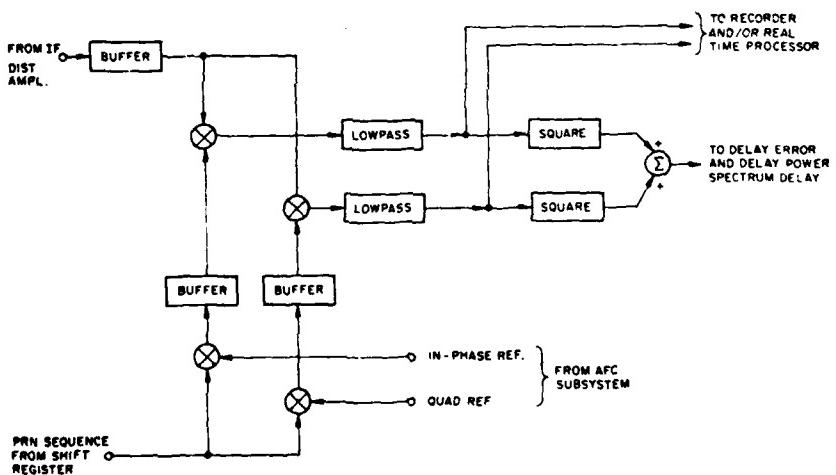

Fig.12   Block diagram of I and Q demodulators
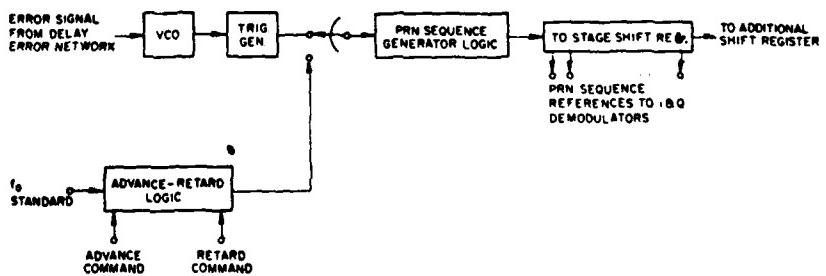
Fig.13   Block diagram of PRN generator and shift registers

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 interconnect. 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 usuable channels on a paper 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 RTCF systems have become more sophisticated and an integral part of the communication transmitter receiver equipment. Micrprocessors 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) Contoller 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. Incorpoated 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 th Remote to ring the called subscriber. The Remote checks the status of the line and if free 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 fl 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 fl and indicates the call will be made on another frequency f2. The Master acknowledges message and returns to idle condition on 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
- Syncompex 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 system 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 "psuedo 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 were 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 broad band 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 messages. 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 on the keyboard which appears on the terminal display panel. The outgoing messages 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 ooprator 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 impemented with a microprocessor and free from drift, aging and 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 the voice channel enabling more than one network to be operational in a 3kHz 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 recognize 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 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 existing HF radio systems to transmit short text messages. The terminal can be connected easily to most conventional HF systems. Modulation and coding 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 messages and optional communications privacy.

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RACE

Subscriber Line Interface System

with the Remote Stations

Figure 1

B

MODERN HF COMMUNICATIONS

UTTL: Square root kalman filtering for high-speed data
received over fading d spersive HF channels

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C m btic ns Systems, Dev   H   Heights, MA.
IEEE Transa tions or Internati C r   cat   vol 1 28.
S ptt 1982. pt   p 75 73.
This Pu   has hee to apply Square r st kalman
filtering to high speed data transmise on on fading
dispersive high-frequency (HF) channels high
d t r   HF transm ssion can te a hi ed by use of a
s ngli   er band width ff c m   c   lation that is
d   s   d with d v lu ion ion   cc t lation (CFE).
 he DFE   ff cients are add   t l   by square
root   n alt n t s   h   l   l ion an arg
m a ization procedures for the   k s root kalman
a gorithms are given. Computer c ulat on indicate
that good error-rate performance can be achieved with
the a gorithms for rapid fading HF radio channels.
82 09/00 82443574

UTTL: Change in the ionization of the F layer under
the effect of high-power radio waves

AUTH: A/COLIAN. S. F.; B/DIMANT IA S.; C/KIK. V. IU.;
D/LOBACHEVSKII. L. A.; E/NIGULIN V   P   I   ;
F/PANCHENKO. V. A.; G/PETROV. S.; H.FOLIMATIDI.
V. P. I/BELINGER. I S. J IU   M I   P4A
 kademi a N uk SSSR. Insti t Zem o a netizma,
rosfery i l'asoro trant n   r tiion A-a emgor cook.
I chno-Issledovat   ki   Inst   t Radio. Moscow.
USSR) Geo agnetizm i Aeronomi a, vol 22. July-Aug.
1982. p. 675-677. in Russian
Multifrequency sounding was used to investigate
density disturbances caused in e F b er maximum
r ion by high-power radi   aves at a frequency of 4.2
MHz. which was clo e to c   c   c   critical
frequenc.) It is shown that a type of disturbance
leads to an appreciable incre e in the magnitude
of the disturbance and the charactor of it's temporal
change agree well with the theoretical results of
Gurevich and Shvart burg (1973) and Ikin and Krotova
(1981). 82/08/00 82443431

UTTL: Solid state high frequen   F   e
AUTH: A/COTTLIEB. I. M   Reston. VA   A rtech Publishing Co..
Inc.. 1982. 255 p
A comparison is conducted between tube and solid-state
technologies, taking into account the characteristics
of radio frequency power   r   e   of radiation
advantages s solid-state RF pow   d vices, the class
C tube R amplifier   the trans stor vers on of the
grounded grid amplifier, a solid-state approach to
high power reliability, tests circuits and

UTTL: Industry looks at m litary exceptionally high
frequency bands
AUTH: A/HOFFMAN. F. I   P.A. A/HCOMSAT General Corp..
Xash region, DC; Military,
Elec r onics/Countermeasures. vol 8 Oct. 1982. p. 38.
42 45.
ABS: Various programs and systems in the developing EHF
communications band usage. including satellites, are
examined. Commercial systems. built by orpanies which
also hold military contracts, are beginning to feature
capabilities formerly reserv d for the military. In
order to et increased c pac ty for voi e video. and
data links. The C-band, from 4 6 GHz is becoming
oversubscribed. leading to the implementation of the
Ku-band. from 11-14 GHz. Saturation of these bands is
expected to occur in the early 1900s. The Japanese
have been experimenting with the 20/30 THz bands since
1970. the Italians are planning th italsat. and ESA
is developing the E-Sat all in the same frequenc/
bands. NASA explored the regime partially with the
ATS 8 spacecraft. but funding has lagged. Plans exist
for an Advanced Communications Technology Satellite.
operating up to 50 GHz. to be launched in 1987 and
function for two years. It is sugg sted that military
users may benefit from leasing dedicated commercial
satellite channels. while the comm rcial operators
will profit from greater operational use of the
equipment. 82/10.00 82443498

UTTL: Measured spectral amplitude of lightning sferics
in the HF. VHF. and UHF bands
AUTH: A/SHCWFERT. T. H.; B.HONELL. M. A.; C/LOTT. G. K.;
JR (AA B nburn University Aubturn. Ala.; C/U.S.
Navy. Chesapeake VA) Cope Aut on Univ. Ala.
IEEE Transacc ols on Elec r Aromat c Compatibility.
vol EMC-24 Aug. 1982. p 256. 261.
ABS: An experimental study was carried out to measure the
spectral amplitude of lightning sferics at 22.5 MHz.
27 5 GHz and 2 25 GHz. This effect included the
detection storage and analysis of RF emission data
to accurately determine the relative electric field
s re ith frequency the m   mum amp o area with a
typical local thunderstorm. These measured spectral
amplitudes are normalized and plotted relative to
similar data from other investigators. These data
surport a 1/F to the 2nd to 1/F to the 3rd slope for
s r ics at frequencies above the HF range. 82/08/00
82443466

systems can be designed to be more robust than HF and more economical than satellite communication systems.
82/05/00 #2432025

UTTL: Disturbance of the ionospheric F-region by powerful radio waves.
AUTH: A/KOZLOV, A. G. PAA: A/Tomskii Gosudarstvennii Universitet, Tomsk, USSR) Radiofizika, vol. 25. no. 2, 1982. p. 127-132. In Russian.
ABS: The paper considers variations of the main parameters of the F-region under the effect of high-power radio waves. The problem is based on a nonstationary mathematical model of an inhomogeneous plasma, producing a self consistent description of the composition dynamics and absorption of charged neutral and metastable components. It is shown that the effect of high-power radio waves on the F-region has a nonstationary character, and that the relative contribution of various mechanisms to the formation of the F-region structure is time-dependent, the duration of pumping. 82/00/00 #2231013

UTTL: The behavior of the signal level at frequencies exceeding the maximum usable frequency.
AUTH: A/DUMBRAVA, Z. F. PAA: A/(Akademia Nauk SSSR, Severo vostochnyi Kompleksnyi Nauchno-issledovatelskii Institut, Magadan, USSR) Radiotekhnika i elektronika, Nov. 1982, p. 1 Fizike Sollstiu no 50, 1982. p. 893- In Russian.
ABS: The behavior of the signal level at the input of a receiver when the maximum usable frequency is slightly exceeded is investigated. A precipitous drop in excess of 40 dB/MHz is observed in the level. In the analysis, records of signal amplitude and maximum usable frequency during the same period simultaneously obtained by the method of oblique us multifrequency sounding are used. 82/00/00 82A30254

UTTL: An electromagnetic environment on an aircraft.
AUTH: A/(de P. D. B. Avison) PAA: A/(Marconi Avionics Ltd., Rochester, England) IEE/Royal Aeronautical Society Conference on Electromagnetic Compatibility, Guildford, England, September 16-18, 1980 Proceedings. (A82-24351 10-33) London, Institution of Electronics and Radio Engineers, 1980. p. 280-291.
ABS: This paper reports an EMC analysis program are carrying out a study aimed at defining the operational

low-frequency solid-state transmitters. The bipolar transistor in RF power applications is considered, giving attention to gain leveling provisions, matching networks, the effect of transistor and tube differences on tuned circuits, output circuits for medium and high-power applications.
Attention is also given to the field-effect transistor in RF power applications, Particular attention being given to low-power applications of transmission-line elements to RF power circuitry, low-power applications, and medium- and high-power applications. 82/09/00 82433645

UTTL: Attenuation with distance and wind speed of HF surface waves over the ocean.
AUTH: A/FORGET, P.; B/BROCHE P. C/DE MAISTRE J. C. PAA: C/(Toulon, Universite, Toulon, France) Radio Science, vol. 17, May-June 1982. p. 599-610. Research supported by the Institut National d'Astronomie et de Geophysique. Centre National de la Recherche Scientifique.
ABS: An HF monostatic experiment is performed in order to evaluate the relative attenuation with distance of radio waves propagating in the ground wave mode over the sea surface, at 7 and 14 MHz. Clear evidence of the wind causing the sea roughness) influence arises from the data, particularly at 14 MHz. The results are compared to Barrick's theoretical calculations, and a good agreement is found for the value of attenuation rates (in decibels per kilometer) and their variation with the wind speed in the downwind direction. Some discrepancies in the data are partially interpreted as due to a fetch influence upon the radar cross section. 82/06/00 82A35037

UTTL: Communications through troposcatter scattering
AUTH: A/INCE A. N. PAA: A/(Istanbul Technical University, Istanbul, Turkey) IEEE Communications Magazine, vol. 20, May 1982. p. 27-43.
ABS: Various scatter systems are reviewed with reference to their underlying physical mechanisms and communications applications. Attention is given to tropospheric scatter, ionospheric scatter, meteor burst, aircraft wind speed scatter, rain scatter, and recently, the scatter channel scattering by aircraft, the scatter channel characteristics diffused and fading, transmission loss, long- and short-term fading, signal distortion, and antenna gain degradation. Also given are the capacities of the scatter systems and the ranges they can put them to the best use, and how the conventional and modern satellite communications. Scatter

electromagnetic environment for aircraft and flight weapons systems. The UK land-based HF environment is described, and methods for predicting field strengths and power densities both in the near and far fields are outlined. Attention is given to the types of arrays commonly used in connection with high-power transmitters, and their principles of operation. Curtain arrays, log-periodic arrays, and rhombic arrays are considered, and typical field strength patterns are given for each type. 80/00/00
82A24365

UTTL: HF system design principles for efficient spectrum utilization
A/CAPPELL, M. PAA: A/(Ministry of Defence/Procurement Executive, London, England) In: International Conference on Radio Spectrum Conservation Techniques, London, England July 7-9, 1980. Proceedings. (A82-23805 09-32) London, Institution of Electrical Engineers, 1980, p. 1-5.
80/00/00 92A23802

UTTL: Investigation of decametric radio wave propagation on transpolar paths during auroral substorms
A/ELE-OVESHCHENSKAIA, N. F.; B/EUENOV, V. A. Radiofizika, vol. 24, no. 11, 1981, p. 1299-1305. In Russian.
ABS: The results from an investigation of the characteristics of the long-range propagation of decametric waves on three transpolar paths are presented. The study was carried out at various phases of the auroral substorm during winter nights in the period 1974-1978. The first radio line covered a distance of 7200 km and had a working frequency of 15 MHz; radio line two had a distance of 8600 km and working frequencies of 12 and 17 MHz; the third path had a distance of 7600 km and a working frequency of about 17 MHz. The distance from the point at which the signal was received to the auroral zone was 670-800 km. Stable regularities in variations of the HF signal intensities were observed during the substorm.
81/00/00 82A18376

UTTL: Parametric decay of high intensity radio waves of high intensity in an ionospheric plasma
A/ABYZOV, A. S.; B/SHAAI, A. S.; C/SOLOGUVNIKOV, G. K. Geomagnetizm i Aeronomiia, vol. 21, July-Aug. 1981, p. 502-606. In Russian.
ABS: The parametric instability of ionospheric plasma in a field of intense radio waves in the meter range is

discussed. In particular, the decay of high-frequency radio waves into Langmuir and transverse electromagnetic (i.e., stimulated Raman scattering) and into ion-acoustic and electromagnetic waves (i.e., stimulated Brillouin scattering) is considered. A calculation of the threshold fields and increments of perturbation growth numbers is presented for the case of waves emitted by ground-based transmitters. It is shown that when the amplitude of the transverse wave is not less than 20 V/m, more than half the energy of the original impulse is transferred to the scattered wave. The spectrum of signal scattering shows a red shift (relative to the frequency of the original signal) to frequencies of excited ion-acoustic waves. The form of the spectrum corresponds to those obtained in laboratory experiments. 81/00/00 81A55103

UTTL: Spectral observations of high frequency radio emission during the Zarnitsa-2 experiment on electron beam injection into the ionosphere
AUTH: A/ZHULIN, I. A.; B/MARKEEV, A. K.; C/MISHIN, E. V.; D/KUZHIN, IU. IA.; E/FOLOUD, L. V. PAA: E/(Akademiia Nauk SSSR, Institut Zemnogo Magnetizma, Ionosfery i Rasprostraneniia Radiovoln, Krasnogorodok, USSR) Fizika Plazmy, vol. 7, Mar.-Apr. 1981, p. 575-582. In Russian.
ABS: Results of the observations of high frequency radio emission in the range 27-51 MHz produced by electron-beam injection into the ionosphere are reported. The dynamic characteristics of the radio emission are explained in terms of a plasma-beam discharge in the vicinity of the rocket. 81/00/00
81A40392

AUTH: UTTL: High-frequency radio wave measurements with the WISP
A/STORED, D.; B/........
A/WEBB, P. G. PAA: A/(Department of Communications, Communications Research Centre, Ottawa, Canada) In: ...... fund 2nd ISEAS .... on active Experiments in Space, Alpbach, Austria, Hungary, June 2-14, 1980.) Advances in Space Research, vol. 1, no. 2, 1981, p. 241-243.
ABS: It has been proposed to apply the WISP for .... The system .... and the transmitter generates signals with variable frequency level, and modulation. The transmitter works into a dipole of variable length (up to 300 m tip-to-tip) and which matched, can deliver 0.5 kW peak power. One receiver is located on the platform and another on the

Recoverable Plasma Diagnostics Package subsatellite. WISP/HF will be used to test a number of fundamental concepts connected with propagation in anisotropic media, including phase and group delay, polarization, power flux, and radio coupling. There is strong interest in research on other and magnetospheric structure as the basis for understanding the dynamics of the neutral atmosphere and the plasmasphere.
81.00/00   81A25057

AUTH: The World Administrative Radio Conference 1979 - Results and impact on defense and national security
A/COLE, A. J.; B/FAIRHEAD, O. L.   PAA: A/(U.S. Department of Defense. Office of the Assistant Secretary of Defense. Washington D.C.).
B/(Systematics General Corp., Falls Church, Va.). In: ICC '80: International Conference on Communications. Seattle, Wash., June 8-12, 1980. Conference Record Volume 1. (A81-32276 14-32) New York, Institute of Electrical and Electronics Engineers, Inc., 1980. p. 7.5.1-7.5.4

ABS: An analysis is presented of the possible impact of the results of WARC-79 on future military communications systems. Specific reference is made to DOD activities in the fields of high frequencies, navigation satellites, radar and weapon systems, terrestrial fixed and mobile services, satellite communications, and the geostationary orbit.
80/06/00   81A32241

AUTH: High-frequency pulsed breakdown of air in intersecting radio beams
A/BORISOV, N. D.; B/GUREVICH, A. V.   PAA: B/(Akademiia Nauk SSSR, Institut Zemnogo Magnetizma, Ionosfery i Rasprostraneniia Radiovoln, Akademgorodok, USSR). Geomagnetizm i Aeronomiia, vol. 20, Sept.-Oct. 1980, p. 841-847. In Russian.

ABS: The production of an artificial ionization region in the atmosphere by means of the multiple breakdown of air in intersecting radio beams is examined. The effects of wind, diffusion, and recombination on the formation and disruption of a steady-state ionization region are considered. It is shown that the ionization region can consist of one or several narrow layers.
80/10/00   81A1126?

AUTH: Calculation of HF radio signal parameters on the basis of tabular models of the ionosphere
A/TOKAR, V. G.; B/ANDREZEN, Iu. I.; C/NIKITIN, M. A.   PAA: C/(Kaliningradskii Gosudarstvennyi Universitet, Kaliningrad, USSR). Geomagnetizm i Aeronomiia, vol. 20, July-Aug. 1980, p. 753-756. In

Russian.   80/08/00   80A45057

UTTL: The effect of an artificially disturbed region of the ionosphere on the oscillation of HF signals.
AUTH: A/EREMENKO, V. A.; B/...; C/...
IU A., B/MATIUGIN, S. N.   PAA: A/(USSR, Institut..., Moscow). Radiofizika, vol. 23, no. 4, 1980, p. 502-504. In Russian.

ABS: When the F region is heated by intense HF signals from a ground transmitter, an artificially disturbed region is formed which has a focusing effect on radio signals. It is here proposed to use measurements of secondary transmitted signals for the detection of artificial perturbations in the ionosphere from 100 km long. The data were averaged to identify the focusing effect on the background of natural and shade fluctuations. The results may be of use in the observation of large-scale disturbances of the ionosphere.
80/00/00   81A26546

AUTH: Unresolved theoretical problems in design of optimal low-pass filter for harmonic suppression in radio transmitter output
A/SOKAL, N. O.   PAA: A/(Design Automation, Inc., Lexington, Mass.). IEEE Transactions on Circuits and Systems, vol. CAS-27, Mar. 1980, p. 235.
80/03/00   81A31774

AUTH: Power MOSFETs for medium-wave and short-wave transmitters
A/IKEDA, H.; B/ASHIKAWA, K.; C/URITA, K.   PAA: ... IEEE Transactions on Electron Devices, vol. ED-27, Feb. 1980, p. 131-134.

ABS: This paper describes design and performance of a high-frequency power MOSFET used in switching-mode power amplifiers in the medium-wave (500 kHz to 1.5 MHz) or short-wave (1.5 MHz to ... MHz) transmitters. A ... of power up to more than 200 V with an ... drain-source ... operating at approximately 1 Ohm, the ... length and the ... The high-frequency power MOSFETs are optimized as well as other ... layer concentration employing the multichannel ... fabricated by RF diode sputter, the ... of geometries at high speed with turn-on and turn-off times of 25 and 25 ns, respectively. A ... fabrication evaluation test was performed to assure the stability and reliability of the device

AUTH: UTIL: A simple method of HF mode prediction
A/BRIDGER, R. B/DAVEY, I. E. PAA: B/British
Broadcasting Corp. London, England In:
International Conference on Antennas and Propagation.
London, England, November 28-30, 1978. Proceedings.
Part 2 (A80-15812 04-32) London, Institution of
Electrical Engineers, 1978. p. 60-63

ABS: A system of HF mode predictions can be carried out by
hand, as a first step in the prediction of field
strengths, by assuming that the ray is geometrically
reflected at the height of maximum electron density of
the layer. This paper discusses the approximations
involved in this assumption, the important modes that
need to be considered for various ranges of ground
distance, and considerations involved in transmitting
to geographical areas, rather than point to point.
78/00/00 80A15901

AUTH: UTIL: HF communication problems at low latitudes due
to steep spatial and temporal electron density
gradients
A/LAKSHMI, D. R. B/AGGARWAL, S. C/PASRICHA, P. K.
D/REDDY, B. M. PAA: D/National Physical Laboratory
of India, New Delhi, India) Indian Journal of Radio
and Space Physics, vol. 8, June 1979. p. 101-105.
79/06/00 80A15074

AUTH: UTIL: Measurement of the angles of arrival of HF
signals in oblique propagation by the phase method
with Doppler filtering
A/AFRAIMOVICH, E. L. B/FILIPPO, I. G. C/VUGMEISTER,
B. O. D/GUZOV, V. P. E/ZARASOV, V. N.:
F/KOBZEV, V. A. G/KHAKHINKO, V. S. H/NOSOV, V. E.
I/PAKHOTIN, I. N. J/SUGLOV, L. P. PAA:
J/Akademiia Nauk SSSR, Inst. for Zemnogo Magnetizma,
Ionosfery i Rasprostraneniia Radiovoln Irkutsk and
Akademgorodok, USSR) (Geomagnetizm i Aeronomiia,
vol. 18, July-Aug. 1978. p. 741-744.) Geomagnetism and
Aeronomy, vol. 18, Feb. 1979. p. 565-567. Translation.
(Previously cited in issue 02 p. 215, Accession no.
A79-12948.) 79/02/00 80A14145

ABS:

AUTH: UTIL: Time- and frequency-domain characteristics of
man-made radio noise affecting HF communications sites
A/CUKNIAS, E. J., JR. B/AKHRIZGI, S., JR.
C/CLEVE, W. R. PAA: B/U.S. Naval Postgraduate
School, Monterey, Calif.) C/Systems Control, Inc.,
Palo Alto, Calif.) IEEE Transactions on
Electromagnetic Compatibility, vol. EMC-21, Aug. 1979.
p. 182-189.

ABS: The short-term properties of man-made radio noise

The test results indicate that phosphosilicate glass
polarization affects the device reliability only when
offset gate layer coating voltage is much lower than the
optimized value. 80/02/00 80A23177

AUTH: UTIL: Low-noise amplification in tactical radio sets
A/MERANT, C. PAA: A/Thomson-CSF, Division
Telecommunications, Gennevilliers, Hauts-de-Seine,
France) Revue Technique Thomson-CSF, vol. 11, Dec.
1979. p. 881-1005. In French. Research supported by
the Ministere de la Defense.

ABS: The paper examines noise generation in broadband
amplifiers of HF and VHF radio sets, and proposes some
solutions for low-noise amplification. Low-noise
performance has been obtained with amplifier models
operating in the 1.5-30 and 20-80 MHz ranges. This
performance has been achieved by the analysis of noise
generation in active components and by the combination
of low-noise amplifying structures in particular, the
use of vertical-structure field-effect transistors has
been found to reduce noise. 79/12/00 80A21688

AUTH: UTIL: Electromagnetic radiation from selected
telecommunications systems
A/TELESPN, R. C. PAA: A/Bell Telephone
Laboratories, Inc., Murray Hill, N.J.) IEEE.
Proceedings, vol. 68, Jan. 1980. p. 21-24.

ABS: The paper describes the instrumentation and
measurement techniques used in a study of
electromagnetic (EMI) energy in radio transmission
systems, and summarizes the data obtained for high
frequency (HF) radio, troposcatter, obtained for high
earth-satellite and microwave radio relay systems.
Power density and electric field measurements were
made at selected facilities, including antenna tower
locations for point-to-point microwave radio
equipment. It was found that the maximum electric
field strengths associated with HF radio systems in
areas accessible to radio personnel was less than 36
V/m. The maximum power density associated with
earth stations and point-to-point microwave radio
systems were less than 1 microwatt/sq cm. In some
cases microwave radio props. in high building floors
and maximum levels due to VHF and UHF transmitters
of a few tens of microwatts/sq cm. 80/01/00
80A22C18

affecting operational HF receivers are described by
simulated three-dimensional circular noise is
categorized into types, and the importance of each
type to HF circuit operation is discussed. The types
of noise are also related to source categories to
simplify source location and control processes.
79/08/00   7 A8905/

UTTL: Some physical constraints on the use of
'carrier-free' waveforms in radio wave transmission
systems

AUTH: A/DAVIS, J. R.; B/BAKER, D. J.; C/SHELTON, J. P.;
D/DENT, W. S.   PAA: D/Div 5, Navy. Naval Research
Laboratory, Washington, D.C.   IEEE, Proceedings,
vol. 67, Jun. 1979, p. 884-890. Contents, p. 890, 891.

ABS: The paper attempts to determine whether inherently
broadband 'carrier free' waveforms such as Walsh
functions have application to radio wave transmission.
It is shown that the band-limiting constraints of
radiating systems limit emission of the LF spectral
components of such waveforms, and severely restrict the
radiation of activity that may be achieved in systems
that use them. It is also shown that dispersion in the
propagating medium poses difficulties for the use of
such waveforms, and an example is presented that
illustrates their impracticality in a case in which
medium-related limits on coherence bandwidth are of
concern.   79/06/00   79A18343

UTTL: Optimization of animal tracking frequencies
AUTH: A/HORWITZ, C. M.   PAA: A/Sydney University, Sydney,
Australia)   IEEE Transactions on Antennas and
Propagation, vol. AP-27, May 1979, p. 393-397.

ABS: The three major factors which limit the performance of
animal radio tracking systems - propagation loss,
antenna detuning, and signal variability - are
described. Propagation loss over antenna heights from
ground level to 2 m is the frequencies between 70 and
1250 MHz has been measured. It is shown high
loss at high frequency. However, an analysis of the
environmental sensitivity of small antennas shows that
high detuning losses are easily obtained at low
frequency. Using a combined radio between 1000 MHz, using a
signal scattering model it is shown that signal
variability, caused by scatter from surrounding
vegetation, rises rapidly with frequency. As a result
the lowest practical frequency is recommended. In the
case of the animal tracking system modeled here,
frequencies of 500-800 MHz appear best.   79/05/00
79A37091

UTTL: Kinetic theory of inhomogeneous plasmas situated
in RF fields.
AUTH: A/SHEVLI, R.   PAA: A/Atomic Energy Organization of
Iran, Fusion Div., Tehran, Iran. IEEE Transactions
on Plasma Science, vol. PS-7, No. 1979, p. 12-20

ABS: The stability of that an electric field increases
on drift instabilities in an inhomogeneous isothermal
plasma with temperatures of ions and temperature
this is treated in the kinetic theory. A mathematical
formulation of the problem is presented, and the
Vlasov theory is used to derive a general dispersion
relation for a multispecies inhomogeneous plasma
subjected to an externally applied RF electric field.
Under specified assumptions the effects of the drift
wave and the effect of the RF electric field are
examined for low frequency range of frequency ranges
in the medium-frequency only electrons are in
resonance and the RF electric field has an appreciable
effect on stabilization of drift waves in this
frequency range because the RF pressure is in phase
with the gas kinetic pressure, increase using an
increase in the oscillation frequency and an
enhancement of the Landau damping of the wave by
electrons.   79/03/00   79A25570

UTTL: Diurnal and seasonal variations in the duty
cycle of meteor forward scatter propagation
AUTH: A/RAO, R. S.; B/RAO, D. G. V. K.; C/RAO, B. R.;
D/RAO, E. B.   PAA: D/Andhra University, Waltair,
India)   C/University Grants Commission, New Delhi,
India)   D/Defence Electronics and Applications
Laboratory, Dehra Dun, India)   Space Sciences
Symposium, Waltair, India, Jan 10-12, 1978.) Indian
Journal of Radio and Space Physics, vol. 7, June 1978,
p. 154-161.

ABS: It is found that the average duty cycle for meteor
forward scatter propagation at 81 MHz between
Dehra Dun and Waltair is a diurnal variation of the
the order of 8. The diurnal variation of the duty
cycle is sinusoidal with a maximum at 0600 and a
minimum at 1800 hours LT, the seasonal variations
show a minimum around February and a general increase
thereafter. The main contribution to the duty cycle is
found to be for communication in the link is found
to be 0400-0600 hours LT every day.   79/06/00
79A24672

UTTL: A study of an 8MHz RB pattern in relation to
ionospheric characteristics / Gadhati
AUTH: A/REVI, M.; B/GADSARA, A. K.
University Gujrat, India. Journal of Radio
and Space Physics, vol 7, Aug 1978, p. 142-144
Research supporting by the Council of Scientific and
Industrial Research.

ABS: Faraday fadings of 40 KHz signals from INTASAT were
recorded at Gujrat during Aug-Oct 1976. Large and
small scale irregularities were studied from the RB
records. The number of large scale irregularities is
found to be greater during the daytime hours than during
nighttime. The number of large scale irregularities
correlation on the observed in size of the
irregularities is observed to be higher during
nighttime than during daytime hours. These
irregularity events do not seem to follow either
electron density variation or the diurnal pattern which
may be expected according to the thin slab scattering
diffraction theory. 78/08/00 78A24666

UTTL: Design of anode power-supply modulators of
radio transmitters based on systems of multizonal
pulse width modulation
AUTH: A/KARZEV, A. V.; B/TARASHIN, A. / Problemy
Tekhnicheskoi Elektrodinamiki, no. 60, 1978, p. 13-16.
In Russian. 78/00/00 78A20416

UTTL: Effect of weak large-scale inhomogeneities in
the auroral ionosphere on short-wavelength radio-wave
propagation
AUTH: A/GELLERG, M. G.; B/VOLLOV, R. N.; C/KUKUSHKINA, R.
S. PAA C/Akademiia Nauk SSR, Polyarnyi
Geofizicheskii Institut, Murmansk, USSR)
Radiofizika, vol. 21, no. 11, 1978, p. 1591-1596. In
Russian. 78/00/00 78A14408B

UTTL: An estimate of the economic benefit from a
civil aviation satellite communication traffic
AUTH: A/PRICE, J. W. PAA B/Royal Aviation Authority.
London, England. In International Conference on
Maritime and Aeronautical Satellite Communication and
Navigation, London, England, March 7-9, 1978.
Proceedings. (A79-17096 05-04) London, Institution of
Electrical Engineers, 1978, p. 29-32.

ABS: This paper considers the implications that a
communication satellite might have on the capacity of
an oceanic air traffic system and the corresponding
benefit in reduced operating cost penalty to the
traffic involved. The data and illustration are drawn

for the North Atlantic, which has the greatest level
of oceanic traffic and the most revolved ATC system
through the E. Pacific system approaches it. This
considers the effect of a communication satellite
on the possibilities of suitable sites for independent
navigation or surveillance are not considered.
78/00/00 78A17091

UTTL: Measurement of the fluctuations in the arrival
angle of decameter radio waves along path of various
length
AUTH: A/GUSEV, V. D.; B/GAILIT, T. A. C/L....., M. L.;
D/OVCHINNIKOVA, N. P. PAA E/M... ....
Gosudarstvennyi Universitet, Moscow, USSR
Geomagnetizm i Aeronomia, vol 18, Sept-Oct. 1978.
p. 932-934. In Russian. 78/10/00 78A16834

UTTL: JTIDS regular systems to ....
AUTH: A/GPASSE, C. L. PAA A/Lockheed ... Palo Alto,
Calif. In International Telemetering Conference,
Los Angeles, Calif., October 1977, Proceedings.
(A79-14226 03-32) Pittsburgh, Pa. Instrument Society
of America, 1977, p. 325-33.

ABS: The Joint Tactical Information Distribution System
JTIDS concept is described for application to the
military's needs for communication, navigation and
identification on equipments which ... effective
economic route. A key element ... ... to achieving
these goals is the surface acquisition via JTIDS
bandpass filter. In order to satisfy the JTIDS
requirements of today, as well as the tactical
Information Exchange System (TIES) of the future. It
is necessary to utilize state of the art SAW
resonator filter designs. It is ... with more
conventional SAW techniques, little technology attention
is given to the JTIDS application ... SAW resonator
filter tank, the realization of a three pole SAW
band filters with the aid of a three pole SAW
resonator filter, a 350 kHz with a weighted
transmission which satisfies JTIDS TACAN requirements.
and a 2.7 MHz weighted communications channel filter.
77/00/00 78A14247

UTTL: Application of adaptive array technology to HF
communications systems
AUTH: A/HANSEN, P. M. PAA A/U.S. Naval Ocean Systems
Center, San Diego, Calif.) In NTC '77, National
Telecommunications Conference, Los Angeles, Calif.,
December 5-7, 1977. Conference Record. Volume 1.
(A79-13307 03-321 New York, Institute of Electrical
and Electronics Engineers, Inc., 1977, p. 04.2-1 to

Ionospheric conditions, including multipath, is
described. Examples of the measurements obtained in
bearing accuracy and variance under operational
conditions are presented.   76 00/00   79A41533

UTTL: Induced precipitation of inner zone electrons. 1
Observations

AUTH: A/VAMPOLA, A. L.; B/KUCK, G. A.   PAA A/(Aerospace
Corp. Space Sciences Laboratory, Los Angeles, Calif.)
B/(USAF, Space and Missile Systems Organization,
Los Angeles, Calif.)   Journal of Geophysical
Research, vol. 83, June 1, 1978, p. 2543-2551.

ABS: Narrow precipitation spikes of energetic electrons
observed in the inner zone drift loss zone during the
1968-1970 period by instrumentation on the OV1-14 and
OV1-19 satellites are shown to have characteristics
which are consistent with pitch angle scattering
produced through a resonant interaction with
propagating VLF transmissions. The angles of the pitch
angle distribution indicates that for virtually all
of the events the electrons last interacted with the
atmosphere in the vicinity of 55-62 degrees east
longitude. The L dependency of the spikes as a
function of energy is consistent with scattering by a
monochromatic wave. The evidence suggests that the
Omega or station UNB located at 11.942 E and operating
at a frequency of 11.2 kHz during this period was
responsible for these precipitation events. 78/06/01
78A41065

UTTL: Comparative statistics of atmospheric radio
and man-made radio noise in the HF band

AUTH: A/FOLTH, F. C.   PAA A/(U.S. Department of Commerce,
Office of Telecommunications, Boulder, Colo.) In:
Electromagnetic compatibility; Proceedings of the
Second Symposium and Technical Exhibition, Montreux,
Switzerland, June 28-30, 1977. (A79-36076 16-32) New
York, Institute of Electrical and Electronics
Engineers, Inc., 1977, p. 241-246.

ABS: A small but significant part of the atmospheric radio noise
data are received at 6.6 kHz, 100 kHz and 5 MHz, as well as
man-made radio noise at 250 kHz, are used. Atmospheric radio
noise levels that originate from active
thunderstorm cells. While the man-made radio noise may
be taken as representative of that from electrically
powered industrial equipment, the atmosphere
including it, both from the activity
and it. Both abstracts give an idea of the noise
environment and to aid in the design of noise
reduction circuitry for HF communications systems.
77/00/00 78A39104

UTTL: 04 2-5
The adaptive array technology discussed in the present
paper can be used to enhance HF communication
capability by providing antenna array gain, element
pattern diversity, discrimination against other user
interference and discrimination against multipath.
This is accomplished by using an additive reference
(pilot signal) to discriminate between useful and
interference signals. The achievable performance
enhancement is analyzed and adjustment ely on the basis of
experimental results obtained for a full-scale
conformal HF adaptive array.   77/00/00   79A13304

UTTL: The occupied bandwidth measuring apparatus using
a sequential spectrum analyzer

AUTH: A/ICHINO, Y.; B/KUBOTA, F.; C/SHICBARA, W.   Radio
Research Laboratories, Review, no. 24, Mar. 1978, p.
205-214. In Japanese, with abstract in English.

ABS: The paper describes a bandwidth measuring apparatus
that includes a sequential analyzer. The device can be
used for both HF and SHF. It prints out the occupied
bandwidth and calculates the envelope curve of the
spectra and the statistical distribution.   78/03/00
79A11239

UTTL: High frequency drift waves with wavelengths
below the ion gyroradius in equatorial Spread F

AUTH: A/HUBA, J. D.; B/CHATURVEDI, P. K.; C/OSSAKOW, S. L.;
D/TOWLE, D. M.   PAA A/(Science Applications,
Inc., McLean, Va.); B/(Navy and University, College
Park, Md.); C/(U.S. Navy, Naval Research Laboratory,
Washington, D.C.); D/(MIT, Lexington, Mass.)
Geophysical Research Letters, vol. 5, Aug. 1978, p.
695-698. Research supported by the U.S. Defense
Nuclear Agency and U.S. Navy.   78/08/00   78A48037

UTTL: Technique for assessing the bearing accuracy of
distant HF transmitters

AUTH: A/JONES, T. B.; B/SPRACKLEN, C. T.   PAA
A/(Leicester, University of Leicester, England)
B/(Institution of Electrical Engineers, Proceedings, vol.
125, June 1978, p. 405-413.

ABS: Travelling irregularities in the ionosphere can
produce large errors in the radio-fixed bearings of
distant HF transmitters, By analysing the small
scale frequency shifts produced in the received
signal during these periods, the measuring errors
can be recognized. A differential doppler technique is
described that can assess the accuracy of the bearings
measured for any transmitter of unknown location. The
performance of the system for a wide range of

AUTH: U:TL Aircraft Satellite Data Link
A/MFFP. B. ; B/ZIMMER. M. ; C-DUBHARUT. J.;
D/MCFADDEN. J. PAA: A/NOAA. National Weather
Service. Silver Spring. Md.). B/(NOAA. National
Hurricane Center. Coral Gables. Fla.); D/(NOAA.
Research Facilities Center. Miami. Fla.) American
Meteorological Society. Bulletin. Mar. 1978. p.
284-290.

ABS: Automated communications via the Geostationary
Operational Environmental Satellite have been employed
to circumvent the interference problems associated
with high frequency radio transmission of weather
reconnaissance data from aircraft. The
Aircraft-Satellite Data Link used for the first time
in August, 1977 to monitor a developing hurricane.
consists of a flat-plate antenna, a UHF transmitter
and a computer/keyboard assembly. The computer permits
message lengths of up to 500 characters. with
transmission at intervals as low as once per minute.
After transmission to the satellite. the messages are
retransmitted to the satellite ground station and then
relayed via a landline to the national hurricane Center.
78/03.00 78A34971

AUTH: UTTL Predicting the compatibility of high frequency
sky wave communication systems
A/SAILORS. D. B.; B/KUGEL. C. P.; C/HAYDON. G. W.
PAA: b/(U.S. Naval Ocean Systems Center. San Diego.
Calif.); C/(NOAA. Institute for Telecommunication
Sciences. Boulder. Colo.) IEEE Transactions on
Electromagnetic Compatibility. v. 19. EMC-19. Aug. 1977.
pt 2. p. 332-343.

ABS: A technique for computing the electromagnetic
compatibility of HF sky wave communication systems is
described the circuit-reliability methodology.
normally used to estimate expected circuit performance
in the absence of interference has been extended to
estimate the expected performance in the presence of
sky wave interference. A set of the compatibility
during co-channel operation called circuit
compatibility. is introduced Circuit compatibility is
the ratio of the estimated circuit reliability
considering the operation of an interfering
transmitter to estimated circuit reliability under
interference free conditions. The techniques involved
to estimate circuit compatibility include the
determination of (1) the ability of simultaneous
transmission by the desired and interfering
transmitters. (2) the joint probability that the
standard MUFs on the desired and interfering paths are
above their operating frequencies. and (3) the joint
probability that the signal-power-to-noise density
ratio is above the receiver's required

signal-power-to-noise density ratio for the desired
grade of service and that the signal-to-interference
is above the signal-to-interference threshold for the
desired emission against the interfering emission.
77/08/00 78A24808

AUTH: UTTL Military communications today and survey
A/MISRA, R. L.; B/HOWE, L. B. PAA: B./RCA.
Government Systems Div. Canden. N.J.) RCA Engineer.
vol. 22. Oct.-Nov. 1977. p. 46-51.

ABS: Attention is given to the Defense Communications
System. which carries common user traffic for commands
and controls in the U.S. Europe and the Pacific. as
well as to some global. Tactical Instrumentation
communications and   duties. Tactical transmission on
Distribution System. The trends are all   d)
satellite links is noted digital circuit   the
advantages of being compatible with large   e
integration technology. providing high message quality
for reconstitution and relay over long distances. and
allowing use of advanced cryptographic techniques.
Portable transceivers with a frequency hopping
anti-jam mode are also considered. 77 11/00
78A21324

AUTH: UTTL Effect of geophysical disturbances on
ionospheric radio wave propagation and emergency
communication
A/DUTTA, A. K.; B/MIDHA, S. K.; C/SAHA, B.; D/SEHRA.
G. S. PAA: D/(Calcutta, University. Calcutta. India)
(Institute of Electronics and Telecommunication
Engineers. Symposium on Electromagnetics. including
Antennas. Propagation and Microwaves. Calcutta.
India. Apr. 1976.) Institution of Electronics and
Telecommunication Engineers. Journal. vol. 23. July
1977. p. 440-444. Research supported by the Council
of Scientific and Industrial   .

ABS: Some trends of a study of the   of geophysical
disturbances due to solar flares   layers and
meteor showers on the to wave   add are
propagation characteristics in the entire frequency
range of radio communication ranging from VLF to HF.
are presented The results reveal that a fadeout in
the band may be accompanied by an enhancement or no
the other band   call in certain cases various
indicate the possibility of employing interband
frequency diversity systems involving VLF. LF, MF and
HF bands. to cope with the fadeouts associated with
different geophysical disturbances. 77/07/00
79A18422

TITL: Triangular enhancement of signal strength of a
high frequency radio wave in relation to travelling
ionospheric disturbances

AUTH: A/SEN, A. K. B/TRIPATHI, S. K. C/DATTA, R. N. PAA.
C/Calcutta University, Calcutta (India) Indian
Journal of Radio and Space Physics, vol. 6, June 1977.
p 146-151. 77.06.00 78A37712

TITL: Ground wave propagation characteristics of
interest to the tactical communicator

AUTH: A/PALMER, F. H. CORP: National Defence Headquarters,
Ottawa (Ontario). Dir ASOD Electromagnetic.
Prediction Predbl. in the Tactical Environ. p 16 (SEE
N82-26727 25 32)

ABS: The effects of terrain and urban cover on ground
wave propagation at frequencies of interest to the
tactical communicator are reviewed. Starting with the
simplest case, that of free space propagation,
complexities and subsequently dealing until most of the
factors which relate to propagation path loss
variability and multipath are ... are addressed. It
is demonstrated that propagation prediction and the
choice of appropriate models is as much an art as a
science. 82.04.00 82N26529

TITL: RF processing

AUTH: A/HARRIS, W. R. ...... Athens CSS
A/ ..... Engineering Centre .... Langley
Research Center Joint Univ ..... for Air
Transportation Res. 1981 p 185-192 (SEE N82-26169
17 01)

ABS: The locating radio frequency processing is discussed. A
diagram which generally illustrates the automatic gain
control is presented. In addition a differentiator for
pulse detection ... communications
.... such as wideband and ... to detect pulse
signals such as logonic. The ... enhances
reception for surveillance and observation of HF
over-the-horizon radar signals others where time
difference estimates between pulse returns are of
interest. The high resolution VLF spectrum which
receives an VLF signal ... over an HF 5581A wave
analyzer detecting signal ... in a very narrow
bandwidth of only 3 Hz is also presented. 82/06/00
82N26211

TITL: Nonlinear phenomena arising from radio wave
heating of the lower ionosphere

AUTH: A/TOMKO, A. A. CORP: Pennsylvania State Univ.,
University Park

ABS: A theoretical and experimental study of the
interaction of high power ... frequency (HF) radio
waves with the lower ionosphere is discussed. The
theoretical calculations of the electron
temperature of the ionospheric plasma can be greatly
enhanced when the plasma is irradiated by a powerful
ground based HF transmitter with an effective radiated
power of the order ... 10 kW. If this plasma heating
is maintained for ... together a few seconds, the
conductivity of the plasma can also be altered. These
temperature and conductivity ... case
significant changes in the plasma conductivity and
wave absorption in the medium. Data from the
absorption experiment clearly show a large (9 dB)
increase in wave absorption at 2.1 kHz due to a high
power (10 kW ERP) HF heating of the ionosphere. The
theorized modulation experiment generated strong
very low frequency (VLF) radiation when the ionosphere
was irradiated by a power ... modulated HF wave. These
VLF signals are believed to be due to HF heating
induced conductivity modulation of the dynamo current
system. 81/00/00 82N28820

TITL: A ... of measuring and recording the
... response of the ionosphere. (Paper)

AUTH: A/CHISM, S. CORP: Communications Research Centre,
Ottawa (Ontario). CSS Communication Technology and
Systems Branch.

ABS: A simple method by which the propagation modes of an
ionospheric communication channels can be measured and
... is described. Experimental results are
... as illustration of the technique.
85/ ... 52 12/01/00 82N1427

TITL: Anti-jam analysis of frequency hopping M-ary
frequency shift keying communication systems in high
frequency Rayleigh fading channels.

AUTH: A/VUONG, D. CORP: California Univ., Los Angeles.

ABS: The performance of such systems ... at High frequency
... by various channels, when ... used to intentional
jamming, are studied. The channels considered are
characterized by Rayleigh fading and additive Gaussian
noise. To combat jamming the communication system hops
over the total spread spectrum bandwidth, which is
many times larger than the instantaneous bandwidth
occupied by the MFSK signal to stay within the HF
band and during any given an ... correct propagation
... the spread spectrum bandwidth is typically

UTTL: Wideband HF channel analyzer
AUTH: A/PE0, D. E.; B/BA110, D. A. CORP: SRI Inc.
ABS: This report is a final report describing a wideband HF channel analyzer. The wideband HF channel analyzer is a digital signal processor designed to be used for wideband high-frequency (HF) channel measurement studies. The analyzer contains an array of digital converters, a digital preprocessor, a digital correlator array, and a post-processor and microcomputer interface. To perform a channel measurement experiment, a transmitter at a remote site broadcasts a pseudo-random, binary-phase-shift-keyed sequence modulated onto a suitable carrier. After complex demodulation at the receiver, the analyzer measures channel delay by correlating the received sequence with its internal reference. Two modes of operation are provided. The Sounder Mode is used to obtain a preliminary characterization of the channel. The Probe Mode is used to more accurately measure the channel delay characteristics.
RPT#: AD-A046224 81/01/00 81N21253

UTTL: PACE - An automatic high-frequency radio telephone system for communication in remote areas
AUTH: A/CROW, E. L.; B/HAILE, G. A. CORP: CRC (Canada) B
CORP: Communications Research Centre, Ottawa (Ontario). CSS: (Communications Systems Branch.)
ABS: A high frequency radiotelephone system offering enhanced performance compared to that of existing systems is described. The enhancements include capability of extending telephone systems without the need for co-operators. Results of preliminary, on-the-air field trials indicate that the system is reliable and exceptionally easy to use.
RPT#: CRC-1338-E 80/12/00 81N20526

UTTL: Radio direction finding on high frequency short duration signals
AUTH: A/SHEPPARD, D. O. CORP: Naval Postgraduate School, Monterey, Calif.
ABS: The feasibility of accomplishing high frequency direction finding against short duration (100-1000 ms) HF skywave signals, using narrow aperture antennas is investigated. Two statistical procedures for estimating the signal bearing are both tested and compared. These procedures employ time averaging to reduce the large instantaneous bearing error caused by the phase and amplitude distortion of the wavefront due to scattering and multipath interference. Results are presented using data collected with the Southwest

(continued)
other users of the spectrum in such an environment. Frequency Hopping FH/MFSK systems seem to be a practical than other wideband spectrum spreading techniques. A FH/MFSK system can be easily programmed to use a noncontiguous band, thereby avoiding strong interfering signals which are usually narrow band signals of friendly signals that should not be interfered with. 81/10/00 82N14441

UTTL: Ionospheric modification by high-power radio waves.
AUTH: A/DUNCAN, L. M. CORP: Los Alamos Scientific Lab., N. Mex Presented at the Ionospheric Effects Symp., Alexandria, Va., 14 Apr. 1981.
ABS: The physics of ionospheric modification by high power radio waves is reviewed. In the context of out current theoretical understanding of disturbance generation mechanisms are qualitatively described. In addition, results of recent experiments are summarized in which ionospheric irregularities are generated and their evolution and decay processes investigated in detail. The effects and potential controlled applications of these HF ionospheric modifications for various RF systems studies and uses are discussed. The C3I scientific community provides an important motivation for these ionospheric modification studies and their increased interaction and active participation in experimental design and interpretation are encouraged.
RPT#: LA-UR 81-870 CONF-810421-2 81/04/00 81N28656

UTTL: Ionospheric predictions for HF radio systems: the future
AUTH: A/BRADLEY, P. A.; B/LOCKWOOD, E. CORP: Science Research Council, Slough (England) CSS: (Rutherford and Appleton Labs.) In AGARD The Phys. Basis of the Ionosphere in the Solar-Terrest. System 13 p (SEE N81-2 5C7 14 42)
ABS: Current prediction procedures are reviewed briefly and those for which they are required are considered. Examples are presented of improvements for long term, short term, and storm prediction predictions. Desirable and likely future improvements in knowledge of ionospheric morphology leading to more accurate predictions are discussed. Specific suggestions are made for extended prediction procedures to aid optimization of communication systems. These involve estimates of additional parameters and improved representations of other factors, in particular, predictions of background interference, signal dispersion and error rates are proposed. 81/02/00 81N23537

Research Institute Coaxial Station (COP HFDF system. It is shown that for a limitation ±1% of data from this system the standard deviation of the bearing estimate for a 200 ms signal varied from 15 to 59 degrees.

RPT#: AD AD-2136   80/05 00   81N16029

AUTH: A/BUSH. C. M.; B/RUSICH. R. K.; C/MELLECKER. C.
CORP: National Telecommunications and Information Administration. Boulder. Colo.   CSS: (Inst. for Telecommunication Science.)

UTTL: HF (High-frequency) radio noise in the topside hemisphere

ABS: The results of studies, undertaken in order to develop a morphological picture of the radio environment in the ionosphere above the peak of the F2 region are discussed. The factors and assumptions that must be considered in assessing the propagation of terrestrially based signals through the ionosphere to satellite heights are summarized. Some general characteristics of the DMSP HF receiver are also given.

RPT#: PB80-162670 NTIA/REP-80-33   80/02 00   80N27591

UTTL: Theoretical feasibility of digital communication over ocean areas by high frequency radio

AUTH: A/HARDEN. G. W.; B/RUSH C. M.; C/TETERS. L. R.
CORP: Transportation System Center. Cambridge. Mass.; National Telecommunication and Information Administration. Boulder. Colo.

ABS: The theoretical reliability of digital data transmission via high frequency radio is examined for typical air traffic routes in the Atlantic and Pacific areas to assist the U.S. Department of Transportation in the evaluation of a system for improving air traffic control over ocean areas. The expected performance of a reference high frequency data transmission system with a permissible error rate of one in a thousand binary error is expressed as a fraction of time that a given theoretical reliability will be equaled or exceeded. The expected performance of air-to-air HF systems is also considered, and it is concluded that these systems should work for the reference communication system out to the line-of-sight range of about 600 km for high-flying aircraft.

RPT#: AD A079424 TSC-FAA-79-26 FAA-EM-79-20   79/11/00   80N27568

AUTH:
CORP: National Bureau of Standards. Boulder. Colo.

ABS: Detailed descriptions are given of the time and frequency dissemination services of the National Bureau of Standards. These services include the broadcasts from radio stations WWV, WWVB, and... a time and frequency... telephone service... Also indicated... 1979 are indicated. A list of time and frequency publications available about the Time and Frequency Division is also included.

RPT#: PB80-147747 NBS-SP-432   79/09 00   80N25494

AUTH: A/LARSEN. T. R.
CORP: Norwegian Defence Research Establishment. Kjeller.   In NASA. Marshall Space Flight Center. Solar-Terrest. Predictions Proc. Vol. 2 p 617 (SEE N80-24673 15-92)

ABS: The effects of D region ionization upon radio wave propagation ... the main ... magnetic storms, auroral disturbances, etc. ... Examples of radio wave characteristics for such conditions are given for the frequencies between the extremes of 1 - 12 000 Hz and ... possibilities for predicting the ionospheric effects.

RPT#: 79/12/00   80N24673

AUTH:
CORP:

ABS: The predictability of those ionospheric parameters relevant to ionospheric-reflected communications is ... the modelling needs for ionospheric communications are covered.

RPT#: 79/12/00   80N24694

UTTL: Methods of improving the performances of HF digital radio systems
A/WATTERSON, C. C. CORP. National Telecommunications and Information Administration, Boulder, Colo. CSS: (Inst. of Telecommunication Sciences.)
ABS: The bit-error-probability performances of HF digital radio systems with respect to channel and equipment additive, multiplicative, and nonlinear distortions were evaluated with respect to the nine system design features that affect the performances. The system design features are the fundamental pulse waveform, the keying method (ASK, PSK, AFSK, FSK, and CPK), the multiplexing method (frequency and concentric multiplexing), the type of demodulator filter (matched and nonmatched nonadaptive filters and quasistatic and dynamic adaptive filters), the detection method (coherent, partially coherent, differentially coherent, and noncoherent), the transmitter power, the antennas, diversity, and error coding. Spectral efficiency (information rate/signal bandwidth) is also evaluated. The best combination of design features is determined.
RPT#: PB80-128606 NTIA-79/29 79/10/00 80N21663

UTTL: A model of natural HF radio noise in severely disturbed propagation environments
A/PRICE, G. H. B/SMITH, G. CORP. SRI International Corp., Menlo Park, Calif.
ABS: A model is developed that describes the impact of severe propagation losses on the natural noise environment at high frequencies.
RPT#: AD-A077619 AD-E300627 DNA-4930T 79/03/31 80N20474

UTTL: New technology to improve HF circuit reliability and availability for remote regions
A/BELROSE, J. S. B/BODE, L. R. CORP. Communications Research Centre, Ottawa (Ontario). CSS: (Dept. of Communications.) In AGARD Spec. Topics in HF Propagation 11 p (SEE N80-19372 10-32)
ABS: The research and development underway at the Communications Center, Ottawa, is described along with that in Canadian industry. Topics discussed include: (1) trail-remote CW radio communications, (2) antennas for base station and land use, (3) HF-SSB transceiver, and (4) SSB tone decoder. 79/11/00 80N19417

UTTL: A mobile HF impulse source locator
A/HERON, M. L. CORP. James Cook Univ. of North Queensland, Townsville (Australia). CSS: (Physics Dept.) In AGARD Spec. Topics in HF Propagation 9 p (SEE N80-19372 10-32)
ABS: A noise pulse locator designed to meet the needs of research into the structure of tropical cyclone (hurricane) systems but which can be easily adapted to storm location and tracking applications is described. The principle and instrumentation of the noise pulse locator are discussed. The resolution, errors, and limitation of the equipment as employed on an Orion aircraft are estimated and the status of application and adaptation of the instrument are noted. 79/11/00 80N19413

UTTL: Principles of HF communication in tunnels using open transmission lines and leaky cables
A/WAIT, J. R. B/HILL, D. A. CORP. Colorado Univ., Boulder. (Cooperative Inst. for Research in Environmental Sciences.) In AGARD Spec. Topics in HF Propagation 4 p (SEE N80-19372 10-32)
ABS: The concept that electromagnetic waves can be guided by tunnels is covered. It is shown that the frequency must be greater than some cut-off value that typically is of the order of 50 MHz when the tunnel contains axial conductors. Its situation is changed dramatically, and there is no cut-off as such, although wave-guide type modes propagate at the higher frequencies. 79/11/00 80N19412

UTTL: Perspective on the prediction of auroral absorption
A/ CORP. Topics in HF communication Sciences, Boulder, Colo. In AGARD Spec. Topics in HF Propagation 10 p (SEE N80-19372 10-32)
ABS: Methods for the calculation of the effects of auroral absorption on high frequency (HF) communication circuits for HF radio links are presented. The geographic distributions, the temporal variations, and the relationships with polar cap absorption on one hand, and with magnetospheric substorms on the other, are briefly examined. 79/11/00 80N19350

UTTL: The geomorphology of the ... breakthrough propagation

AUTH: A/GOODMAN, J. M. CORP Naval Research Lab., Washington, D. C. CSS: (Communication Sciences Div.) In AGARD Spec. Topics in HF Propagation 13 p (SEE N80-19372 10-32)

ABS: The morphological behavior of HF ... waves that propagate through the ... ... and observed at satellite levels is described, using realistic models ... the types of information about the tropospheric structure that could be deduced from satellite measurements of ground-based HF signals of HF noise sources are studied. 79/11/00 80N19385

UTTL: The influence of ionospheric models on calculations of decametric wave propagation

AUTH: A/FLEURY, R.; B/GOURVEZ, P. CORP Centre National d'Etudes des Telecommunications, Lannion (France). In AGARD Spec. Topics in HF Propagation 14 p (SEE N80-19372 10-32)

ABS: The ionospheric F2 layer critical plasma frequency foF2 is one key geophysical parameter affecting those involved in HF wave propagation. This means that the efficiency of numerical methods available for HF waves propagation simulation (e.g. ray tracing) is greatly limited by the uncertainties inherent to the foF2 input models. Observed planetary distributions of monthly median foF2 and those calculated by current models which give values of order bars on foF2 are compared systematically. Further on accurate ray tracing method used with F2-region models in which foF2 is perturbed in amounts equal to typical values of its error bars helps to assess the influence of these error bars on HF waves propagation results. 79/11/00 80N19383

UTTL: Hybrid ray-mode formulation of tropospheric propagation

AUTH: A/CHO, S. H.; B/MIGLIORA, C. G.; C/FELSEN, L. B. CORP Polytechnic Inst. of New York, Farmingdale. CSS: (Dept. of Electrical Engineering.) In AGARD Spec. Topics in HF Propagation 15 p (SEE N80-19372 10-32)

ABS: A method of analyzing radio propagation using a combination of ray and mode is proposed, using as an illustration the propagation of radio waves in an elevated tropospheric duct. Its emphasis is given to how the hybrid ray-mode approach could be readily applied to a number of different propagation related problems across the frequency spectrum. 79/11/00 80N19382

UTTL: Tropospheric effects on HF propagation

AUTH: A/GIBSON, ... H. CORP ... Co. Syracuse, N.Y. In AGARD Spec. Topics in HF Propagation 14 p (SEE N80-19372 10-32)

ABS: An evaluation is made of the effect of tropospheric refractive bending on the propagation of HF radio waves. The troposphere is ... the conductivity ... refraction in the troposphere ... transverse ordinary mode of propagation, the electron density being represented by a Chapman model. Ray tracings are performed based on the assumption that the propagation medium is stratified into layers of refractive index. The elevation and distance and the true and virtual reflection heights are calculated as a function of surface conductivity and various electron density profiles. The relationship between radar range and target ground distance is also examined. The existence of long range propagation paths beyond 4500 km for specific tropospheric and ionospheric conditions is discussed. 79/11/00 80N19380

UTTL: Augmentation of HF propagation

AUTH: A/BARRETT, T. J. CORP Mission Research Corp., Santa Barbara, Calif. In AGARD Spec. Topics in HF Propagation 13 p (SEE N80-19372 10-32)

ABS: ... that chemical release might be employed to alter the application channels for tactical critical periods in which tropospheric communication to be used in the event of an emergency ... an accurate emergency. The technique, while important, does provide communication covering over short areas unfortunately 3000 mile range in a nuclear environment, the chemical release would alter the use of ... nuclear blackout ... to provide information to do so to blackout caused by a particle ionized natural gas choice of optimum cloud location. This technique does not replace any current communication system but does provide a way to maintain communications for a few minutes, which might otherwise be blocked out for hours. Potential applications of this technique include communication with radio fleets, subforces, SAC, and other tactical communications. 79/11/00 80N19379

UTTL: Comparison of measured and predicted MUF's at a remote location

AUTH: A/BARTHOLDT, T.   CORP: Forsungsinstitut der Deutschen Bundespost, Darmstadt (West Germany).   In AGARD Spec. Topics in HF Propagation 6 p (SEE N80-19372 10-32)

ABS: The times at which the signals are received first and last by ionospheric propagation during the diurnal cycle during the time where the frequency of the transmitted signal is equal to the 10th MUF (maximum usable frequency). Values of these transition times are taken from the field strength records of a path from England to Germany and then compared with the predicted values. A further comparison is made between the observed MUF transition times and the ionogram critical frequency at the appropriate path mid-point, i.e. both in real and for the above mentioned path. Good agreement is obtained between the observed critical frequencies and the measured MUF's. It is suggested that information obtained from signal-strength records can lead to substantial improvements in MUF predictions. This is of considerable importance since operational requirements usually exist for real-time situations and not for the average conditions computed by prediction programmes.
79/11/00   PGN19378

UTTL: Assessment of HF communications reliability

AUTH: A/MASLIN, N. M.   CORP: Royal Aircraft Establishment, Farnborough (England).   CSS: (Radio and Navigation Dept.)   In AGARD Spec. Topics in HF Propagation 12 p (SEE N80-19372 10-32)

ABS: The concept of circuit reliability for an HF sky-wave link is discussed. Its frequency dependence is considered together with the relationship to the median received signal-to-noise ratio for a given HF circuit. The importance of a study of this kind is the ability to be able to quantify changes that could be made for an HF circuit, in terms of reliability improvements, and hence to make decisions of the cost effectiveness of increasing transmitter power, antenna efficiencies, directional antenna design.   79/11/00   80N19377

UTTL: Real-time adaptive HF frequency management

AUTH: A/FENWICK, R. B.   B/WOODWARD, F. J.   CORP: Barry Research Corp. Sunnyvale, Calif.   In AGARD Spec. Topics in HF Propagation 14 p (SEE N80-19372 10-32)

ABS: In 1972 the United States Air Force began a program aimed at improving tactical HF communications. The approach involved measurement in real time of the important unknowns: propagation, noise, and spectrum

occupancy, and adapting circuit use at a frequency, in real-time, to the conditions imposed. A potential problem with this approach is causing harmful interference to other spectrum users. An exercise called TROPHY DASH III was conducted to assess the ability of a particular technique to permit real-time selection of frequencies while yielding acceptably low interference. TROPHY DASH III results are viewed as significant. First, in spite of extensive notification to other spectrum users, negligible interference was reported. Based upon these results, real-time adaptive sharing of the HF spectrum in the tactical theater appeared both valid and feasible for a limited number of high priority users, and procurement of the AN/TRQ-35(V) Tactical Frequency Management System was authorized.   79/11/00   80N19379

UTTL: Modern HF communication for low-flying aircraft

AUTH: A/BRUNE, J. F.   B/RICCIARDI, B. V.   CORP: Army Avionics Research and Development Activity, Fort Monmouth, N. J.   CSS: (Communication and Sensor Div.)   In AGARD Spec. Topics in HF Propagation 15 p (SEE N80-19372 10-32)

ABS: An application of the HF propagation phenomenon that can provide relatively short range, reliable, terrain independent communications is described. It has been shown that communications out to ranges of 50 km under varying terrain conditions, to and from low flying aircraft, is an extremely difficult problem especially for air mobile tactical forces. The US Army uses the top-of-the Earth communications (NOE Com-) system program conducted extensive tests and analysis using the HF media for tactical communications with low flying aircraft. The use of the near-vertical-incidence skywave (NVIS) portion of the HF channel for aircraft communications is described. The HF-NB radio systems have the capability of operating in either a ground wave or NVIS mode. For the NVIS mode, the energy is directed vertically to the ionosphere and returned to the surface of the Earth. The NVIS mode provides umbrella type coverage. Because of NVIS propagation, HF/NB systems with appropriate antennas have the capability of providing communications coverage out to ranges greater than 50 km in any type of terrain. The NVIS mode is terrain independent. The characteristics of the HF NVIS mode are considered and the features required of a modern HF radio system to make efficient and practical utilization of the already overcrowded HF band are noted.   79/11/00   80N19390

AUTH: A/NASLIN, L. M. CORP: Royal Aircraft Establishment.
Farnborough (England). CSS: (radio and Navigation
Dept.) In AGARD Spec. Topics in HF propagation 13
p (SEE N80-19372 10-32)

ABS: HF (2-30 MHz) radio communication is a principal means
of beyond line of sight communication to aircraft.
There are particularly serious problems for the small
aircraft. Many factors degrade the overall received
signal-to-noise ratio both at the ground and in the
aircraft. To achieve satisfactory results, careful
consideration should be given not only to the terminal
equipment but also to the long term geographic
planning and to the propagation of the frequencies to
be used over robust HF links. The worst HF
communication problems occur for a short range
air-ground sky-wave link at night which requires
frequencies at the low end of the HF band. Working
over a longer range link increasing the 'optimum
working frequency', thus avoiding the poor antenna
efficiencies and generally high external noise
levels. It is shown that good frequency management,
good antenna directivity and the use of a number of
geographically separated remote receiving stations are
vital in providing satisfactory communications
reliability to the small aircraft. 79/11/00
80N19374

AUTH: A/BURGESS, B. CORP: Royal Aircraft Establishment.
Farnborough (England). In AGARD Spec. Topics in HF
Propagation 6 p (SEE N80-19372 10-32)

ABS: High frequency radio is a prime means for beyond
line-of-sight air-ground communications and will
remain so such into the foreseeable future.
representing Satellite communications where this
latter system is employed various aspects that make
up an HF communications system and the contribution
that they make to the overall system performance are
reviewed. The needs of the modern user in the use of
HF communications are considered with emphasis on the
changing requirements on the HF link that that the use of
digital communications brings. Error rate performance,
avoidance of interference, HF prediction techniques
for northern latitudes, and the role of channel
evaluation techniques are addressed in order to
acquaint the research worker with the status and
trends in airborne HF communications and to indicate
where further work may profitably be undertaken to
eventually improve system performance. 79/11/00
80N19373

AUTH: A/COYNE, V. J. P2A A/LEDOC Griffiss AFB, N.Y.)
CORP: Advisory Group for Aerospace Research and
Development. Neuilly-Sur-Seine (France) Sym. held
in Lisbon, 28 Sep. - 1 Oct. 1972

RPT: AGARD-CP-263 ISBN-92-835-0273-3 AD-AD65/2 79/11/06
80N19372

AUTH: A/MAEDA, R CORP: Radio Research Labs, Tokyo
(Japan). In NOAA Solar Terrestrial Predictions Proc.
Vol 1 p 212-222 (SEE N80-18480 09-92)

ABS: Radio wave prediction services in Japan
communications are outlined with respect to prediction
methods, performance and use of the current
prediction is based on the CCIR atlas combined with
some modifications matching to a computer system. The
principal service is the monthly medium prediction
issued regularly three months in advance for various
circuits in operation. In addition for short
term and disturbed communications and real time
prophetic sounding data are used on theory and
prediction is expected to be integrated in the future
prediction service 79/06 00 80N18480

AUTH: A/AKASOFU, S. I. C. NAKAMURA, T.
CORP: Solar Terrestrial Predictions Proc.
(Japan). In NOAA Solar Terrestrial Predictions Proc.
Vol 1 p 112-201 (SEE N80-18480 09-92)

ABS: The Hiranch Branch of RR predictions techniques are
examined separately for the 27 day recurrent storm
events and prediction for the first time. Civilities in
predictions on the correct so far are the accuracy of
the storm predictions during the term are evaluated.
In addition to the storm prediction on short-term
predictions of HF radio propagation conditions are
conducted at Hiraiso. The HF propagation predictions
are techniques concerned in magnetic storm prediction
applications of the magnetic storm prediction
79/18/00 80N18471

AUTH: A/KOS, Z. B/STASIEWICZ, K. CORP: Polish Academy
of Sciences, Warsaw. CSS: (Space Research Centre.)
In NOAA Solar Terrestrial Predictions Proc. Vol 1 p
07-71 (SEE N80-18480 09-92)

ABS: A provisional method for determining the hours and
frequencies at which solar flares cause major
radiopath absorption increase is presented. This
technique is incorporated in a computer program for
monthly radio propagation prediction 79/06/00
80N18468

AUTH: A/DEHROLDT, T. CORP: Forschungsinstitut der
Deutschen Bundespost, Darmstadt (West Germany). In
NOAA Solar-Terrest. Predictions Proc., Vol. 1 p
25-43 (SEE N80-18452 09-42)

ABS: Long term and short term HF propagation prediction
service is described. The long term prediction method
is computer based. The short term prediction method is
still a manual method which relies mainly on a number
of continuous field strength recordings and on the
forecaster's ability to interpret the recordings and
to combine this information with all available
solar geophysical data. Examples of the predictions
are given and are compared with actually observed HF
propagation conditions. 79/08/00 80N18465

AUTH: A/SCHEIBE, M. CORP: Mission Research Corp., Santa
Barbara, Calif.

ABS: The relative importance of various reaction rates are
examined using the results of a recent MELT code
calculation for a megaton range burst at 250 km
altitude. New rate coefficients are developed for the
critical reactions involved in the ion decay and the
uncertainties in these rate coefficients are examined.
With the use of a number of calculations, some of them
rather crude, these rate coefficient uncertainties are
translated into uncertainties in communications
degradation effects for HF and higher frequencies and
some of these are significant. Areas in which
improvement is needed, with regard to both reaction
rate data and the chemistry treatment in MELT, are
noted.
RPT#: AD-A078924 AD-E300633 MRC-R-458 DNA-4971F 79/05/31
80N18279

UTL: Radio communication utilizing the base of a
striated barium plasma
AUTH: A/FUCKS, G. J.; B/SCOTT, L. D. C/SCALE, D. H.
D/NORMAN, W. R. CORP: Mission Research Corp., Santa
Barbara, Calif.

ABS: In conjunction with the DNA barium releases, average
and short term HF propagation...
80N11299
RPT#: AD-A072167 AD-E300455 MRC-R-401-R DNA-4670T 78/07/00

AUTH: A/SOUTH, F. T. CORP: Rice Univ., Houston, Tex.
ABS: A 430 MHz backscatter radar is used to diagnose the
...
80N34024

UTTL: Propagation at medium and high frequencies. 1:
Practical radio systems and travelling media
AUTH: A/BRADLEY, P. A.   CORP: Appleton Lab., Slough
(England).   In AGARD Aerospace Propagation Media
Modelling and Prediction Schemes for Mod. Commun.,
Navigation, and Surveillance Systems 21 p (SEE
N79-27385 18-32)
ABS: The principal phenomena associated with ground wave
and sky-wave propagation at HF and HF are discussed.
Particular consideration is given to transmission loss
and coverage range, the dependence on
ground-reflection properties, the state of the
ionosphere and the earth's magnetic field. Regular
temporal changes depending on time-of-day, season and
solar activity, together with short-term fading are
noted. The importance of antenna design in system
operation is emphasized. Practical usage of the MF and
HF bands is illustrated. Propagation parameters
susceptible to modelling were reviewed, followed by an
examination of requirements for long-term propagation
models for system design, frequency allocation, and
assignment. Short-term models are shown to be of
potential value for frequency management, but the
logistic difficulties of producing and disseminating
results from these models are emphasized.   79/05/00
79-27386

UTTL: Aerospace propagation media modelling and
prediction schemes for modern communications,
navigation, and surveillance systems.   CORP: Advisory
Group for Aerospace Research and Development.
Neuilly-Sur-Seine (France).   AGARD Lecture Ser., held
at London, 4-5 Jun. 1979, Boulder, Colo., 14-15 Jun.
1979
RPT#: AGARD-LS-99 ISBN-92-835-1322-3 AD-AG71130   79/05/00
79-27385

UTTL: Marconi aerial output M-33-5500-01 evaluation:
Comparison of this active antenna with two previously
evaluated active antenna
AUTH: A/THOMLESS, E. A.   CORP: Naval Ocean Systems Center,
San Diego, Calif.
ABS: The Marconi M-33-5500-01 antenna has significantly
better reception of night-time signals than the Baystore
or BW antenna. But as a rule, active antennas
self-generated interference at ... interference is
... where several nearby ... emissions occu-
simultaneously. Despite this ... this
antenna is judged to be well engineered and properly
designed and optimized for integration into HF
receiving systems. Because of its small size, it could
serve as a backup receiver antenna for HF and lower

frequencies aboard navy ships. Relationships a
presented for apparent effective height antenna noise
power on output, antenna noise figure, antenna gain,
operating noise figure of a complete receiving system,
and intermodulation intercept points, ratios, and
product levels.
RPT#: AD-A066449 NOSC/TR-361   78 10 25   79-25279

UTTL: On approaches to robust detection for HF
communications
AUTH: A/MORRIS, J. M.   CORP: Naval Research Lab.,
Washington, D. C.
ABS: This report proposes robust detection theoretic
approaches to HF signal receivers ... of these
approaches require a uniformly ... most detection
theory results to the specifics ... the HF
communication problem. One approach recommended is
however, it demands more original effort. The report
contains a survey of recent robust detection theory
research and discusses the applicability of the
research results to HF additive interference models;
it concentrates on the known empirical case to highlight
the additive interference problem.
RPT#: AD-A065030 AD-E000/70 NRL-8349   79 01 25
79-23329

UTTL: Exact ray paths in a multisegment quasiparabolic
ionosphere. Equations for HF ray paths and in
displaying propagation conditions for real-time
minicomputer based assessment systems.
AUTH: A/HILL, J. R.   CORP: Naval Ocean Systems Center, San
Diego, Calif.
ABS: Exact ray path calculations in a multiple
quasiparabolic ray ionosphere are derived for use in
radio ... minicomputer based propagation assessment
system. Closed ... path forms are derived by
quasiparabolic ... of the ionospheric layers
having electron density profiles and valid to layers
having a minimum. Ray forms are derived ...
demonstrating implementation of the equations on a
minicomputer (SEL 810A) and a microcomputer (Tektronix
4051). The microcomputer program is BASIC) used to
... Guide, the ray program listings is included.
RPT#: AD-A063904 NOSC/TR-266   79/01/15   79-22687

AUTH: A/RUSH, C. M. CORP: Institute for Telecommunication
Sciences, Boulder, Colo. In AGARD Operational
Modelling of the Aerospace Propagation Environ.. Vol.
1 and 2 9 p (SEE N79-18094 09-31)

ABS: The fundamentals employed to produce the ionosphere and
radio propagation conditions used in short-term and
long-term analysis are described. These techniques range
from the highly statistical approach such as given in
CCIR REPORT 252-2 for estimating skywave field
strength for radio waves at frequencies greater than
2.0 MHz to the application of basic physical
principles to forecast the ionospheric response to a
solar disturbance. The approach taken at both
short-term and long-term levels for prediction
techniques, with observations of various
ionospheric/radio propagation parameters is described
and methods for further improvement of the predictions
are addressed. 78/11/00 79N18113

AUTH: A/LAUCHE, H. B/LANGE-HESSE, G. CORP:
Max-Planck-Institut fuer Aeronomie, Katterburg Lindau
(West Germany). In AGARD Operational Modelling of
the Aerospace Propagation Environ.. Vol. 1 and 2 5 p
(SEE N79-18094 09-31)

UTTL: Variation of the green line oxygen airglow
emission rate as a precursor indicative of wintertime
absorption anomaly of HF radio waves.

ABS: A radiometric sion of the E layer obtained from the
ground at nighttime during the winter anomaly campaign
at Aeronomic O. Spain, winter 1975-76 is comparison of
those optical data with the HF waves ionospheric
absorption values of the following day indicates that
a special kind of variation of the airglow emission
can be used as a precursor indicative for an increase
of the ionospheric radio absorption on the
following day. 78/11/00 79N18108

AUTH: A/GOUIELARD, C. B/CARATORI, J. CORP: Laboratoire
d'étude des transmissions ionospheriques, Cachan
(France). In AGARD Operational Modelling of the
Aerospace Propagation Environ.. Vol. 1 and 2 24 p
(SEE N79-18094 09-31)

UTTL: Modeling the atmosphere in problems concerning
the management of HF transmission networks.

ABS: The atmospheric parameters are determined by studying the
relation between any two parts of a several thousand
kilometer zone centered on a station. The method
developed is based on the study of ionospheric models
and takes into account the horizontal ionization
gradients. The model chosen is discussed and compared
with other models. Its use in measuring transmission

AUTH: A/KAISER, J. CORP: Royal Aircraft Establishment,
Farnborough (England) Transl. into ENGLISH from
Rundfunk-Tech. Mitt. (West Ger.). v. 19, no. 1, 1975 p
23-50

UTTL: Near field problems with high power transmitters

ABS: Unwanted interactions between the electromagnetic
field radiated from a transmitter and the environment
are discussed for the MW frequency range. Dangers
occurring in the near field of the transmitter
include: the direct radiation effect of radiation on
the human body; unacceptable interference voltages
produced in electrically conductive structures in the
transmitter's radiation field; detonation of explosive
substances and mixtures; and interference to
electrical and electronic equipment. The very presence
of structures can also disturb the transmitter's
radiation pattern. Methods for protecting property
adjacent to a transmitter are included.

RPT#: RAE-LIB-TRANS-1982 BR65966 70/07/00 79N21247

AUTH: A/DERBOLDT, T. CORP: Forschungsinstitut der
Deutschen Bundespost, Darmstadt (West Germany). In
AGARD Operational Modelling of the Aerospace
Propagation Environ.. Vol. 1 and 2 12 p (SEE
N79-18094 09-31)

UTTL: HF short-term field-strength predictions and
their agreement with observations

ABS: Daily HF field-strength predictions for two European
and three overseas paths based on solar-geophysical
data, real time field strength measurements of distant
HF-transmitters on 26 frequencies, the geomagnetic
variations indicated on a local magnetometer in real
time, the critical frequencies of the ionosonde at
Lindau, and solar observations from Wendelstein and
geomagnetic observations from Witout are discussed.
The distinctive feature of the predictions is that
they are forecast in the form of a quality figure
which is related to the daily expected field strength.
All paths are monitored on four or five frequencies
and the daily quality figure for each path is obtained
by calculating the mean of the 24 hourly values of the
field strength of that frequency having the highest
field strength for each or the individual hours. This
is done on the assumption that the optimum frequency
would be used for communication. This value does not
agree with the mean of the field strength of the five
frequencies. The quality figure in steps of 0.1
(corresponding to 0.5 dB) is forecast for the
above mentioned 5 circuits on each workday before
noon. Reliability of the forecasts is considered.
78/11/00 79N18112

AUTH: 4/BARCLAY, L. W.   CORP: Directorate of Radio
Technology, London (England).   In AGARD Operational
Modelling of the Aerospace Propagation Environ.. Vol.
1 and 2 9 p (SEE N79-18094 09-31)

ABS: Prediction of the monthly median received
signal/external noise ratio as a function of frequency
and of the system characteristics is considered along
with the prediction of the probability that a
particular grade of service will be achieved. Specific
topics covered include: the upper frequency limit;
signal to noise ratio; circuit reliability; time
availability; service probability; long term
predictability; and multimode propagation. The value of
probability prediction of circuit performance to the
user is discussed.   78.11.00   79C18105

AUTH: 4/BRADLEY, P. A.   CORP: Appleton Lab., Slough
(England).   In AGARD Operational Modelling of the
Aerospace Propagation Environ. Vol. 1 and 2 12 p
(SEE N79-18094 09-31)

ABS: The roles of predictions in system planning, frequency
management, and frequency assignment are discussed.
The need for a compromise between prediction
complexity and accuracy is stressed, with regard to
available resources and uncertainties in current
knowledge of the behaviour of the ionosphere. It is
concluded that no single prediction procedure can be
optimum for all purposes, in terms of
approaches based on empirical data as to allowances which
make due account of the separate physical factors known to
be of importance are considered. Features of the
latest prediction method in the CCIR are discussed and
those areas where further studies are needed or
improvements seem possible in the foreseeable future

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height'd distribution of electron concentration and
methods of determining raypaths ... the importance of
ionospheric tilts and off great-circle propagation.
Developments in the prediction of auroral absorption
and of sporadic-E attenuation and reflection losses
are described. The difficulties ... such as encountered
for predictions at frequencies close to the maximum
usable frequency are highlighted and the need for an
'above the MUF' loss allowance is reviewed. Means of
extending predictions to combine data for different
propagation modes and to give assessments of the
likelihood of multipath are also considered.
78/11/C0   79N18104

AUTH: 4/ROEHRBULLER, J.   CORP: Naval Ocean Systems
Center, San Diego, Calif.   CSS: (EM Propagation Div.)
In AGARD Operational Modelling of the Aerospace
Propagation Environ. Vol. 1 and 2 9 p (SEE N79-18094
09-31)

ABS: An environmental prediction and assessment system
(EPAS), comprised of a variety of real time sources of
solar/geophysical data and a ... which collects,
processes, and selectively disseminates these data to
regional propagation assessment terminals is
described. (Operational environmental models used in
FROPHET, a real time assessment terminal, and PROPHET
products, are discussed. Results of testing the
terminal and the EPAS concept in an operational
environment are indicated.   78.11.00   79N18097

AUTH: 4/ LINCOLN, J. ...   CORP: Air Force Global Weather
Central ...   In AGARD Operational
Modelling of the Aerospace Propagation Environ.. Vol.
1 and 2   7 p (SEE N79-18094 09-31)

ABS: ... environmental forecasting abilities of the Space
Environmental Support System (SESS) of the Air Weather
Service ... are described including (1) forecasting
and specification of ionospheric variability; (2)
forecast hazard specification of solar flare and solar
particle events; and (3) forecasting of geomagnetic and
... such events. Customer requirements for support include
forecasts with lead times ranging from hours to
months, real-time notification of solar and
geophysical events within minutes, forecasts tailored
to specific user requirements, and detailed
data analysis studies. Specification and prediction
models currently in operational use ... the worldwide

solar and geophysical observing network, and the data
handling and processing system are described. The
development and present status of operational
forecasting and skills in the areas of High Frequency
propagation, vertical electron density profiles, total
electron content, solar and geophysical indices, and
solar radiation are discussed. Future military
applications and use of space environment support are
also discussed along with associated technologically
deficient areas. 78/11/00   79N18095

AUTH: A/DAVIES, K.   CORP: National Oceanic and Atmospheric
Administration, Boulder, Colo. (SS.) iSpace
Environment Lab.) In AGARD Operational Modelling of
the Aerospace Propagation Environ Vol. 1 and 2 19 p
(SEE N79-18064 09-31)

ABS: Long term ionospheric predictions required for the
design of radio circuits, as motion of radio
frequencies, power requirements, and antenna design
and short term ionospheric predictions (forecasts)
intended for use of station operators are discussed.
It is shown that temporal and geographical systems are
adequate for the long term prediction of frequency
allocation on an international basis, and for the
design of many long distance communications circuits
on such frequencies, as well as broad planning of
navigation systems on very low frequencies and the
design of broadcasting service on medium frequencies.
Short term predictions are shown to be inadequate
because forecast service centers can often warn military
communicators and use of warning in order to provide for
alternative means of communication to initiate or
delay rescue operations, and to inform users of the
cause of their transmission problems. 78/11/00
79N18095

AUTH: A/SOTCHEK, H.   PAA. Advisory Group for Aerospace
Command)   CORP: Advisory Group for Aerospace Research
and Development, Neuilly-Sur-Seine (France).   Meeting
held at Ottawa 24-28 Apr. 1978

TITL: Operational Modelling of the Aerospace
Propagation Environment Vol. 1 and 2

RPT#: AGARD-CP-238-VOL-1 AGARD CP-238-VOL-2
ISBN-92-835-0224-8 AG-AGARD-CP ..B, 11/00   79N18064

TITL: Spectral characteristics of earth-space paths at
2 and 30 GHz
AUTH: A/BAXTER, R. A.; B/HODGE, D. B.   CORP: Ohio State
Univ., Columbus.   (SS.) (Electro...ence Lab.)

ABS: Spectral characteristics of 2 and 30 GHz signals
received from the Applications Technology Satellite 6
(ATS-6) are analyzed in detail in elevation angles
ranging from 0 deg to 44 deg. The spectra of the
received signals are characterized by slopes and peak
frequencies. Statistics of the spectrum widths are
presented as probability density as function on angle is
investigated. The 2 and 30 GHz spectral shapes are
contrasted through the use of scatter diagrams. The
results are compared with theoretical results from
fluctuation theory. The average signal slopes are in
close agreement with theory. A close elevation angle is
from the average value at any given elevation angle is
quite large.
RPT#: NASA-CR-159417 ESL-784249-7 7N 08/00 79N17073

TITL: Comparison between the plasma line spectra in
the F and sporadic E regions
AUTH: A/ABB. GR. 1   CORP: ... ... Sao Paulo (Brazil)
(Faculty). Sao Jose dos Campos (Brazil)

ABS: Comparisons of the spectral lines observed in the F
and E regions of the ionosphere when excited by intense
hf radio waves are presented.
RPT#: INPE-1299-PE/143   7N/06/00   79N16481

TITL: ... ... simplified computer method for long-term
calculation of HF sky-wave ...
AUTH: A/PICCHIARA, A. A.   CORP: ... ... U.S. R.   FAA
A/Institution Technologico de Aeronautica, Sao Jose
dos Campos, Brazil)   CORP: Instituto de Pesquisas
Espaciais, Sao Jose dos Campos (Brazil)

ABS: A ... ... computer method is designed to supply the
necessary parameters to design an HF radio
predictions. The plan of the method is to
distribute long-term median values into numerical
values to allow the design of reliable frequencies,
transmitter power and antenna times. The possible
influences of the parameters used are examined.
RPT#: ... 1377 IN/175   7N 10/00   79N14279

TITL: Long-term frequency predictions 2GHz to 2GHz
... ...
AUTH: A/BEALS, E. D.   CORP: Air Force Global Weather
Central Offutt AFB, Nebraska

ABS: High frequency (3-30 MHz) radio waves used for long
range communications networks are disturbed by mercy of

natural variations in the ground upper atmosphere,
more specifically, the ionosphere. This guide explains
these variations and describes various products
available from the Air Force Global Weather Central
(AFGWC) which can [...] to [...] operator of
[...] to [...] predictive soft [...] of [...] physical activity
which can affect HF circuit reliability. The guide is
basic and was designed with the field communicator in
mind, but contains information of value to circuit
managers and supervisors.

RPT#: AD-A057126 AFGWC-TM-78-001   79/01/00   79N10673

AUTH: A/KAZTOR I J   CORP: Inst. Tec. de Pesquisas
      Espaciais. Sao Jose dos Campos (Brazil).

UTTL: Spectrum of the excited plasma line in sporadic-E

ABS: A powerful HF radio wave was used to heat the local
     plasma at the ionospheric height of wave reflection.
     During a 1976 heating experiment, the plasma line
     spectrum was observed at the sporadic-E. The spectrum
     and the conditions of the experiment are presented.
     The growing mode instability is dominant in the
     sporadic-E plasma lines.

PPT#: INPE-1259-PE/141   79/06/00   78N28720

AUTH: A/SCALE. D H   CORP: Mission Research Corp., Santa
      Barbara. Calif.

UTTL: An ambient HF radio [...]

ABS: An approximate method for the calculation of high
     frequency sky/wave radio propagation geometry is
     developed. The method is inexpensive in comparison to
     ray trace techniques yet capable of accounting for
     instantaneous local transition to conditions including
     tilting and off great circle propagation. It is thus
     appropriate for use in absorption of system response
     to nuclear effects on HF propagation, including
     alterations of mode geometry. Errors inherent in the
     approximations are shown to be small compared to
     errors caused by inaccuracies in ambient or disturbed
     ionospheric models.

RPT#: AD-A062576 AD-E300150 MRC-R-340 DNA-4420T   77/09/00
      78N24.30

UTTL: High frequency radiowave propagation in the
      ionosphere

AUTH: A/JONES, T. B.   CORP: Leicester Univ. (England).
      CSS. (Dept. of Physics.)   In AGARD Recent Advan. in
      Radio and Opt. Propagation for Radi Commun.
      Navigation and Detection System. 16 p (SEE N78-23316
      14-32)

ABS: The effect of the ionosphere on the performance of a

---

high frequency radio communication systems. Reflection
at oblique incidence, multipath propagation, and
maximum and lowest usable frequencies are discussed
along with [...] methods of
[...] are predicted [...]
tracing a number of [...] tracing the
complete channel performance [...] system
performance are discussed.   78 [...]   78N23323

AUTH: A/RUBIO, P.   CORP: Army Electronics Command, White
      Sands Missile Range, N. Mex.   CSS. (Atmospheric
      Sciences Lab.)

UTTL: Ionospheric D-region electron density data
availability

ABS: A survey of the availability of ionospheric D-region
     electron density data has been collected and is
     reported here. A table which lists worldwide locations
     where D-region ionospheric electron density profiles have
     been obtained is included. [...]
     electron density data [...]
     [...] of global distribution and [...] of both
     intermediate and long-path high frequency (3 to 30
     MHz) Army communications channels. Worldwide maps are also
     included which serve to illustrate localities where
     D-region data have been recorded and worldwide sections
     and paths where high frequency communications channels
     are currently active.

RPT#: AD-A045275 ECOM-DR-77-6   77/12/00   78N14698

AUTH: A/TACKER, B.   CORP: Stanford Research Inst., Calif.

UTTL: Ground and characteristics of ionospherically
     propagated HF atmospherics from normal and severe
     thunderstorms.

ABS: A class of atmospherics of extremely short duration
     (pulse width of 0.5 [...] associated with
     thunderstorms classified as severe and thus likely to
     produce tornadoes, were studied in California. The
     antenna was scanned across regions within a one-hop
     propagation distance likely to develop tornadoes at
     distances greater than [...] km [...]
     [...] the [...] detection of [...]
     [...] a class of backscatter [...] analysis
     of the data recorded in these detections [...] periods
     for which severe storm events were correlated led to the
     discovery of a class of radio atmospherics
     [...] different from the normal atmospherics during
     thunderstorms activity. [...] analysis of this new
     type take the form of narrow [...] series impulses
     having certain distinctive characteristics.   77/00/00

78N14226

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Lecture Series 127 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 the lectures is to survey problems and progress in the field of HF COMMUNICATIONS. The lectures cover needs of both the civil and military communities for high frequency communications. Concepts of real time channel evaluation, system design, as well as advances in equipment, in propagation, and in coding and modulation techniques are covered. 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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