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Signaling Issues in the Design of a High-Frequency (HF) Intratask Force Communication Network

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The High-Frequency Intratask Force (HF ITF) Network will use frequency hopping (FH) spread spectrum techniques to provide protection from jamming and interception of messages. The use of spread spectrum signaling leads naturally to the use of code division multiple access (CDMA) techniques, which can be used to provide both multiple access capability and jamming resistance. The use of spread spectrum signaling has had considerable impact on the HF ITF Network, not only from the standpoint of waveform design, but also in regard to network organization and control.

After discussing the complex nature of the HF channel and coding considerations, we present the basis for our choice of FH as the spectrum spreading mechanism and Frequency Shift Keying (FSK) as the modulation scheme. We also discuss the antijamming (AJ) performance provided by M-ary FSK signaling with convolutional coding and optimum diversity, and the resulting communication range that is achievable in a jammed environment.

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The use of wideband FH-CDMA signaling raises issues that are not encountered in narrowband networks. We examine these issues with emphasis on the question of the number of signals that can share a wideband channel simultaneously, while maintaining acceptable performance levels. We also address considerations relating to the design of multiple access protocols that are well suited for FH channels.

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EXECUTIVE SUMMARY

A High Frequency (HF) Intratask Force (ITF) Network is being designed as a robust, survivable, antijam (AJ) communication network for the interconnection of mobile task force elements. The HF ITF Network must provide both point-to-point and broadcast communication modes, and it must support the voice and data traffic requirements of many diverse scenarios. It must reliably link the task force platforms, despite the network connectivity changes that result from platform motion, jamming and other interference, platform destruction, and changing propagation conditions. Furthermore, it must exhibit graceful degradation under severe stress.

A task force may consist of as many as 100 platforms (ships, aircraft, and submarines), which usually span a region with diameter of up to 500 km, and which generally travel together as a unit. The use of the HF groundwave medium (2 to 30 MHz) is dictated by its Extended Line of Sight (ELOS) communication range as well as by its natural survivability properties in postnuclear detonation environments.

The HF ITF Network will use spread spectrum signaling techniques to provide protection from jamming and interception of messages. The use of spread spectrum signaling leads naturally to the use of code division multiple access (CDMA) techniques, which can be used to provide both multiple access capability and jamming resistance. The use of spread spectrum signaling has had considerable impact on the HF ITF Network, not only from the standpoint of waveform design but also in regard to network organization and control.

In our studies we quickly identified three areas of design issues that, although interdependent, could, initially, be pursued in parallel. These areas are:

- The architectural organization of the network;
- The signaling issues (i.e., waveform and coding considerations); and
- The multiple access protocols for the sharing of communication resources.

In a preliminary report [1], the problems encountered in the design of an HF ITF Network were identified and reviewed. In a later report [2], a preliminary system concept for the HF ITF Network was presented. Separate reports have discussed in detail the architectural structure and organization process of the network [3] and the ability of the network to reconfigure itself in response to jamming [4]. A future report will be devoted to multiple access protocols.

In the present report we address the signaling issues that arise in the design of the HF ITF Network in greater detail than we did in [2]. In particular, we demonstrate the relationships among signaling, network organization, and multiple access. Note that we are using the term signaling to refer to waveform design and performance considerations, whereas the same term is often used to refer to communication link setup procedures.

The dominant factor in determining many of the ITF Network's features that distinguishes this network from other radio networks is, in fact, the special nature of the HF medium. The HF radio channel is a fading and dispersive medium, which supports propagation via both groundwaves and skywaves. The ITF Network will rely primarily on the use of groundwaves to connect nodes, because, although groundwave attenuation varies with frequency and sea state conditions, it is much more predictable and less dispersive than skywave propagation, which is supported at any time only over a

portion of the HF band. Skywave signals must be considered as a source of multipath interference. The use of skywaves to supplement groundwave paths within the ITF Network and to link the ITF Network with other networks will be considered in the future.

In addition to multipath, sources of interference in the HF channel include atmospheric noise, galactic noise, local platform noise, and other-user interference. In the networking environment other-user interference may be divided into two classes, i.e., interference resulting from platforms external to the network, and interference caused by other network members or other platforms with which the network is communicating. Interference caused by other task force platforms is characteristic of a networking environment, and it is clear that network management schemes are needed to minimize the effects of such interference. A major part of the HF ITF networking study has, in fact, been directed toward this problem. Interference arising from platforms external to the network and interference due to jamming cannot be controlled by the network. Since jamming represents the most significant limitation and threat to HF communication, we have given special attention to this problem.

We have considered the signaling aspects of point-to-point communication and of multi-user communication for both narrowband and wideband environments. The following point-to-point discussion clearly brings forth many of the issues. The subsequent multi-user discussion complements this understanding by identifying additional issues that are unique to networking. Study of both narrowband and wideband networking considerations is necessary, since, as new equipment is developed, wideband signaling will gadually replace the narrowband signaling that is used at the present time. As a result, it is necessary to use a hybird form of operation in the interim.

Point-to-Point Communication

Narrowband signaling is not believed to be capable of providing satisfactory protection from anticipated jamming threats, even with the use of adaptive antenna arrays and selective band use. The use of wideband (spread spectrum) signaling, however, can provide considerable protection from jamming. In addition, spread spectrum signaling provides selective addressing capability, CDMA capability, inherent privacy and security, low interceptibility, and high resolution ranging. We have concluded that, among the candidate schemes for spreading in the HF channel, a pure Frequency Hopping (FH) system is the most practical choice for use in the ITF Network, largely because FH systems are more robust than Direct Sequence (DS) systems. Specifically, FH systems are virtually immune to problems caused by differences in relative signal strength, are less sensitive to dispersion, and do not require a contiguous bandwidth. Also, in these systems the spreading factor can be considerably greater, and the pseudonoise (PN) code acquisition is easier to achieve and more difficult to disrupt. A hybrid FH-DS system can combine many of the advantages of FH and DS systems, but is more difficult to implement.

Another issue in any form of signaling is the selection of the modulation method. Noncoherent M-ary Frequency Shift Keying (MFSK) is a practical scheme for the HF ITF Network because of its feasibility of implementation, its robustness with respect to fading and interference, and its compatibility with a wide range of hopping rates.

Bit error rate (BER) performance under worst case partial band (WCPB) noise jamming has been evaluated as a function of the equivalent signal to noise ratio for FH-MFSK signaling. Performance tradeoffs were developed for several alphabet sizes ($M = 2, 4, 8$), many values of diversity, several coding schemes, both hard and soft decision receivers, the availability or lack of Jammer State Information (J.S.I.),* and for nonfading as well as Rayleigh fading channels. From these results it is concluded that coding and/or diversity are essential for jamming resistance in the HF ITF Network employing FH-MFSK, and that the use of coding greatly reduces the need for diversity.

*J.S.I. is knowledge at the receiver of which symbols are jammed.

Most of our results are based on the use of a soft decision receiver with J.S.I. In this case, when operating under a high quality data reception criterion (BER = 10^{-5}), antijam (AJ) coding/diversity gains of 35 to 37 dB are obtained by using constraint length 7 convolutional coding of rate 1/2 to 1/3 with an optimum diversity of 2* for the nonfading channel, and 3 or 4 (depending on alphabet size) for the Rayleigh fading channel. If the data reception criterion is relaxed to a medium quality level (BER = 10^{-3}) the AJ coding/diversity gain is 16 to 18 dB, again for codes of rate 1/2 to 1/3. For the nonfading channel no diversity is required beyond that inherently provided by the coding process; for the Rayleigh fading channel a diversity of 2 or 3 is optimum.

We have analyzed in detail the tradeoffs necessary to achieve acceptable levels of AJ performance. For example, under some reasonable assumptions, it is possible to operate links at a jammer to signal power ratio between 25 dB (at a data rate of 2400 b/s) and 40 dB (at 75 b/s) for a nominal case of 8-ary FSK signaling, a nonfading channel, 5 MHz spread bandwidth, 10^{-5} BER, soft decision receiver, WCPB noise jamming, and known jammer state. Results for 4-ary FSK are within 0.7 dB of those for 8-ary FSK. When the BER requirement is relaxed to 10^{-3}, and all other conditions are unchanged, the achievable J/S ratio is increased by about 1.5 dB.

Our estimates of jammer to signal levels at which satisfactory communication can be maintained may have to be reduced somewhat after some issues related to practical implementation are examined. For example, in the case of a hard decision receiver, if J.S.I. is available, the performance is approximately 1.3 to 1.5 dB worse than that of the ideal soft decision receiver with J.S.I. over a wide range of BER. If J.S.I. is not available, there is further degradation, much of which can be eliminated through the use of diversity of 2 or greater.

Estimates of the communication range that can be achieved in a jammed environment have been derived, based on a model for groundwave propagation loss in conjunction with the maximum jammer to signal power levels that can be tolerated.

Multipoint (Network) Communication

In considering multipoint or network communication, the signaling issues become considerably more complex because of the potential interference among the users. In a narrowband architecture, there are a number of frequency bins to be managed *(switched)* and distributed to the users. The new issue then is the development of link management schemes that permit the efficient apportionment of these frequency bins to the users via either dedicated or shared links. In a wideband architecture, such as that envisioned for the HF ITF Network, the use of spread spectrum signaling provides an inherent natural means of multiplexing the different users with acceptable levels of interference. This is achieved by CDMA techniques, in which, for the proposed choice of frequency hopping, each FH pattern corresponds to a distinct code. CDMA operation is usually asynchronous, and therefore it is possible for two or more users (using different hopping patterns) to transmit simultaneously at the same frequency, resulting in loss of data. The loss of data caused by such *frequency hits* can be handled via the use of coding and/or diversity.

The number of local transmitters using quasi-orthogonal FH patterns that can use the same wideband channel simultaneously, while maintaining acceptable performance levels, is approximately one tenth of the number of frequency bins into which the channel is divided and over which the users are hopping. The exact number depends on factors such as signal to noise ratio, modulation scheme, coding, diversity, and acceptable performance criterion, e.g., BER or probability of successful packet reception. Considerable improvement over this number can be achieved if side information is available, i.e., knowledge of which symbols have been corrupted by hits. Frequency reuse at distant parts of the network is of course possible. Nevertheless, the limitation in interference rejection capability that can be

* A diversity of m means that each symbol is redundantly transmitted m times.

provided by a pure FH-CDMA system requires the development of additional control and channel-sharing ideas. Important issues specifically related to the use of FH-CDMA include:

- Synchronization requirements and hopping rates;
- The generation of FH patterns;
- Interference in multiple user FH channels;
- The assignment and distribution of codes (i.e., FH patterns) to platforms;
- Contention among signals using the same code;
- Multiple access protocol considerations for FH networks;
- The coordination of different types of network traffic.

Baseline HF ITF Network System Concept

The work accomplished thus far in the design of the HF ITF Network has resulted in a baseline concept for the architecture of the network, for the signaling methods, and for the multiple access protocols. This provides a comfortable framework for the study of the remaining design issues. Table ES.1, taken from [2], indicates our recommendations for a survivable HF ITF Network design.

Outline of Report

In Section 1 of this report we present an overview of the HF ITF Network. We briefly discuss the HF ITF Network operational requirements and constraints, and we describe the Linked Cluster Architecture, which is central to the HF ITF Network concept.

The design of the HF ITF Network has been greatly influenced by the properties of the HF band. In Section 2 we discuss the most important features of both groundwave and skywave HF propagation, as they relate to the HF ITF environment.

Error control coding is needed to correct errors caused by background noise, other-user interference, jamming, and fading. In Section 3 we discuss coding considerations for the HF ITF Network.

In Sections 4 and 5 we address considerations relating to the use of narrowband and wideband signaling, respectively. Our emphasis has been on wideband signaling because of the need to provide an AJ capability. In particular, in Section 5 we present our reasons for the choice of FH as the spectrum spreading mechanism and FSK as the modulation scheme.

In Section 6 we discuss basic power budget issues, and we present a communication range model for HF groundwave communication in a benign environment.

In Section 7 we discuss the AJ performance that can be achieved through the use of FH M-ary FSK signaling and M-ary convolutional coding. We demonstrate that an AJ capability can be achieved for HF ITF Network links. In Section 8 we demonstrate that narrowband signaling schemes are extremely vulnerable to jamming. In Section 9 we use the results of Sections 7 and 8 in conjunction with a model for HF groundwave propagation loss to obtain a model for the communication range that can be achieved in a jammed environment.

In Section 10 we discuss the issues that arise in a networking (or multilink) environment, with emphasis on wideband signaling. The major issues include the ability of a wideband FH channel to support a number of users simultaneously, as well as multiple access protocols that are suitable for such a channel.

In Section 11 we discuss the issues associated with the synchronization of FH signals, including their impact on the overall network performance.

Table ES.1 — Baseline HF ITF Network System Concept

NETWORKING: Robust Network Design	**SIGNALING**: Robust Waveform Design
• **Number and Types of Nodes**: 2 to 100; ships, aircraft, and submarines.	• **Frequency Band**: HF (2 to 30 MHz).
• **Task Force Dispersion**: Approximately 500 km diameter (ELOS communication ranges).	• **Propagation Medium**: HF groundwave.
• **Network Structure**: Overlay of several (approximately five) networks in different portions of HF band.	• **AJ Scheme**: Frequency hopping spread spectrum.
• **Network Architecture**: Overlapping clusters of nodes within each of these several frequency subbands.	• **Modulation**: M-ary FSK (M = 4 or 8 are possible choices).
• **Network Control**: Hybrid (centralized local control within clusters, and distributed operation among clusters).	• **Receiver Detection**: Noncoherent.
• **Intercluster Operation**: Protocols will include features of broadcasting, contention, and reservation. They must be compatible with frequency hopping spread spectrum signaling.	• **Coding and/or Diversity**: Required to combat jamming, fading, and other-user interference. Both convolutional and block coding are being considered.
• **Intercluster Operation**: Dedicated links over backbone network and auxiliary links when possible; demand-assigned or contention-based when necessary.	• **Signaling-Related Parameters**: — Data Rates: 75 to 2400 b/s — Acceptable BER for data: 10^{-3} to 10^{-5} — Typical hopping bandwidth (bandwidth of one of the subbands): 2 to 5 MHz.
• **Topological Changes**: Adaptive and robust network control (continual updating of connectivities).	• **Achievable AJ Protection Ratios***: Worst Case Partial Band — 2400 b/s — 75 b/s Jammed Nonfading Channel: 24.9 dB $\leqslant$ J/S $\leqslant$ 39.9 dB Rayleigh Fading Channel With Broadband** Jamming: 22.7 dB $\leqslant$ J/S $\leqslant$ 37.7 dB

*For the case of convolutional coding with optimum diversity, 8-ary FSK, 5 MHz spread hopping bandwidth, BER = 10^{-5}, soft decision receiver, and known jammer state (i.e., receiver can determine which symbols are jammed).
**Broadband noise jamming is worst case partial band noise jamming for the Rayleigh fading channel.

Finally, in Section 12 we present our conclusions from this research, and we outline our future plans.

Ten appendixes provide more detailed discussion and analysis where needed.

Future Efforts

Our future efforts will proceed within the framework provided by the baseline concept presented here. A major effort will be devoted to the development of channel access protocols that are consistent with the Linked Cluster Architecture and with the constraints imposed by the use of FH-CDMA signaling. A concept of operation that encompasses all aspects of network design will be developed.

A number of design issues specifically related to signaling must be investigated further. Among the most important are the acquisition and maintenance of synchronization of the frequency hopped signals. Other signaling issues include the choice of the alphabet size M for M-ary FSK signaling, the hopping bandwidth, and the number of simultaneously functioning Linked Cluster networks. Channel models will be enhanced to account for the joint effects of jamming, other-user interference, and background noise.

SIGNALING ISSUES IN THE DESIGN OF A HIGH-FREQUENCY (HF) INTRATASK FORCE COMMUNICATION NETWORK

1.0 INTRODUCTION

1.1 HF ITF Network Background

A High Frequency (HF) Intratask Force (ITF) Network is being designed as a robust, survivable, antijam (AJ) communication network for the interconnection of mobile task force platforms. These platforms will consist of ships, aircraft, and submarines. The HF ITF Network must provide both point-to-point and broadcast communication modes, and it must support the voice and data traffic requirements of many diverse scenarios. Furthermore, it must exhibit graceful degradation under severe stress, including jamming, platform loss, and heavy traffic requirements. The use of the HF groundwave medium (2 to 30 MHz) is dictated by its Extended Line of Sight (ELOS) communication range as well as by its natural survivability properties in postnuclear detonation environments.

In our studies we quickly identified three areas of design issues that, although interdependent, could be pursued, initially, in parallel. These areas are:

- The architectural organization of the network;
- The signaling issues (i.e., waveform and coding considerations); and
- The multiple access protocols for the sharing of communication resources.

An understanding of each of these areas is vital to the design of the HF ITF Network. Furthermore, as our studies have progressed, we also have developed a better understanding of the relationships among these areas.

The proposed network organization, which we call *Linked Clusters,* has a hybrid structure that mixes distributed and centralized control by grouping together sets of platforms that are within one-hop distance from a central platform into a centrally controlled cluster, and by allowing distributed operation among *cluster heads.* These clusters are established by the execution of a fully distributed algorithm [3,5-10].

The HF ITF Network will use spread spectrum signaling techniques to provide protection from jamming and from interception of messages. The use of spread spectrum signaling leads naturally to the use of Code Division Multiple Access (CDMA) techniques, which can be used to provide both multiple access capability and jamming resistance. The use of spread spectrum signaling has had considerable impact on the HF ITF Network, not only from the standpoint of waveform design but also in regard to network organization and control.

In a preliminary report [1], the problems in the design of an HF ITF Network were identified and reviewed. In a later report [2], a preliminary system concept for the HF ITF Network was presented. Separate reports have discussed in greater detail the architectural structure and organization process of the Network [3] and the ability of the network to reconfigure itself in response to jamming [4]. A future report will be devoted to multiple access protocols.

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In this report we address the signaling issues that arise in the design of the HF ITF Network in greater detail than we did in [2]. Note that we are using the term signaling to refer to waveform design and performance considerations, whereas the same term is often used to refer to communication link setup procedures. In particular, we demonstrate the relationships among signaling, network organization, and multiple access.

1.2 HF ITF Network Requirements and Constraints

In this section we briefly review some of the major operational requirements and environmental constraints imposed upon the HF ITF Network when used as the primary ELOS communication system in the intratask force environment. These topics are discussed in greater detail in [1].

A task force consists of as many as 100 mobile platforms that travel together as a unit; usually they are located within a region 500 km in diameter. The network is characterized by a variable topology that results from changes in the HF radio connectivity of the platforms; these connectivities change as a result of varying radio wave propagation conditions and noise levels, hostile jamming, interference from other members of the task force and other sources, platform destruction, and platform mobility.

The network must be able to support both voice and data traffic of several priorities with various security requirements at acceptable error rates. Both point-to-point and broadcast modes of operation must be supported. An internetting capability with other military communication networks must also be developed.

Survivability and robustness are among the most important requirements of the HF ITF Network; the network must degrade gracefully under stress conditions. Network degradation will usually be caused by loss of nodes or degradation (or loss) of links. The major threat to network links is hostile jamming; however, link quality is also affected by changing propagation conditions, platform mobility, and other-user interference. Furthermore, the increased traffic requirements during periods of crisis can cause network overloads that result in degraded performance.

1.3 The Linked Cluster Architecture

The issue of survivability has greatly influenced the design of the network. The use of decentralized network control reduces the vulnerability associated with a single central controller. Furthermore, the proposed architecture is based on the use of fully distributed algorithms that enable the task force platforms to self-organize into a reliable network structure, and to continually monitor the changing connectivities for the maintenance of such a structure. A disadvantage presented by a completely distributed control structure, however, is that significant communication resources are needed to maintain consistent data bases at each platform. This problem is especially significant in the HF band, where data rates are often limited to, at most, 2400 b/s. Therefore, we have proposed a hybrid structure known as the Linked Cluster Architecture. This structure, illustrated in Fig. 1.1, consists of clusters of platforms within communication range of local controllers known as cluster heads. The architectural profile of the network, at any given moment, consists of clusters that are linked to each other via gateways. We note that, when used with typical shipboard antennas, HF is a broadcast medium, i.e., it is nominally omnidirectional rather than highly directional as is SHF. Therefore, all platforms within communication range can simultaneously monitor the transmissions of, e.g., their cluster head.

The *backbone network* is defined to be the set of cluster heads and gateways, and the point-to-point links that connect them. It is represented by the striped links shown in Fig. 1.1. The *primary network* is defined to be the backbone network plus those links that connect cluster members with their heads. It is represented by the striped plus solid links. Network links that are not part of the primary network are known as *auxiliary links.*

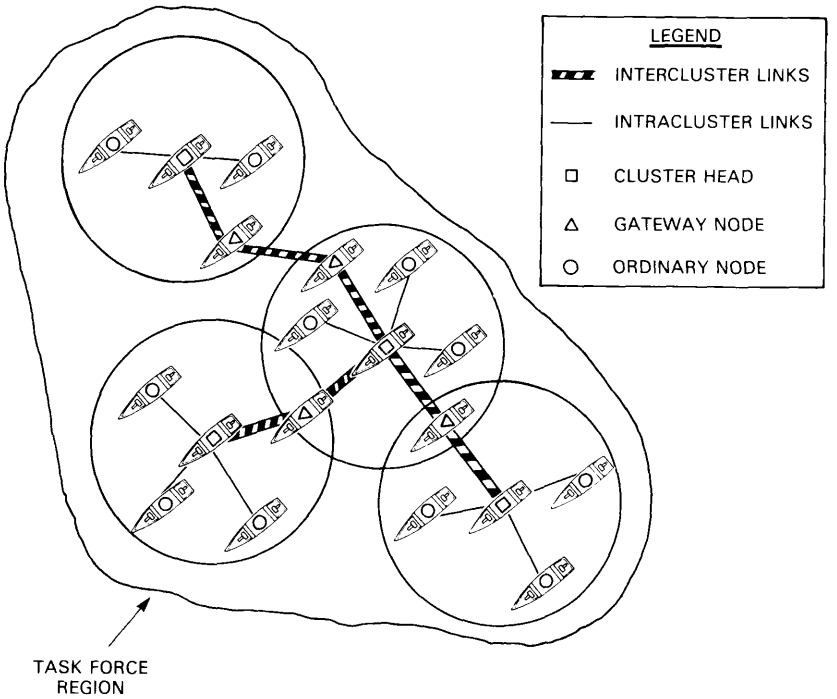


Fig. 1.1 — Network organizational structure—Linked Cluster Architecture

The network organization algorithms make use of connectivity information exchanged between neighboring platforms. The algorithms consist of two time division multiple access (TDMA) frames, after which each platform is able to determine whether it should assume the role of cluster head or gateway, or to remain an ordinary node [3,5-10]. The algorithms may be executed periodically to form new network structures that are based on current connectivities. We emphasize that although a hierarchical network structure is implemented, the network organization algorithms themselves are, in fact, totally distributed.

The HF groundwave medium is characterized by a communication range that varies as a function of frequency. Typically, communication range decreases as frequency increases. The Linked Cluster Architecture takes advantage of this apparent shortcoming of the HF medium by partitioning the HF band into a number of subbands, each with a bandwidth of a few megahertz over which the groundwave communication range is approximately constant. The organization algorithm is run consecutively for each subband, thus producing a set of overlaid connectivity maps that give rise to a set of simultaneously operating networks. The HF ITF Network consists of this set of individual networks that are defined in separate subbands. Network management schemes are presently under development to coordinate the operation of each of the individual networks into an effective overall ITF Network structure. We note that the network organizational algorithms are run independently in each subband. Thus, while one of the Linked Cluster Networks is reorganizing, the others maintain communication by using their most recently derived network structure. This capability of preserving a communication structure during the reorganization process adds considerably to the robustness of the HF ITF Network.

A networking structure inherently provides resistance against jamming by providing relays as they are needed along the communication paths. The use of one or more relays located between the source and destination platforms results in shorter links, and therefore improved signal to interference levels. Furthermore, an adaptive routing capability often permits networks to send traffic along paths that avoid the most heavily jammed portions of the network.

In particular, the HF ITF Network will periodically reorganize itself to respond to topological changes (caused, e.g., by platform motion and hostile jamming) by designating a new set of clusters and routes, as has been demonstrated by Baker et al., [11,4]. The ability to maintain a sufficiently high degree of connectivity is, of course, limited by the AJ performance of the individual network links. Only limited AJ performance is expected to be obtainable with narrowband signaling, and thus we have proposed spread spectrum signaling for the HF ITF Network.

2.0 HF CHANNEL CHARACTERIZATION

The design of the HF ITF Network has been greatly influenced by the properties of the HF band (2 to 30 MHz). The HF radio channel is both fading and dispersive, with propagation occurring via both groundwaves and skywaves. The HF ITF Network will rely primarily on the use of groundwaves to connect nodes. Groundwave attenuation varies with frequency and sea state conditions, but it is generally more predictable than skywave propagation, which is supported at any time only over a portion of the HF band. Skywave signals will typically be considered as a source of multipath interference in our studies, although the use of skywaves to advantage as a form of diversity reception may be considered in the future.

2.1 HF Groundwave Propagation

The HF medium has been designated as the primary medium for ELOS intratask force communication, largely because of the ELOS propagation ranges (up to several hundred kilometers) of HF groundwaves. Another advantage of HF groundwave propagation is that it is expected to be only minimally affected by nuclear detonations. Besides being more predictable than skywave propagation, groundwave propagation has the advantage of having little dispersion; the use of groundwaves thus permits considerably greater signal bandwidths, and therefore data rates, than the use of skywaves. Furthermore, the use of HF groundwaves avoids the need to use relays external to the task force, such as geostationary satellites operating at UHF, SHF and EHF, that are potentially vulnerable to physical attack.

Figure 2.1, taken from Barrick [12], illustrates typical groundwave path loss over a smooth sea as a function of range, as frequency is varied between 3 and 50 MHz. Vertically polarized waves are most effective for groundwave propagation, because horizontally polarized waves are greatly attenuated at the sea-air interface. It is clear from this figure that groundwave attenuation increases as frequency increases. The lower end of the HF band (2 to 6 MHz) is therefore especially attractive for ELOS ITF communication.

The effects of rough sea are illustrated in Fig. 2.2 [13], which shows that propagation loss increases as sea state increases, and that this effect becomes more pronounced as frequency increases.

Despite the apparent advantages of the lower part of the HF band, it would not be advisable, for a number of reasons, to restrict all of our groundwave communication to these frequencies. This region is subject to skywave interference at certain times of the day (mainly at night); it would become overcrowded if all platforms restricted their communication to this subband; finally, the network would be more susceptible to jamming because only a small part of the available spectrum would have to be covered by a jammer. Our approach, as discussed in Section 1, has been to divide the HF band into a number of subbands, over each of which the propagation loss, and therefore connectivity, is relatively constant.

2.2 HF Skywave Propagation

HF skywave propagation, which relies on the refraction of radio waves by the ionosphere, is normally used for Beyond Line of Sight (BLOS) communication at ranges as great as thousands of

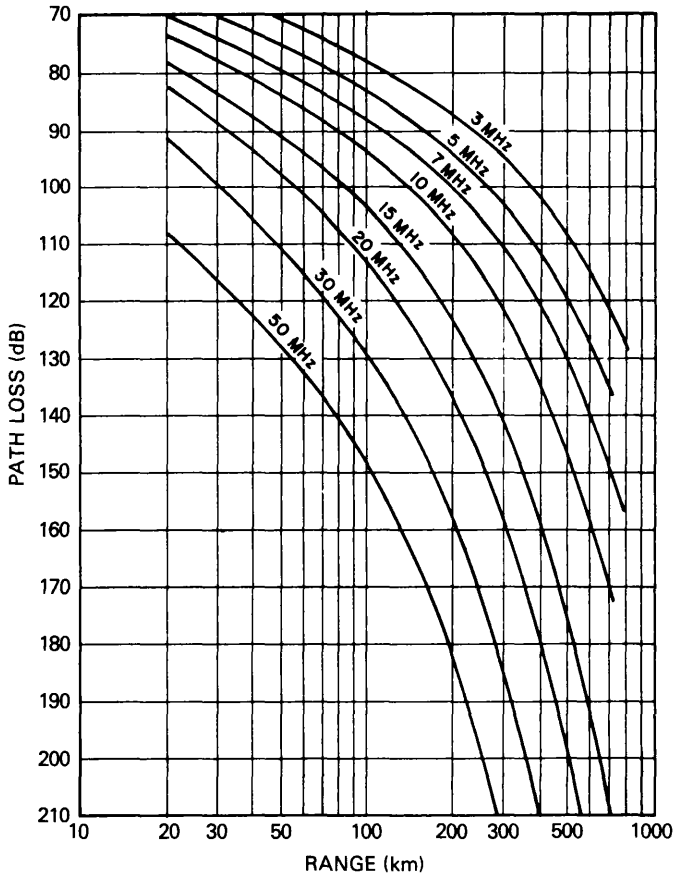


Fig. 2.1 — HF groundwave path loss over a smooth sea (from [12])

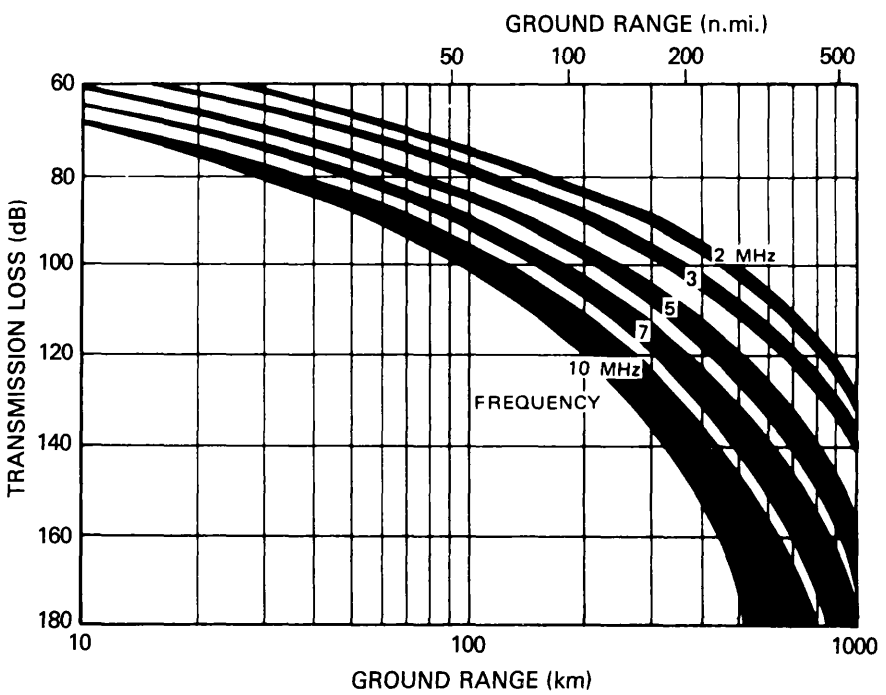


Fig. 2.2 — Path loss across the ocean for sea states from 0 (top of each black band) through 6 (bottom of each black band) (from [13])

kilometers. Skywave signals are often generated along with the desired groundwave signals, and in such cases they will appear at the receiver as interference in the form of multipath signals. Skywave propagation must be considered in the design of the HF ITF Network, not only because it is a source of interference, but also because the network will have to interface with other networks via long haul skywave links, and because it may be necessary to include skywave links within the HF ITF Network.

The diurnal variation of the maximum usable frequency (MUF) and lowest usable frequency (LUF) over a *typical* skywave link of several thousand kilometers is shown in Fig. 2.3. Also shown is the frequency of optimum transmission (FOT), which is usually defined for conventional narrowband signaling techniques as 85% of the MUF. The links of the ITF environment are not *typical,* however, since transmission ranges are at most hundreds rather than thousands of kilometers. The MUF for transmissions within the task force will not be much greater than the critical frequency (i.e., the highest frequency that is reflected at vertical incidence) in most cases, because the angle of incidence is nearly vertical; typically, the MUF for intratask force communication will be less than 10 MHz.

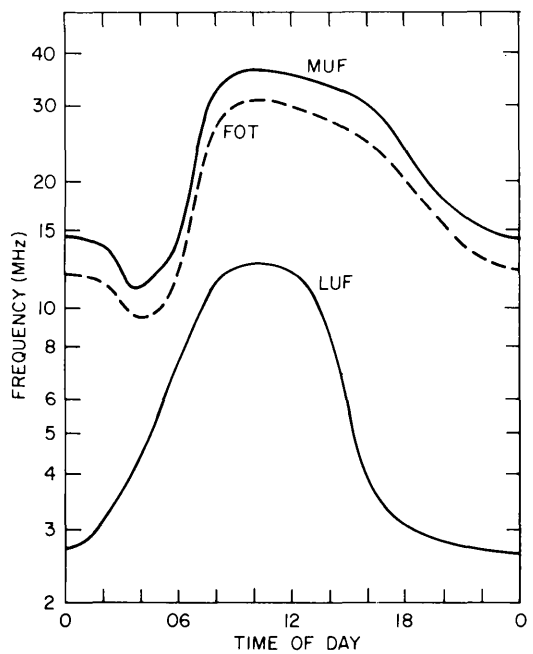


Fig. 2.3 — Diurnal variation of the maximum usable frequency (MUF), the lowest usable frequency (LUF), and the frequency of optimum transmission (FOT) on a typical long-range HF skywave link

Knowledge of the portion of the HF band over which skywave propagation is supported can often permit the choice of an operating frequency that either uses or avoids this mode. In the ITF environment, where we are relying mainly on groundwaves, multipath signals can be eliminated by avoiding the use of this band, as can the effects of skywave signals caused by other-user interference and jamming. An important consideration is that, although the MUF for true ITF propagation ranges is usually relatively low because of the short ranges involved, the network must cope with external sources of interference. Therefore, from an interference viewpoint, the *typical* curve of Fig. 2.3 may in fact be appropriate. Also, when Low Probability of Intercept (LPI) and Limited Range of Intercept (LRI) communication modes are being used, the region of the HF band that supports skywave propagation should be avoided. The choice of transmission frequency to avoid undesired skywave propagation is known as *propagation tactics* [14].

Typical curves, such as those shown Fig. 2.3, are of limited use because the characteristics of the HF channel depend not only on frequency and time of day, but also on range, geographical location,

season, and sunspot number. For example, during periods of high sunspot activity, the critical frequency may be twice that at low sunspot activity for otherwise identical conditions [15]. Also, *pathological* effects such as magnetic storms, sudden ionospheric disturbances, and sporadic *E* layer propagation make the estimation of channel characteristics difficult.

2.3 Sources of Interference in the HF ITF Environment

The primary sources of HF interference in the ITF environment are: atmospheric and galactic noise, locally generated platform interference, multipath interference resulting from undesired skywave propagation, other-user interference, and hostile jamming.

A considerable amount of data is available for characterizing natural atmospheric and galactic noise in the HF band. Satisfactory techniques have been devised for computing atmospheric HF noise levels at various geographic locations, and for various seasons and hours of the day [16-18]. Typically, atmospheric noise predominates below about 15 MHz, and galactic noise predominates above this frequency. The noise distributions are non-Gaussian, nonwhite, nonstationary, and highly impulsive, and therefore it is difficult to characterize them mathematically [19]. Furthermore, locally generated platform noise is very difficult to predict, and may be considerably greater in amplitude than the ambient noise. Noise levels in frequency subbands that support skywave propagation are considerably higher at night, when the absorbing *D* layer disappears, than they are during the daytime. In practice, the communication range that can be achieved depends on frequency, transmitter power, data rate, modulation and demodulation schemes, receiver structure, antenna gains and/or losses, other system losses, acceptable error rates, and interference levels. The effects of the other sources of interference are much greater than the effects of atmospheric and galactic noise in the stressed environment that we are assuming for the ITF Network. Therefore, estimates of communication range based solely on background noise can only be considered as representative of a best-case environment, rather than a realistic one. Curves of this type (such as those presented in Section 6) are useful, however, to illustrate the dependence of communication range on frequency.

Noise and distortion considerations relating to the hardware design of a new wideband HF system architecture are discussed in [19a and 20]. Of all the issues considered, the one most directly related to the *networking* area is that of the required separation between simultaneous transmitting and receiving frequencies on any platform. Under the proposed design, it is believed that successful reception can be achieved if a frequency separation of at least 2.5% or 100 kHz (whichever is larger) is maintained. It is also important to note that locally generated electromagnetic interference (EMI) rises as transmitter power increases; however, the new wideband architecture is expected to reduce the EMI considerably.

Multipath interference can be troublesome whenever the undesired skywave signals are greater than about 10 dB below the desired groundwave signals. Two basic situations may arise. If the differential time delay between groundwave and skywave signals is short in relation to the symbol duration, and the signals are of comparable magnitude, then the noncoherent combining of the signals may result in fading. If the delay is larger, then intersymbol interference will occur. Methods to avoid such multipath effects include: the use of spread spectrum signaling techniques such as direct sequence and frequency hopping; the use of modulation techniques that are robust with respect to such interference (e.g., noncoherent frequency shift keying (FSK)); the use of channel equalizers, or avoidance of those parts of the HF band that result in such multipath.

Measurements of skywave/groundwave received signal power ratios over seawater paths in the band from 2 to 10 MHz were performed by Hipp and Green [21]. Some of their major conclusions are as follows. Groundwave propagation was shown to be generally predominant (i.e., greater than 10 dB above the skywave signal) for the majority of the measured data set in the 2 to 4 MHz frequency range, and for terminal-to-terminal path distance less than 400 km. However, the sunrise and sunset periods often provide good support for 2 to 4 MHz skywave propagation, resulting in interference even at

ranges less than 400 km. In the range from 4 to 10 MHz, mixed skywave/groundwave propagation can be anticipated over a large percentage of the 24-hour diurnal period. Their measurements agree, generally, with theoretical predictions by Freres, Sailors, and Taylor [22]. Experiments are currently under way by NRL researchers to further characterize relative signal strengths as well as the coherence bandwidths for groundwave and skywave signals [23]. The results of such studies must be used with care, however, because of the often unpredictable nature of the HF skywave channel.

Other-user interference is usually the principal source of interference in the HF band, and can be tens of dB greater than the atmospheric noise level [19,24,25]. In the networking environment there are two main classes of other-user interference:

- Interference caused by platforms external to the network,
- Interference caused by other network members (or other platforms with which the network is communicating).

The main characteristic of the former is that it is not controllable; it is part of the environment in which our network must function, and it will be considered to be part of the background noise level. Like atmospheric noise, other-user interference levels in frequency bands that support skywave propagation are considerably higher at night than during the daytime.

Interference that results from other task force platforms is characteristic of a networking environment, and it is clear that network management schemes are needed to minimize the effects of such interference. A major part of the HF ITF networking study has, in fact, been directed toward this problem. Other-user interference can be avoided if orthogonality is somehow provided among the transmitted signals in the time domain, frequency domain, or joint time-frequency domain. In practice, this orthogonality is difficult to maintain; however, there are time, frequency, and code division multiple access techniques that can achieve acceptable levels of performance in many applications as is discussed in Section 10. The problem of other-user interference is especially severe in the portion of the HF band that supports skywave propagation, because in this case all members of the network can potentially interfere with each transmission, rather than the limited subset that is within groundwave propagation range of a particular platform. Point-to-point measurements, such as those of Hipp and Green, are no longer adequate to assess the interference levels for a particular path, since interference can come not only from the same platform that is transmitting the desired signal, but also from platforms at virtually any range (including those external to the task force, and in fact from all parts of the world).

Jamming considerations are similar to those for other-user interference. However, jammers are obviously not controllable, and therefore they are not subject to network management techniques, nor can their signals be modeled simply as part of the background noise. The network must be designed to circumvent this form of interference. Even when only groundwave links are used in the network, the effects of jamming signals that propagate via skywaves must be considered, since it is usually not feasible to separate the propagation modes at the receiver. The frequency band over which the network will be susceptible to skywave jamming signals will be determined by the relative positions of the jammer and the platforms being jammed. Although the MUF for task force communication ranges is usually relatively low, the MUF corresponding to distant jammers is considerably higher (it would be similar to that shown in Fig. 2.3) resulting in troublesome interference over a wider part of the HF band than would be expected on the basis of the MUF for communication between task force platforms.

3.0 CODING CONSIDERATIONS FOR THE HF ITF NETWORK

As discussed in Section 2, the HF band is characterized by large amplitude highly non-Gaussian interference, which results in the corruption of signals. Error control coding is therefore needed to

reduce the bit error rate (BER) to acceptable levels. In this section we discuss some of the considerations related to the use of coding in the HF ITF Network. Besides the modeling of channel disturbances, important considerations include the use of FH signals and the use of packet switching rather than continuous data streams.

A real-world channel model for the HF ITF Network is illustrated in Fig. 3.1. As a result of the magnitude and variety of the noise sources and their possible presence or absence, the BER can be considerably greater than on an additive white Gaussian noise (AWGN) channel with the same average noise spectral density. Furthermore, the channel is no longer memoryless. That is, the raw channel error rate (i.e., the fraction of binary symbols received in error), $p(t)$, coming out of the MFSK demodulator will be time-varying. In general, besides the essentially random errors caused by atmospheric noise, there may be bursts of errors created by either jamming, other-user interference (caused by either task force members or other HF users), multipath propagation, fading, locally generated platform noise, or any combination thereof. The modeling of the HF channel and the evaluation of code performance on this channel are therefore very difficult. The purpose of the interleaver/de-interleaver operations is to make the channel appear memoryless to the decoder, so that errors will appear to be random, thereby facilitating the decoding process in many cases. The channel may then be characterized simply by some raw channel error rate, p. Alternatively, some coding schemes can take advantage of the bursty nature of the error process, in which case interleaving would not be desirable. In general, a burst error correcting code does not perform well on random error channels, and vice versa.

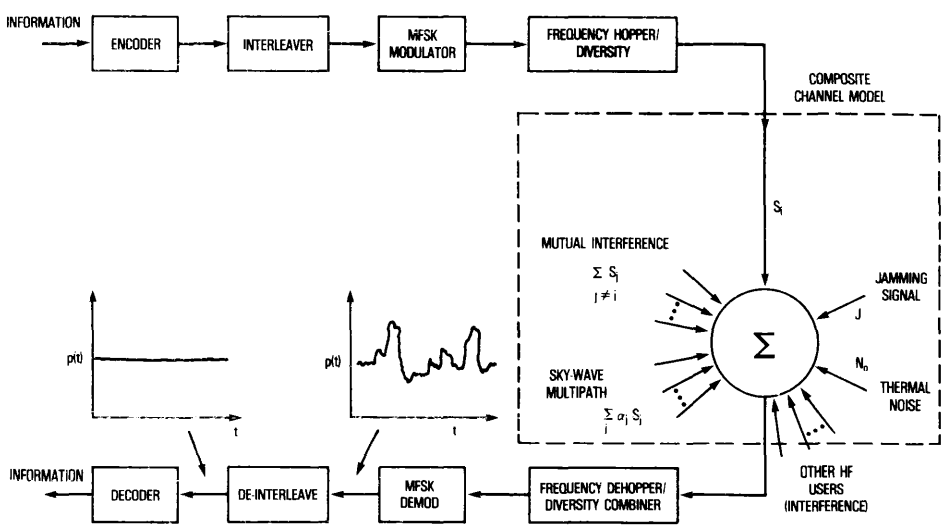


Fig. 3.1 — Generalized channel model for the HF ITF Network

Link performance depends not only on channel propagation conditions and interference, but also on waveform design and signaling format. First, BER performance depends on the type of modulation that is used. We have proposed noncoherent FSK for the HF ITF Network for the reasons discussed in Section 5.2. More subtle considerations relate to the packet-switched nature of much of the HF ITF Network traffic. The degree of interleaving that is necessary depends on the number of symbols transmitted per hop as well as on the characteristics of the interference. The use of single relatively short packets (known as datagram traffic) limits the degree of interleaving that is possible. Thus, although an assumption of ideal interleaving may be valid (or nearly so), when dealing with stream traffic where one can interleave over a relatively large number of packets, it may not be valid when dealing with datagram traffic. Even in the case of stream traffic, however, interleaving over a large time period may result in excessively large encoding delays.

The use of any coding scheme on a FH channel generates some degree of frequency diversity, because symbols corresponding to each information bit are then transmitted in more than one frequency bin, thereby reducing the effect of frequency-dependent losses.

The use of coding has a tremendous impact on link performance. For example, for the M-ary convolutional coding schemes discussed in Section 7, the E_b/N_0 required to achieve a BER of 10^{-5} on an AWGN channel is about 10 dB. The corresponding value of p for these cases is typically about 0.1 or somewhat greater.

Several classes of coding approaches are available for consideration:

- Automatic Repeat Request (ARQ)
- Block codes
- Convolutional codes
- Concatenated codes

The selection of one or more of the coding schemes listed above depends on our understanding of the channel disturbances. Also, different criteria are applicable depending on the nature of the traffic. For example, voice traffic can tolerate considerably higher error rates than data can, but it cannot tolerate excessive delays. Data may or may not be time-critical, and it may be subject to different BER performance requirements, such as *high quality* (10^{-5}) or *medium quality* (10^{-3}). In Appendix B we discuss these coding schemes, and we demonstrate that each has advantages and disadvantages. At this point, although we are not yet in a position to decide on the coding scheme that will be used in the HF ITF Network, we can make a number of observations.

3.1 Summary of Coding Considerations

The use of ARQ in combination with some form of forward error control (FEC) scheme is a possibility for those network services that can tolerate the delay that is associated with the retransmission process. In cases where such delays cannot be tolerated (e.g., voice or time critical data), or if provision to request retransmission cannot be implemented, then one must rely exclusively on FEC techniques.

The choice of a coding scheme must be made in conjunction with the waveform design. A major consideration for FH systems is the number of symbols (binary or M-ary) that are transmitted per hop. If a single symbol is transmitted in each hop, then codes designed for random errors are generally best, although there can be sources of interference that affect much of the hopping bandwidth, thereby necessitating some burst error correction capability. If several symbols are transmitted per hop, then a code with a burst error correcting capability should be used, or alternatively, interleaving may be used to randomize the errors to make them correctable by a random error correcting code. However, the degree of interleaving that is possible may be limited in packet-switched systems.

Reed-Solomon (RS) codes are extremely powerful burst-error-correcting block codes. They have excellent error-correction capability in burst error environments, but they are perhaps not as good as some other schemes in random error environments. An important property is that they are characterized by an extremely low probability of undetected error. Their use in slow FH systems in wideband multi-user channels is thus very attractive. They also perform very well in jamming environments. A possible disadvantage of the use of Reed-Solomon codes in slow FH systems (such as those discussed in Appendix I) is that they place a constraint on the signaling scheme. For example, the use of a RS-(31,15) code would require a signaling scheme in which one or more RS symbols (32-ary symbols or a multiple of 5 binary symbols) are transmitted in each hop. The corresponding packet size is constrained to be an integral number of codewords. Whether or not these factors are detrimental to system operation depends on the modulation scheme being used, packet size requirements, etc.

Convolutional coding is normally used on random error channels, although there are convolutional codes and decoding algorithms designed for bursty or combined random/bursty channels. Convolutional codes are normally designed for long data streams, but can be adapted for packet-switched applications either by reinitializing the coder and decoder for each packet or by using a *tail-biting* scheme.

When convolutional coding is used, it is easier to implement soft decision detection than in block coded systems. However, in the complicated interference environment anticipated for the HF ITF Network it is difficult to evaluate a maximum likelihood metric, thus degrading the quality of soft decision estimates.

In Appendix F we demonstrate the performance that is achievable by using FH-MFSK signaling in worst case partial band (WCPB) noise jamming with M-ary convolutional coding and one M-ary symbol transmitted per hop. M-ary convolutional coding can actually be considered as a concatenated coding scheme consisting of an M-ary modulator (inner code) and a binary convolutional encoder (outer code). In Section 10.3.3 and Appendix I we demonstrate the performance of Reed-Solomon codes in multiple-user channels for slow FH systems.

The use of concatenated coding with Reed-Solomon outer coding and convolutional inner coding may be well-suited for the HF ITF Network. The Reed-Solomon outer coding would not only take care of burst errors, but also, with high probability, detect uncorrectable packet errors. The inner convolutional code would take care of the random errors caused by atmospheric noise and other disturbances. Alternatively, one could consider a concatenated scheme in which the inner code is nominally designed to provide a BER of 10^{-3}, and the outer code, used only when necessary, reduces the BER to 10^{-5}.

In conclusion, we have noted some of the advantages and disadvantages of convolutional codes and Reed-Solomon block codes. At the present time both Reed-Solomon and convolutional coding schemes are considered as viable alternatives for the HF ITF Network. There is, in fact, little difference in the complexity associated with the implementation of block vs convolutional codes. The final choice of a coding scheme for the HF ITF Network will be made in conjunction with future HFIP and possibly other waveform/coding design studies. In the meantime, we can use the performance results obtained thus far, as presented in this report, as guidelines for anticipated signaling performance.

4.0 NARROWBAND HF LINK COMMUNICATION CONSIDERATIONS

We now consider the problem of HF point-to-point communication in the ITF environment, neglecting until Section 10 the problems associated with networking issues such as link establishment procedures and other-user interference.

Our emphasis in the HF ITF studies has been on wideband signaling, because of its improved AJ and LPI performance. Although presently available equipment uses only narrowband signaling—there have been few changes in existing HF equipment in the past twenty years—we have assumed that a wideband capability will be available when the HF ITF network is implemented. However, since the evolution from narrowband to wideband systems is likely to occur over a period of many years, and since narrowband systems are also needed for communication with nodes that are not part of the U.S. Navy (e.g., analog voice interface with civil or nontreaty organizations) and possibly for operation under benign conditions even when wideband equipment is available, the HF ITF Network must be designed to use both wideband and narrowband communication.

Our intent in this section is not to provide a comprehensive treatment of all narrowband signaling techniques and equipment, but rather to provide a brief overview, as well as a basis for the wideband

signaling discussion of Section 5. The major signaling considerations include the transmission frequency, modulation technique, and AJ performance.

The signal bandwidth of HF systems currently in use is, typically, about 3 to 6 kHz. The choice of transmission frequency is normally based on the desired propagation mode (i.e., groundwave vs skywave) as well as on the operating environment (i.e., atmospheric noise, other-user interference, and jamming). If only groundwave propagation is desired, then a good choice of frequency would be one at which: a strong groundwave signal is propagated; there is no skywave propagation (i.e., the frequency is below the LUF or above the MUF corresponding to the particular communication path), to avoid multipath signals and reduce other-user interference; there is little noise or other-user interference (i.e., the user should look for quiet subbands). Alternatively, if a skywave path is desired, then a frequency near the frequency of optimum transmission (FOT, which has been defined as 85% of the MUF for the desired propagation path), at which multipath dispersion is minimal, would be best if conventional modulation techniques are used. At lower frequencies the skywave itself experiences multipath with propagation via *high* and *low* waves; at slightly higher frequencies than the FOT one would risk losing the channel because of the possibility of fluctuations in the MUF.

A number of modulation schemes that are appropriate for narrowband communication over fading HF channels have been developed. Four such techniques that have been or will be incorporated in a large number of modem equipments are:

a. Advanced Narrowband Digital Voice Terminal (ANDVT) data modem (data rates of 300, 600, 1200, and 2400 b/s).
b. MIL-STD-188C data modem (600, 1200, or 2400 b/s).
c. TADIL-A/LINK-11 data terminal set (1364 or 2250 b/s).
d. AN/UCC-1 Fleet Broadcast modem (16 data streams of 75 b/s each; total rate = 1200 b/s).

The first three of these techniques use multitone differentially coherent phase shift keying (DPSK), while the fourth uses multitone binary frequency shift keying (FSK).

Jewett [26] discusses modulation and coding considerations for the ANDVT voice modem. The problem of intersymbol interference is handled by using sufficiently long guard times (approximately 4 ms). However, fading caused by multipath signals is still a problem.

Other modulation techniques have also been proposed or used briefly for narrowband HF communication. The AN/GSC-10 (KATHRYN) system used coherent phase shift keying (PSK) [27,28]. The ANDEFT/SC-320 system used frequency differential PSK rather than time differential PSK [29]. More recently, a number of serial tone modems with equalizers have been under development.

A detailed discussion of the various modulation schemes that are suitable for the HF channel would be far beyond the scope of this report, especially since the AJ requirements imposed on the HF ITF Network will require the use of wideband signaling. The reader is referred to a survey of modulation techniques [30], and to a discussion of a number of modems [31]. MIL-STD-188C, MIL-STD-188-200, and MIL-STD-188-342 are also of interest.

The jamming protection that can be provided when narrowband signaling is used is not expected to be adequate for the HF ITF Network. It is straightforward for a jammer to detect the presence of a narrowband signal, and then transmit high-energy narrowband jamming signals (either noise or tones) to disrupt communication. In contrast, significantly better AJ performance can be achieved by using wideband signaling.

There are methods, other than coding, that can provide some degree of AJ capability for narrowband signals. For example, if diversity transmission is used (i.e., transmitting the same signal

simultaneously at two or more frequency subbands), the jammer will have to jam this set of subbands. The implementation of such diversity requires the use of additional equipment at both the transmitting and receiving platforms, and as a result bandwidth requirements increase. As in the nondiversity case, the jammer would attempt to determine which subbands are being used and then jam them. In fact it would often be easy for the jammer to do so. However, some significant AJ performance gains may be expected if the jamming signal arrives via skywave and is subject to independent fading in the different subbands.

Another method for achieving protection from jamming is based on the use of adaptive antenna arrays to null out jammers. This approach has the advantage of being compatible with the current narrowband HF communication architecture. Unfortunately, adaptive array techniques have several limitations that argue against exclusive reliance on this method for achieving AJ performance. Most important among these limitations are that only a small number of jammers can be nulled out at the same time with reasonable size arrays, and that it is difficult to achieve protection when the jamming comes from the same direction as the desired signal. Recent simulation results obtained by the Naval Ocean Systems Center (NOSC) indicate that with the combined use of adaptive arrays and adaptive equalization it may be possible to null jammers that are collinear with the communicating platforms when significant multipath propagation modes are present from either the jammer or the communicator [32]. However, if one is operating in a portion of the HF band that does not support skywave propagation, then no such discrimination is possible.

Compton [33] has recently demonstrated that an array that adapts to polarization can null an interfering signal coming from the same direction as the desired signal, if the signals have different polarizations. However, HF groundwave propagation is predominantly vertically polarized, as a result of the more rapid attenuation of horizontally polarized signals in this propagation mode, and hence this technique would appear to be usable only in the unlikely circumstance that the jamming is predominantly horizontal in polarization.

5.0 WIDEBAND HF LINK COMMUNICATION CONSIDERATIONS

We noted earlier that wideband (i.e., spread spectrum) signaling is needed in the HF ITF Network to provide adequate protection from jamming. In this section we summarize the considerations that have led us to recommend frequency hopping (FH) as the spreading mechanism and frequency shift keying (FSK) as the modulation technique. Spread spectrum signaling and modulation in the HF ITF environment are discussed in greater detail in Appendixes C and D, respectively.

We note that the complete design and specification of the wideband HF waveform to be used in the HF ITF Network is not part of this task. Such a detailed waveform design, as well as a hardware (including modem and overall architecture) design, is being undertaken as part of the HF Improvement Program (HFIP) [34,19a,20]. An understanding of waveform design and coding issues is essential, however, to the design of the HF ITF Network, and therefore we have devoted considerable effort to this area. Furthermore, it is hoped that some of the results presented here will be useful to the ultimate HFIP waveform design.

5.1 Spectrum Spreading Technique: Frequency Hopping

One of the fundamental issues in the design of a spread spectrum system is the choice of the spectrum spreading mechanism. The three forms of spread spectrum signaling that we have considered are direct sequence (DS), frequency hopping (FH), and hybrid FH-DS.

The primary criteria by which a spread spectrum system should be evaluated are its ability to provide AJ capability at acceptable data rates, its ability to function in the particular environment for which it has been designed (in our case the HF ITF environment), and the practicality of its implementation.

By considering performance under each criterion we have concluded that, although each of the spreading mechanisms has its advantages, a pure FH system is the most practical choice for use in the HF ITF Network. The basis for this choice is provided in Appendix C.

The main advantages of FH signaling as compared with DS signaling are:

a. FH systems are virtually immume to the *near-far* problem, and are therefore better suited than DS systems for multiple access applications.
b. Pseudonoise (PN) code acquisition (synchronization) is easier to achieve and more difficult to disrupt for FH systems.
c. FH systems are less sensitive to the dispersive nature of HF channels.
d. FH systems do not require a contiguous bandwidth.
e. Wider bandwidths, and therefore greater processing gain can be achieved with FH systems.

Also, when noncoherent signaling is used (e.g., noncoherent FSK, which is proposed for the HF ITF Network) there are no carrier phase acquisition problems.

We have also considered hybrid FH-DS systems. But, although in certain respects they provide some advantages over a pure FH system, they are more difficult to implement.

Thus our conclusion is that a pure FH system is the most appropriate for the HF ITF Network. We note in Appendix A that a hopping rate of 2400 hops per second will considerably reduce the effects of multipath fading caused by interference. A high hopping rate is also advisable to prevent repeat-back jamming, as well as to facilitate the use of coding and diversity. If the hopping rate is limited to at most several hundred hops per second, satisfactory performance is still expected in many cases. However, when operating in frequency bands that support considerable skywave propagation one would have to rely more on coding, interleaving, and modulation techniques that are robust with respect to multipath interference.

The next question is the choice of modulation method. Frequency Shift Keying (FSK) is generally used in pure noncoherent FH systems and is proposed for the HF ITF Network as well. We shall now discuss the reasons for this choice.

5.2 Data Modulation Method: Noncoherent FSK

In Appendix D we discuss issues that relate to the choice of the data modulation method. The primary reasons for our recommendation to use noncoherent FSK are:

a. feasibility of implementation;
b. robustness with respect to channel properties (such as fading and interference); and
c. compatibility with hopping schemes in which either several symbols are transmitted per hop, one symbol is transmitted per hop, or several hops are transmitted per symbol.

We have investigated binary and M-ary FSK. Quaternary FSK (QFSK) is well suited to satisfy the maximum anticipated data rate requirement of 2400 b/s. For example, if a hopping rate of 2400 hops per second is used, then 2400 quaternary symbols per second would be transmitted, thus resulting in a transmitted symbol rate equivalent to 4800 binary symbols per second. If rate 1/2 coding is used, the data rate is the desired 2400 b/s.

In Section 7 and Appendixes E, F, and H we discuss in detail jamming, coding, and diversity for binary, quaternary, and 8-ary FSK. The waveform design must take into consideration the complex interrelationships among many variables, such as the adversary's jamming strategy, channel fading, M-ary signaling, diversity, coding, and interleaving. In the subsequent sections we examine these issues as they relate to the HF ITF Network.

6.0 COMMUNICATION RANGE MODEL FOR A BENIGN ENVIRONMENT

Although HF groundwave propagation loss is fairly well predicted by Barrick's model, as discussed in Section 2, it is still difficult to predict the achievable communication range because of variable (and highly non-Gaussian) interference that arises primarily from atmospheric and galactic noise, other-user interference, and noise generated on naval platforms, as well as uncertainties in system losses that include deviations from nominally omnidirectional antenna patterns. In this section we address the communication range that is achievable in a benign environment.

6.1 Basic Power Budget Considerations

We now summarize a preliminary power budget analysis that illustrates the relationship between transmitted and received energy per bit as a function of frequency, range, and data rate. The bit error rate (BER) performance of a digital system operating in additive white Gaussian noise (AWGN) is a function of E_b/N_0, the ratio between the received energy per bit and the background noise spectral density. Various modulation and coding schemes are normally compared on the basis of their BER performance as a function of E_b/N_0. Because the HF interference environment, which arises primarily from atmospheric noise and other-user interference, is highly non-Gaussian, it is no longer possible to obtain an accurate mathematical characterization for the channel by using a single number, i.e., a value for N_0. Furthermore, locally generated platform noise as well as intermodulation products and other sources of disturbance must be considered as part of the interference process. What can be done is to use a value for N_0 that represents the average noise spectral density, and then estimate the value of E_b/N_0 that is required to achieve a specified BER. One way to do so is to simply use a *representative* value for N_0. Another is to add an appropriate margin to the E_b/N_0 value for AWGN to take care of non-Gaussian effects.

An accurate channel characterization would require a statistical description of the interference processes, including a model for burst vs random nature of the disturbances, as well as a model for fading, if any. The value of E_b/N_0 that is required to achieve a specified BER rests not only on the channel characteristics, but also on the signaling scheme. It depends on the type of modulation that is used. In addition, it depends on the coding scheme, including the ability of the code to cope with burst errors and/or random errors. Interleaving is often used to randomize bursty error processes, and it is especially important when more than one symbol is transmitted per hop, but the degree of interleaving that can be achieved may be limited in packet-oriented systems. Thus, it may be necessary to distinguish between packet-switched and circuit-switched applications when developing a channel characterization. The value of E_b/N_0 required also depends on the use of hard decision vs soft decision detection and the ability, if any, of the receiver to detect and excise disturbances that are not signal-like.

In this section we consider a benign environment in which there is no fading or jamming. Jamming, which is potentially the most disruptive source of interference in the HF ITF Network, is addressed in later sections.

The first step in the determination of link performance is the estimation of the received power of the desired groundwave signal. The received signal power can be expressed in terms of the transmitted power and the gains and losses sustained along the transmission path:

$$P_R = (P_T Q)/L_P,$$

where,

P_R = received signal power,

P_T = transmitter power,

L_P = propagation loss and,

Q = quantity representing net effect of system gains and losses.

Typically, P_T may be about 1 kW, although the new wideband architecture will permit control of transmitter power from much lower levels (desirable for LPI and LRI applications) to as high as 16 kW (which may sometimes be needed to combat jamming) [20]. We assume that all communicating platforms are surface vessels; thus L_P can be estimated from Barrick's curves shown in Fig. 2.1. The effects of rough sea are easily incorporated into the propagation loss model [12]. The parameter Q represents the combined effect of transmitter power distribution losses, receiver processing losses, antenna pattern effects, and possibly other system losses.

We consider the performance of a groundwave link in an unjammed environment. The primary sources of interference are atmospheric noise and other-user interference, which we shall combine into the value N_0 for the noise spectral density. We have noted that the quantity E_b/N_0 is of primary importance in digital systems. It can be expressed in terms of the received signal to noise density ratio, P_R/N_0, as follows:

$$E_b/N_0 = P_R/(RN_0)$$

where R is the information data rate in bits per second.*

Figure 6.1, which illustrates the received energy per bit for a smooth sea and several data rates, assuming binary signaling and a transmitter power of 1 kW (and neglecting transmitter and receiver gains and losses, i.e., Q = 1) was obtained simply by relabeling the vertical axis of Fig. 2.1. The E_b/N_0 ratio (in dB) is then obtained by subtracting N_0 (in dBW-s) from the value of E_b (in dBW-s) shown on the curve, although it is difficult to agree on an appropriate value for N_0, as we have noted. The effects of system gains or losses can be incorporated into the model by simply shifting the vertical axis in accordance with the value of the parameter Q.

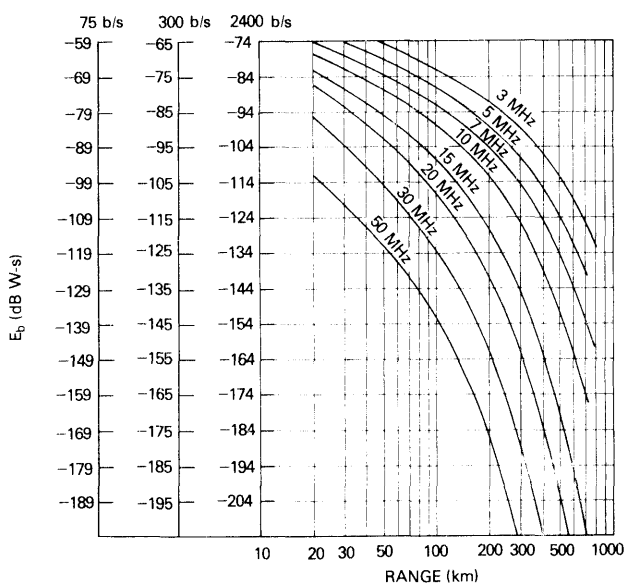


Fig. 6.1 — HF groundwave received energy per bit as a function of frequency, range, and data rate; Transmitter power = 1 kW; binary signaling; omnidirectional antennas; smooth sea

*Note that although the use of coding and/or diversity will increase the number of binary symbols that are transmitted, and the use of M-ary modulation will result in fewer transmitted symbols than in a binary system, R always refers to the actual information data rate in bits per second before the application of modulation, coding, and diversity.

6.2 Narrowband vs Wideband Considerations

Care must be taken to ensure that an acceptable value of E_b/N_0 is maintained over a sufficiently large portion of the hopping bandwidth of a wideband HF channel. It is for this reason that we have divided the HF band into a number of subbands, as we have discussed in Section 1. Over each of these subbands the propagation loss, and therefore connectivity, is relatively constant. However, the background interference levels, and therefore E_b/N_0, will usually vary over the frequency bins that comprise each subband.

It is sometimes argued that greater communication ranges can be achieved in benign environments by using narrowband signaling rather than wideband signaling. The basis for this argument is that one can find a quiet channel to operate in, which may be characterized by noise levels tens of dB less than the average noise density over that portion of the HF band. Such reasoning is valid only if interference levels remain relatively constant. A sudden increase in noise level in a narrowband channel can occur when another (not necessarily hostile) user begins to transmit in the channel, or when skywave propagation conditions change, thus supporting additional sources of interference. Communication can therefore be easily disrupted. Considerable delay may then result in finding another quiet channel.

A FH signal, on the other hand, will hop over the entire subband, and will be able to overcome the effects of a number of noisy frequency bins (and hence lost bits) through the use of forward error correction coding, provided that an adequate E_b/N_0 is maintained in a sufficient number of frequency bins. If for example, Reed-Solomon codes of rate 1/2 are used, then a correct codeword decision can be made despite errors in 25% of the hops, and virtually all codeword errors are detectable, even if they are not correctable, as discussed in Appendix B. If one can decide which frequency bins are unreliable, then correct codeword decision can be made despite the erasure of 50% of the hops. The code rate can be lowered if necessary, or diversity can be used. In some cases it may be possible to adjust the FH patterns to avoid the worst of the slots. The use of Reed-Solomon codes in a multiuser environment is discussed in Section 10.3.3 and Appendix I. The use of frequency hopping with coding and/or diversity provides robustness against frequency dependent interference because symbols corresponding to each information bit are transmitted in more than one frequency bin. The improved performance of FH systems is especially apparent when jammed environments are considered, as is discussed in Sections 7 through 9, where convolutional coding is considered.

Significant losses can arise as a result of nulls in the transmitting and receiving antenna patterns. Although HF shipboard antennas are designed to be nominally omnidirectional, antenna patterns often have significant nulls, as great as 20 dB or more, as a result of interaction with the structure of the ship. Typically there are a few broad nulls at the lower end of the band, and an increasing number of narrower nulls as frequency increases.

We first consider the effect of such nulls on narrowband systems. Results from Smyth [35] cited in Dobson [36] indicate that if 95% azimuth coverage is required, a margin of about 13 dB should be added to alleviate the effects of the nulls. An alternative solution might be simply to change the operating frequency, because antenna patterns can change significantly at HF if the operating frequency is varied by as little as a few hundred kHz. It may be difficult to do so, however, because some scheme is needed to coordinate the choice of frequency at the source and destination platforms. We note that the existence of nulls can result in link outages as transmitting and receiving platforms change orientation with respect to each other.

In the case of broadcast mode (i.e., multidestination) communication, it is desired that all platforms within a nominally circular region receive the transmissions of a single platform. Use of a single narrowband frequency channel will often result in gaps in coverage, particularly at the upper end of the

HF band. Such gaps cannot be eliminated by simply changing to another transmission frequency, because then they will simply reappear elsewhere.

The use of adaptive arrays in narrowband systems can reduce the depth of nulls, thereby making antenna patterns more nearly omnidirectional. Such systems, however, require the use of a number of antennas, which may or may not be available for this purpose.

We now turn our attention to FH systems. We noted earlier that the use of a single narrowband frequency leaves one vulnerable to the severe attenuation experienced in antenna nulls. If, however, one hops over a wide bandwidth (e.g., 1 MHz or greater), only a relatively small fraction of the hops will normally be located in nulls, and thus no spatial direction would experience, on the average, significantly greater losses than any other. The use of frequency hopping, thus, again provides some degree of robustness with respect to frequency dependent system losses by providing frequency diversity.

We note that frequency hopping can be used in conjunction with adaptive arrays, thereby providing many of the advantages of both of these techniques. In such a hybrid system it would be necessary to generate a different set of weights for each frequency bin as a result of the dependence of the antenna pattern on frequency at HF. This need to adapt at each frequency bin may, depending upon the ability of the system to change weights rapidly, result in delay at each hop during which no data is transmitted. As a result, the hopping rate may be constrained to be lower in a hybrid FH adaptive array system than it might be in a pure FH system. The data rate could thus be considerably lower than in a pure FH system as a result of decreased time efficiency, unless one hopped sufficiently slowly, in which case many of the benefits of frequency hopping might be lost.

6.3 A Range Model

We now present a sample power budget calculation in which we assume a transmitted power level of 1 kW, path loss corresponding to sea state 6 (extremely rough sea), data rates of 75 and 2400 b/s, noncoherent binary FSK, rate 1/2 convolutional coding with constraint length 7, a soft decision receiver structure, interleaving to combat burst errors, and BER = 10^{-5}.

In this model we assume that the only losses are propagation loss and the quantity Q representing *miscellaneous system losses.* The former are shown in Fig. 6.2, which has been obtained from Fig. 2.1 by incorporating the additional loss corresponding to sea state 6 (from [12]), and the latter are assumed to be 3 dB. The same range model is used for both narrowband and wideband signaling. We do not consider interference caused by other network users to be part of the background noise level. For the case of narrowband signaling, we implicitly assume that network management schemes ensure that neighboring users do not transmit in the same frequency bin simultaneously.

The signal design parameters, sea state, BER, and system losses have been defined earlier in this section. For rate 1/2 convolutional coding with constraint length 7, the required E_b/N_o in a Gaussian noise environment is 9.7 dB (see e.g., [37]), which we round up to 10 dB. We have chosen the noise density N_0 based on one of the noise occupancy distributions presented in Brown [24]. The communication range model obtained is illustrated in Fig. 6.3 as a function of frequency for data rates of 75 and 2400 b/s. This model should be considered as being representative of a region characterized by relatively high noise levels, rather than as a definitive or worst-case model for HF groundwave propagation over seawater. Noise levels in quiet areas are typically tens of dB lower than assumed here, resulting in increased communication range.

The communication range that is achievable in a multiple user environment, in which users have to contend with interference from other task force members in addition to the disturbances discussed here, is a topic for further study. In Section 10.3.3 and in Appendix I we consider the effect of a

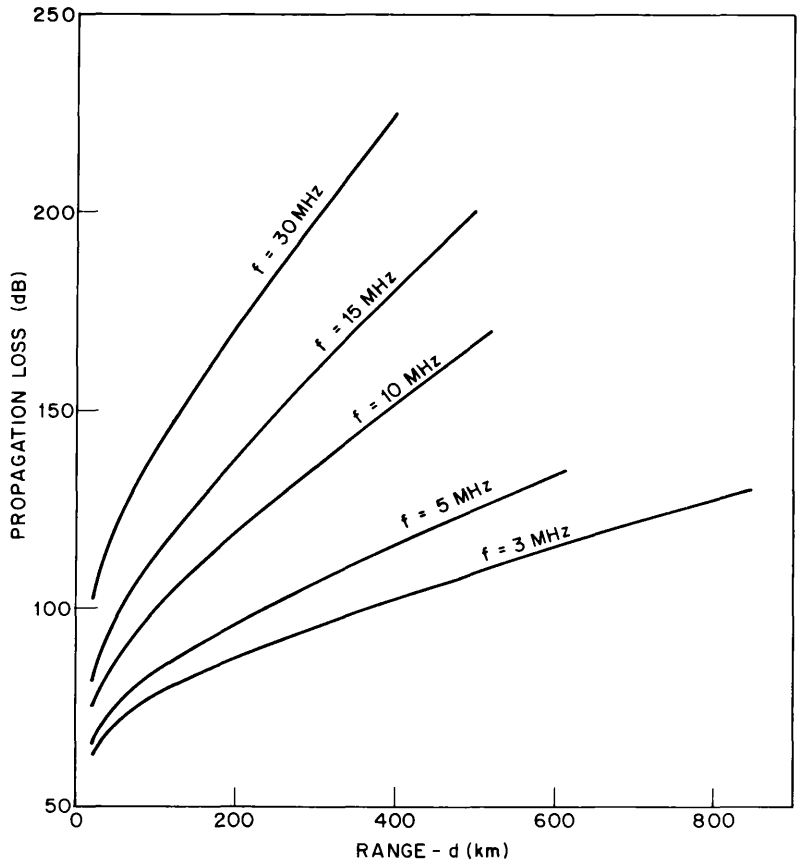
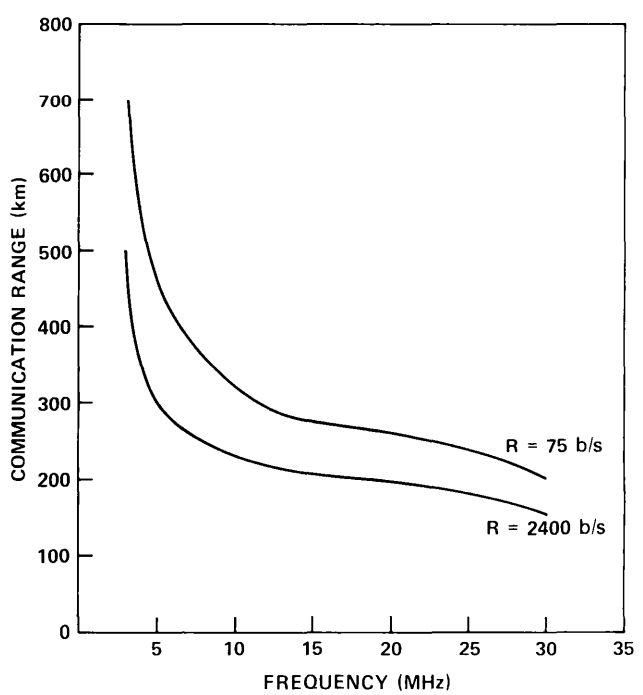


Fig. 6.2 — Groundwave path loss over a rough sea (sea state 6) as a function of distance for several frequencies in the HF band

Fig. 6.3 — HF groundwave communication range for our model



number of other users on correct packet reception. In the future, this approach will be combined with a model for HF propagation and disturbance characteristics.

Estimates of communication ranges that are achievable in a benign environment are of limited use to the HF ITF Network, which must be designed to operate under severe stress. These estimates, however, are useful in assessing the impact jamming has on link performance. The primary limitation on achievable communication range will be caused by hostile jamming, although fading is also an important concern. Therefore, it is necessary first to develop an understanding of how jamming affects communication link performance, and then to develop modulation, coding and diversity techniques, including methods to adaptively vary data rates, that are suitable for operation in a jammed environment.

7.0 AJ PERFORMANCE OF FH-MFSK SIGNALING OVER NONFADING AND RAYLEIGH FADING CHANNELS

In this section we illustrate the AJ link performance that can be obtained in a FH-MFSK system, using signaling parameters that are characteristic of the HF ITF environment. We begin in Section 7.1 with a discussion of basic AJ signaling considerations, and in Section 7.2 we present a summary of AJ performance tradeoffs. These results are based on three appendixes. In Appendix E we discuss the performance of an uncoded system. In Appendix F we demonstrate the improved performance that can be achieved through the use of coding and diversity. These two appendixes deal with soft decision receivers and known jammer state, as defined below. In Appendix H we demonstrate the procedure used to obtain performance results for hard decision receivers.

Again, the quantity E_b/N_0 is of fundamental importance in the jamming environment, as it was in the benign environment. However, it must be redefined in terms of an equivalent energy-per-bit to noise density ratio. Initially, AJ performance will be expressed in terms of BER as a function of an equivalent bit energy to jammer noise density ratio, and then in terms of the jammer to signal power ratio that can be tolerated at the receiver. The performance results that we present here are based on Omura's model [38], which we have applied to the HF ITF Network. We consider the following cases:

a. Broadband noise jamming and worst case partial band (WCPB) noise jamming,
b. several alphabet sizes (i.e., values of M in MFSK),
c. many values of diversity,
d. several coding schemes,
e. both hard and soft decision receivers,
f. jammer state known or unknown (i.e., in the former case the receiver knows whether or not each received symbol has been jammed), and
g. nonfading and Rayleigh fading channels.

7.1 Basic AJ Signaling Considerations

An AJ communication system must provide resistance against jamming signals that have substantially more power than the desired signals. In a FH system, AJ capability is achieved by pseudorandom hopping of the instantaneously narrowband signal across a wide frequency band, and by use of coding and diversity. The jammer may choose to jam across the entire frequency band, resulting in little power in each frequency bin (broadband jamming), or it may choose to concentrate the same total power over a smaller fraction of the band (partial-band jamming), thereby placing more jamming power into each of the jammed frequency bins. In this report we consider noise jamming for both broadband and worst case partial band (WCPB) cases. In a WCPB jamming strategy, the jammer chooses the fraction of bins that are jammed to maximize the resulting BER. Partial band jamming is usually preferable, although Omura [38] and Avidor [39] have recently demonstrated that broadband noise jamming

is, in fact, the WCPB noise jammer in uniform Rayleigh fading channels, i.e., channels in which propagation conditions are uniform across the entire channel bandwidth. Multitone jamming is also a realistic threat; its impact on link performance is a subject for future investigation. The use of multitone jamming, however, is not expected to result in significantly different communication ranges than noise jamming.

Further protection from jamming in networks such as ours is achieved by using relaying and adaptive routing, as was recently demonstrated by Baker et al [11,40].

We assume that noncoherent MFSK signaling is used, with one MFSK tone transmitted per hop. The basic signal parameters are:

$M = 2^K$ = alphabet size

K = number of binary symbols per MFSK symbol

$= \log_2 M$

R_h = hopping rate (hops per second)

We note here that the bandwidth of each frequency bin is MR_h Hz (assuming orthogonal tone spacing, and neglecting the need for guard bands to reduce adjacent channel interference), and that $K = \log_2 M$ binary symbols are transmitted per hop.

We assume that the hopping rate is sufficiently fast so that repeater jamming is not possible. We also assume that the frequency hopping patterns are pseudorandom in nature, and therefore cannot be predicted by the jammer.

The basic AJ system parameters are as follows:

W = total spread spectrum signal bandwidth (number of frequency bins times their individual bandwidth),

R = information data rate in bits per second,*

S = received signal power, and

J = received jammer power summed over all frequency bins over which the signal is hopping.

The received energy per information bit is:

$$E_b = S/R,$$

and the average noise spectral density is:

$$N_0 = J/W.$$

The *equivalent bit energy to noise density ratio* is defined as:

$$E_b/N_o = \frac{S/R}{J/W} = \left(\frac{W}{R}\right)\left(\frac{S}{J}\right) = \frac{PG}{J/S},$$

where $PG = W/R$ is the *processing gain,* and J/S is the received jammer-to-signal power ratio.

*We again note that R always refers to the actual information data rate in bits per second before the application of modulation, coding, and diversity.

This formulation is valid for a wide variety of signal and jammer waveforms, although system performance would of course depend on the specific waveforms that are used. The formulation is useful because it allows the comparison of different types of AJ systems on the basis of the E_b/N_0 required for a given BER. Note that E_b refers to the received energy per information bit rather than per M-ary chip or per binary channel symbol. In all of our discussions on AJ performance the term E_b/N_0 refers to the equivalent E_b/N_0. For the purpose of this analysis, noise sources other than jamming have been neglected.

Care must be used in the interpretation of the term *processing gain* in conjunction with FH systems. A quantitative definition of processing gain such as that used here is valid only if the signal is hopped uniformly over the bandwidth W, and the jammer is incapable of predicting the hopping pattern or of following the signal as it hops from one frequency bin to another. It is assumed that the jammer knows everything about the signal characteristics (e.g., the hopping bandwidth, modulation/coding/diversity scheme, and received signal levels) except the actual FH pattern. For example, the definition we use is valid for the broadband and partial band jamming examples that are considered in this report. Although other definitions of processing gain are, in fact, possible (e.g., it is sometimes defined as the number of frequency bins over which the signal is hopped), the definition we have used is consistent with the problem formulation based on the equivalent E_b/N_0, and it facilitates the comparison of a number of signaling schemes on the basis of their AJ performance, as we demonstrate in Appendix G.

We compare different signaling schemes on the basis of their *tolerable J/S ratio,* i.e., the maximum J/S ratio that can be tolerated by the AJ communication system for a specified BER:

$$J/S|_{\max}\ (\text{dB}) = PG\ (\text{dB}) - E_b/N_0\ (\text{dB}),$$

where the value of E_b/N_0 used in this equation is that which is required to obtain the desired BER and is dependent on the signaling scheme that is used. Clearly, AJ performance can be improved by increasing the processing gain, which can be accomplished either by increasing the hopping bandwidth (and thus the number of frequency bins), or by lowering the data rate. As an example, for a total spread bandwidth (which does not have to be contiguous for FH systems) of 5 MHz and a data rate of 2400 b/s we have $PG =$ 33.2 dB. Lowering the information data rate to 75 b/s increases the processing gain by a factor of 32, or 15 dB, to 48.2 dB. In a practical system the data rate, although not necessarily the hopping rate, may have to be lowered during periods of severe jamming. Although the values of processing gain that are achievable at HF are considerably smaller than those at UHF, SHF, or EHF (because of the wider bandwidths achievable in those frequency bands), AJ performance can be considerably better than it is in the narrowband signaling case.

The overall AJ performance of a communication system depends not only on $J/S|_{\max}$, but also on the actual received signal and jammer powers, as well as on various system losses. Our studies have emphasized the evaluation of $J/S|_{\max}$ because this quantity reflects the dependence of system performance on the modulation/coding/diversity scheme. Clearly, BER performance can also be improved by increasing the received signal power, but such an option is often not available to the communicator, and if it were it could be exercised regardless of the signaling scheme that is used. The signal and jammer power levels will enter into our analysis in Section 9, where we present a communication range model for a jammed environment.

The AJ performance curves that are presented in Section 7.2 have been generated under the assumption of an *ideal** soft decision receiver that can detect which of the symbols it receives have been corrupted by jamming; the same assumption has been made by a number of other researchers in

*We sometimes refer to the soft decision receiver with *J.S.I.* as the *ideal* receiver, because its performance is better than that of other receiver structures. However, we acknowledge that the use of this term is imprecise; e.g., a hard decision receiver with no implementation losses is also, in some sense, ideal.

this area (e.g., [41]). The knowledge of whether or not particular symbols are jammed, a form of *side information,* is referred to as *Jammer State Information,* or *J.S.I.* It is expected that the performance of practical receiver structures can approach that of such an ideal receiver, although some small processing losses are expected. The performance of soft decision receivers without *J.S.I.* is poor, as discussed in Appendix H, and is not considered. In Section 7.3 we discuss the performance that is achievable using hard decision receivers, for both cases in which *J.S.I.* either is or is not available.

We have considered both nonfading and Rayleigh fading channels. We noted earlier that HF groundwaves experience little dispersion. The main source of dispersion in groundwave links is skywave multipath interference, much of which can be avoided at ITF ranges by hopping at 2400 hops per second or faster. Even at lower hopping rates, the effects of skywave interference can be minimized through the use of noncoherent FSK signaling, with one (possibly M-ary) symbol transmitted per hop (or, alternatively, several symbols transmitted in parallel in each hop). Since we will be relying primarily on groundwave links in the HF ITF Network, a nonfading channel model is adequate to evaluate AJ performance for most cases of interest if a hopping rate of 2400 hops per second or greater is in fact used, and is expected to provide reasonably accurate results even for lower hopping rates. For frequencies at which skywave propagation is not supported, the nonfading model would be valid at all hopping rates.

Since some skywave links may be used either within the network or for communication to external points, an understanding of the AJ performance over fading channels is also essential. A fading model may also be appropriate at low hopping rates when operating at frequencies at which skywave propagation is supported. Although a Rician (i.e., specular-plus-Rayleigh) fading model might be more appropriate than a Rayleigh model in some cases (where the groundwave signal is the specular component and the skywave signal is the Rayleigh component), a Rayleigh model is useful because it provides a worst-case bound for Rician channel performance.

7.2 AJ Performance Tradeoffs for Soft Decision Receivers

Before summarizing our major results, it is useful first to review the many variables affecting AJ performance.

The tolerable J/S ratio may be expressed as:

$$J/S|_{\max} = f(M, m, r, W, R, BER)$$

where,

M = M-ary alphabet size,

m = diversity per coded MFSK symbol,*

r = composite code rate (defined in Section B.4.1).

W = spread hopping bandwidth, and

R = information data rate.

The resultant bit error rate can be expressed as,

$$\text{BER} = f(E_b/N_0 | \text{ detection technique channel model, decision strategy, J.S.I.})$$

*m is the number of times that each M-ary symbol produced by the encoder is redundantly transmitted. The reader is reminded that two diversity parameters (m and L — see section E.1.1) and three code rate parameters (r, r_i, and r_0 — see Section B.4.1) are used in this report.

Our analyses assume the use of noncoherent detection, with and without *J.S.I.*, and apply to WCPB noise jamming on both nonfading and Rayleigh fading channels. The results for the fading channel apply also to the nonjammed case in which the Rayleigh fading channel is affected only by AWGN, since WCPB jamming in a Rayleigh fading channel is actually broadband jamming. Both soft and hard decision receiver structures are a part of this analysis.

To present some AJ tradeoffs we specify the following variables as an example.

W = 5 MHz,
R = 2400 b/s,
Soft decision receiver,
Jammer State Information *(J.S.I.)* known,
Worst case partial band (WCPB) noise jamming,
BER = 10^{-5} and 10^{-3}.

The two values of BER listed above correspond to high-quality and medium-quality data reception, respectively. The AJ performance under these two criteria differs somewhat, and it is discussed separately in the following subsections.

7.2.1 AJ Performance Under a High-Quality Data Reception Criterion (BER = 10^{-5})

Figure 7.1 shows the AJ performance of a FH-MFSK signaling design under the conditions described above, and using the high-quality data reception criterion. Results for nonfading as well as Rayleigh fading channels are shown. In either case, it is apparent that some form of coding and/or diversity is needed to make the system capable of functioning in a jammed environment. The coding and diversity improvement gains are represented by the difference between one of the upper curves and the lowermost curve in this figure. Clearly, coding provides more performance improvement than does diversity, as expected. However, this advantage is most apparent at the lower alphabet sizes, and it diminishes somewhat as M increases. The curve illustrating the performance of a system with coding and optimum diversity represents the best operational AJ performance that is possible. It is within 3 dB of the maximum achievable value based on the R_0 bound. Note that optimum diversity refers to the diversity value m that results in the lowest E_b/N_0 value.

The curves for *coding only* and *coding and optimum diversity* shown in Fig. 7.1 are based on the use of convolutional coding of constraint length 7. For binary signaling we assume the use of Odenwalder's optimum rate 1/2 code [42]. For M-ary signaling we assume the use of Trumpis's composite rate $r = 1$ M-ary codes [43], which correspond to an equivalent binary code rate r_0 of 1/2 and 1/3 for $M = 4$ and 8, respectively (see Table B1 in Appendix B).

Based on the results summarized in Fig. 7.1, either the $M = 4$ or $M = 8$ signaling designs (with rate $r = 1$ coding and using diversity) represent reasonable system choices. It must be noted that the use of coding and diversity, for fixed data rate R, results in faster hopping rates and therefore greater frequency bin bandwidth. In a practical system, additional diversity normally is achieved by lowering the data rate while maintaining a fixed hopping rate. Thus, there is a tradeoff between data rate and BER performance, and the use of optimum diversity values may in some cases result in unacceptably low data rates.

Table 7.1 summarizes the AJ performance for binary, 4-ary, and 8-ary alphabets as a function of diversity for a fixed data rate of $R = 2400$ b/s and a hopping bandwidth of 5 MHz. The hopping rate (R_h) and frequency bin bandwidth (BW) are included in the table to illustrate the impact of the use of diversity on signal design parameters. In the nonfading case the optimum diversity is $m = 2$ for the desired BER = 10^{-5}. The use of $m = 2$ requires a halving of the data rate if the hopping rate is to

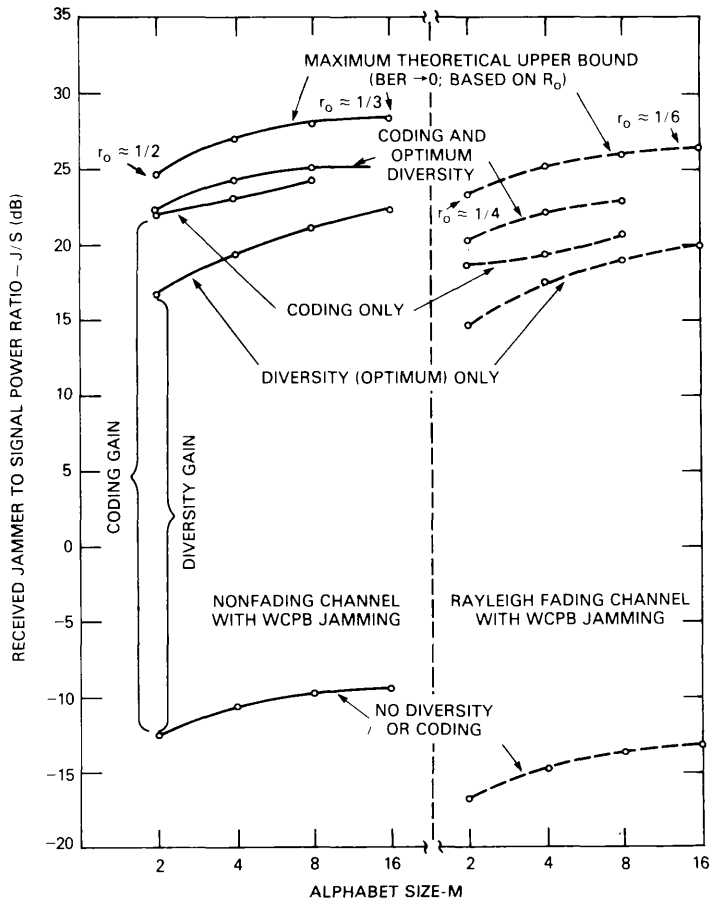


Fig. 7.1 — FH-MFSK AJ signaling tradeoffs for WCPB noise jamming on nonfading and Rayleigh fading channels under a high-quality data reception criterion (BER = 10^{-5}); R = 2400 b/s; W = 5 MHz; soft decision receiver; J.S.I. known

Table 7.1 — Tolerable J/S Protection Ratios for Noncoherent FH-MFSK; Data Rate Fixed at 2400 b/s; High-Quality Data Reception (BER = 10^{-5}). (Hopping Bandwidth = W = 5 MHz; Convolutional Code of Constraint Length 7; Soft Decision Receiver; J.S.I. Known)

M	r_0	r	m	R_h	BW (Hz)	$J/S\|_{max}$(dB) WCPB Noise Jamming	$J/S\|_{max}$(dB) Rayleigh Fading
2	1/2	1/2	1	4800	9600	22.0	18.6
			2	9600	19200	22.3*	20.0
			3	14400	28800	21.7	20.2*
4	1/2	1	1	2400	9600	23.1	19.2
			2	4800	19200	24.3*	21.4
			3	7200	28800	24.0	21.9
			4	9600	38400	23.8	22.0*
8	1/3	1	1	2400	19200	24.2	20.5
			2	4800	38400	24.9*	22.3
			3	7200	57600	24.4	22.7*

Note: *indicates optimum value of diversity (m)

remain fixed, as we have just discussed. Therefore, it may be better to use a no-diversity ($m = 1$) system to avoid the data rate reduction, although such a system would require slightly higher E_b/N_0, i.e., an additional (0.3, 1.2, 0.7) dB for $M = (2, 4, 8)$. This tradeoff between data rate and E_b/N_0 will have to be evaluated for each specific application.

In the Rayleigh fading case the optimum diversity is 3 or 4, but little improvement is gained by increasing m to values greater than 2. The use of diversity is more important on Rayleigh fading channels than it is on nonfading channels, because the increase in E_b/N_0 values required in a no-diversity system as compared to $m = 2$ is (1.4, 2.2, 1.8) dB for $M = (2, 4, 8)$.

All of these results have dealt with a 2400 b/s data rate. The AJ performance for other data rates and spread hopping bandwidths can be derived easily from these results by use of the relationship:

$$J/S|_{\max} = (W/R)/(E_b/N_0).$$

Once a particular modulation/coding/diversity scheme and BER requirement have been chosen, the necessary E_b/N_0 value may be determined from one of the curves presented in Appendixes E and F for the uncoded and coded cases, respectively. If the spread hopping bandwidth W is fixed, then $J/S|_{\max}$ is inversely proportional to the data rate R. For example, if R and R_h are each lowered by a factor of 2 (thereby preserving the diversity value m), then the tolerable J/S ratio is increased by 3 dB. Performance for other values of m can be derived from the curves shown in Appendix F. If we allow the data rate to vary from 75 b/s to 2400 b/s, the tolerable J/S is shown in Fig. 7.2; e.g., by lowering the data rate from 2400 b/s to 75 b/s we gain 15 dB of J/S protection. It will be necessary to develop schemes that are capable of responding to channel disturbances by lowering data rates when necessary, and restoring them to higher levels whenever possible. To summarize, it is theoretically possible, when noncoherent FH 8-ary FSK signaling is used with rate $r_0 = 1/3$ convolutional coding and optimal diversity, to attain the following protection ratios for data rates between 75 b/s and 2400 b/s:

	2400 b/s		75 b/s	
•	24.9 dB	$\leqslant J/S\|_{\max} \leqslant$	39.9 dB	(WCPB noise jamming; No fading; $m = 2$)
•	22.7 dB	$\leqslant J/S\|_{\max} \leqslant$	37.7 dB	(Rayleigh fading; Broadband* jamming; $m = 3$).

Some of the hopping rates and bandwidths shown in Table 7.1 may be too high to be implemented in a practical system, particularly at HF where the number of frequency bins is severely limited. Table 7.2 summarizes the AJ performance as a function of diversity for a fixed frequency bin bandwidth of 9600 Hz, and therefore a fixed hopping rate for each alphabet size, M. The diversity parameter m is inversely proportional to the data rate for the present case of a fixed hopping rate. Tradeoffs between data rate and J/S performance can easily be determined by using this table.

7.2.2 AJ Performance Under a Medium-Quality Data Reception Criterion (BER = 10^{-3})

Certain types of information transmitted within the ITF Network will not necessarily require the stringent BER performance (10^{-5}) discussed in the previous subsection. AJ performance results based on a medium quality (BER = 10^{-3}) data reception criterion are shown in Fig. 7.3 for nonfading and Rayleigh fading channels. The interesting distinction in this case is that some minimal protection from jamming (albeit very small) is possible without using coding or diversity. For any practical implementation, however, coding and/or diversity are recommended because of the enhanced jamming protection ratios they provide, as shown in Fig. 7.3. The coding and diversity improvement gains are represented by the difference between one of the upper curves and the lowermost curve in the figure. Clearly,

*Broadband noise jamming is WCPB noise jamming for the case of the uniform Rayleigh fading channel.

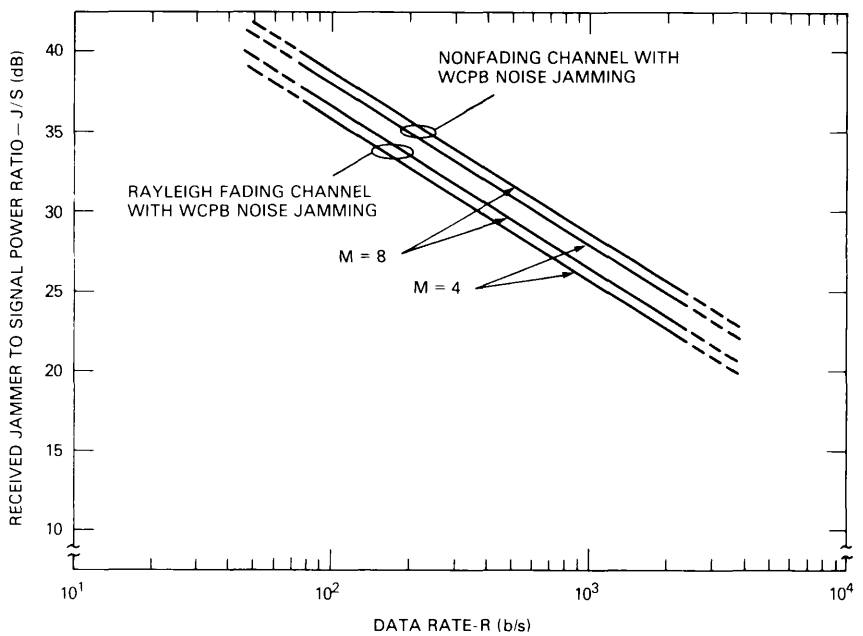


Fig. 7.2 — Maximum J/S protection ratio under a high-quality data reception criterion (BER = 10^{-5}) as a function of data rate. Rate r = 1 coding and optimum diversity; W = 5 MHz; soft decision receiver; J.S.I. known

Table 7.2 — Tolerable J/S Protection Ratios for Noncoherent FH-MFSK; Frequency Bin Bandwidth Fixed at 9600 Hz; High-Quality Data Reception (BER = 10^{-5}). (Hopping Bandwidth = W = 5 MHz; Convolutional Code of Constraint Length 7; Soft Decision Receiver; J.S.I. Known)

M	r_0	r	R_h	m	$R\,(b/s)$	$J/S\|_{max}$ (dB) WCPB Noise Jamming	$J/S\|_{max}$ (dB) Rayleigh Fading
2	1/2	1/2	4800	1	2400	22.0	18.6
				2	1200	25.3*	23.0
				3	800	26.5	25.0*
4	1/2	1	2400	1	2400	23.1	19.2
				2	1200	27.3*	24.4
				3	800	28.8	26.7
				4	600	29.8	28.0*
8	1/3	1	1200	1	1200	27.2	23.5
				2	600	30.9*	28.3
				3	400	32.2	30.5*

Note: *indicates optimum value of diversity (m)

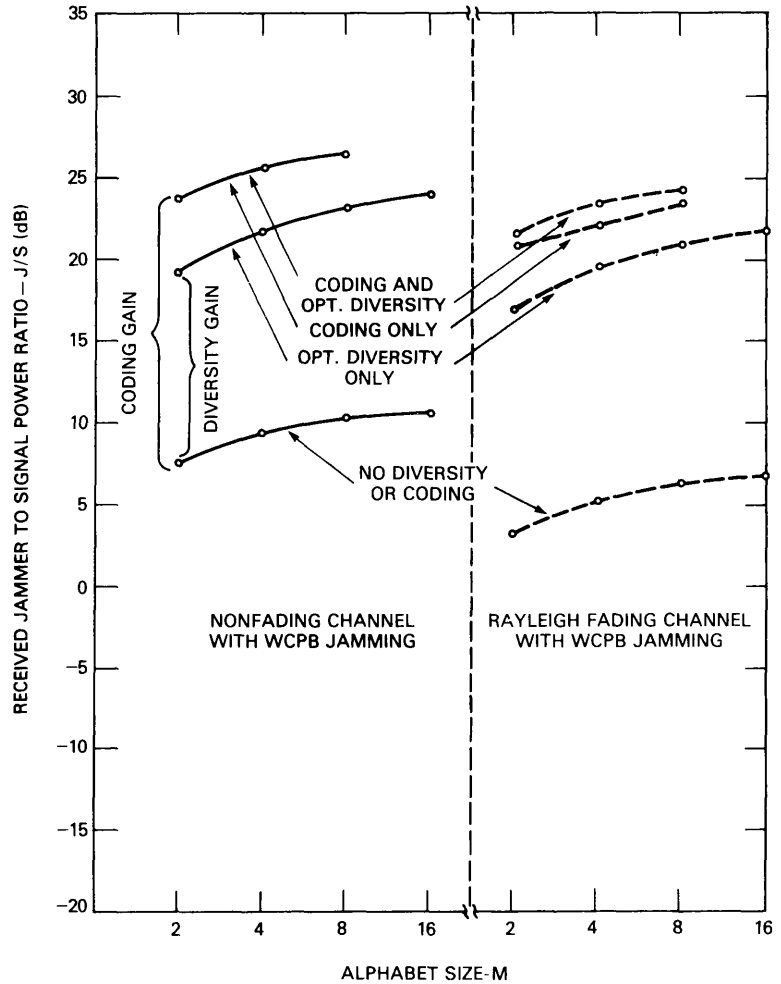


Fig. 7.3 — FH-MFSK AJ signaling tradeoffs for WCPB noise jamming on nonfading and Rayleigh fading channels under a medium-quality data reception criterion (BER = 10^{-3}); R = 2400 b/s; W = 5 MHz; soft decision receiver; J.S.I. known

these coding/diversity gains become smaller when the BER requirement is increased from 10^{-5} to 10^{-3}. This decrease in gain is evident by comparing Figs. 7.1 and 7.3 which indicate *coding only* gains of about 34 to 35 dB for BER = 10^{-5} and only 16 to 18 dB for BER 10^{-3}. Table 7.3 summarizes the AJ performance for binary, 4-ary, and 8-ary alphabets as a function of diversity for medium quality data reception, and a data rate fixed at R = 2400 b/s.

This table indicates that in a coded system the use of diversity results in no performance enhancement on the WCPB noise jammed nonfading channel, and only modest improvement (about 1 dB) on the Rayleigh fading channel. This result tends to reinforce our earlier statement in which we indicated that coding is more effective than diversity in combating WCPB noise jamming on nonfading as well as fading channels.

As in the case of high-quality data reception, these results have dealt with a 2400 b/s data rate. Figure 7.4 illustrates the maximum tolerable AJ protection ratios as a function of data rate. To summarize, it is theoretically possible, when noncoherent FH 8-ary FSK signaling is used with rate $r_0 = 1/3$

Table 7.3 — Tolerable J/S Protection Ratios for Noncoherent FH-MFSK;
Data Rate Fixed at R = 2400 b/s;
Medium-Quality Data Reception (BER = 10^{-3}).
(Hopping Bandwidth = W = 5 MHz;
Convolutional Code of Constraint Length 7;
Soft Decision Receiver; J.S.I. Known)

M	r_0	r	m	R_h	BW (Hz)	$J/S\|_{max}$ (dB) WCPB Noise Jamming	$J/S\|_{max}$ (dB) Rayleigh Fading
2	1/2	1/2	1	4800	9600	23.8*	20.8
			2	9600	19200	23.3	21.6*
4	1/2	1	1	2400	9600	25.5*	22.0
			2	4800	19200	25.5	23.3
			3	7200	28800	24.9	23.5*
8	1/3	1	1	2400	19200	26.6*	23.5
			2	4800	38400	26.1	24.2*

Note: *indicates optimum value of diversity (m)

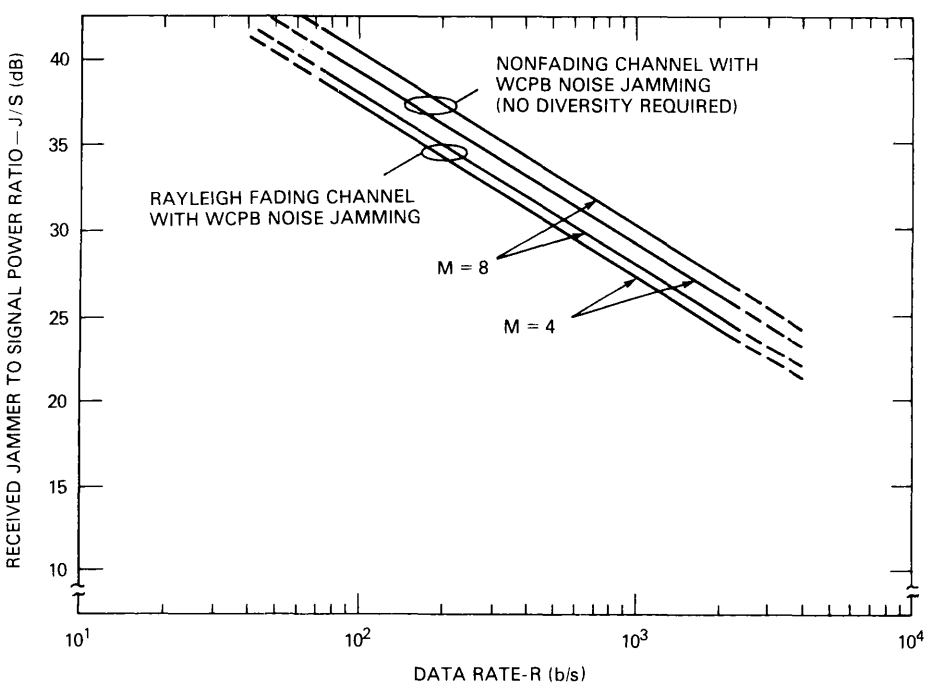


Fig. 7.4 — Maximum J/S protection ratio under a medium-quality data reception criterion (BER = 10^{-3}) as a function of data rate. Rate $r = 1$ coding and optimum diversity; $W = 5$ MHz; soft decision receiver; J.S.I. known

convolutional coding and optimal diversity, to attain the following protection ratios for data rate between 75 b/s and 2400 b/s.

2400 b/s 75 b/s

- 26.6 dB $\leqslant J/S|_{max} \leqslant$ 41.6 dB (WCPB noise jamming; No fading; No diversity required)
- 24.2 dB $\leqslant J/S|_{max} \leqslant$ 39.2 dB (Rayleigh fading; Broadband jamming; $m=2$)

These protection ratios are only about 1.5 dB greater than those achievable for the BER = 10^{-5} case illustrated earlier for coded/optimum diversity FH-MFSK signal designs. Hence, allowing a higher BER does not permit the system to withstand a significantly greater jamming threat for this case of FH-MFSK signaling with coding and optimum diversity. For a signal design without either coding or diversity, however, the opposite is true. In this case, for BER = 10^{-5} the system is not jam-resistant, but by increasing the acceptable BER to 10^{-3} the system will achieve some minimal protection from jamming (see Figs. 7.1 and 7.3).

Table 7.4 illustrates the tradeoffs between data rate and AJ performance when the frequency bin bandwidth is fixed at 9600 Hz.

Table 7.4 — Tolerable J/S Protection Ratios for Noncoherent FH-MFSK; Frequency Bin Bandwidth Fixed at 9600 Hz; Medium-Quality Data Reception (BER 10^{-3}). (Hopping Bandwidth = W = 5 MHz; Convolutional Code of Constraint Length 7; Soft Decision Receiver; J.S.I. Known)

M	r_0	r	R_h	m	R(b/s)	$J/S\|_{max}$ (dB) WCPB Noise Jamming	$J/S\|_{max}$ (dB) Rayleigh Fading
2	1/2	1/2	4800	1	2400	23.8*	20.8
				2	1200	26.3	24.6*
4	1/2	1	2400	1	2400	25.5*	22.0
				2	1200	28.5	26.3
				3	800	29.7	28.3*
8	1/3	1	1200	1	1200	29.6*	26.5
				2	600	32.1	30.2*

Note: *indicates optimum value of diversity (m)

7.2.3 Remarks on the Performance of a Soft Decision Receiver with J.S.I.

Based on the results presented in this section, we can make the following observations:

Without coding or diversity:

- A FH-MFSK design provides no protection from jamming when high-quality data reception (BER = 10^{-5}) is required, and only minimal protection for medium-quality data reception (BER = 10^{-3}).

With coding* and/or diversity:

- The convolutional coding gain alone is approximately 34 to 35 dB for high-quality data reception; for medium-quality data reception the coding gain decreases to about 16 to 18 dB.
- Convolutional coding alone is always better (by about 2 to 5 dB) than diversity alone.
- When convolutional coding is used, adding diversity improves AJ performance only slightly (1 to 3 dB for the Rayleigh fading channel, and 0 to 1 dB for the nonfading WCPB noise jammed channel).
- When convolutional coding and optimum diversity are used, the AJ performance improves only slightly (about 1.5 dB) by lowering the BER requirement from 10^{-5} to 10^{-3}.

Conclusion:

- Coding and/or diversity are essential to a jam-resistant HF ITF FH-MFSK signal design.

7.3 AJ Performance Tradeoffs for Hard Decision Receivers

The AJ performance results presented in the preceding subsections are based on the use of an ideal soft decision receiver in which jammer state information is available, and therefore represent optimistic upper bounds on jam-resistant performance. Although we expect the performance of practical receivers to approach that of the ideal receiver, in practice there will be some degradation, e.g., that which results from the inability to obtain perfect *J.S.I.* Therefore, we have evaluated the performance of hard decision receivers, both with and without *J.S.I.*, to provide lower (pessimistic) bounds on performance. In Appendix H we present the evaluation technique, which is based on the use of curves for the cutoff rate parameter R_0 used in conjunction with the BER curves obtained for the ideal soft decision receiver in which *J.S.I.* is available. We also present curves for quaternary ($M = 4$) signaling for several diversity values ($m = 1, 2, 3, 4$).

In Table 7.5 we present the tolerable J/S protection ratios for a high-quality data reception criterion for a nonfading channel and for the three receiver structures of interest, i.e., a soft decision receiver with *J.S.I.* and hard decision receivers with and without *J.S.I.* The same signaling parameters discussed earlier in this section are assumed again. We limit our discussion to diversity values of $m = 1$ and 2, because the use of higher diversity values results in, at best, very little improvement in terms of E_b/N_0 values, while necessitating a significant decrease in data rate.

As we move down the table, the systems considered represent decreasing sophistication. The degree of improvement gained through the use of diversity in a soft decision receiver with *J.S.I.* depends on alphabet size, as we see by comparing systems (1) and (2). The improvement is (0.3, 1.3, 0.6) dB for $M = (2, 4, 8)$. Going from system (1) to system (3), which corresponds to going from soft decision to hard decision while retaining *J.S.I.* and dual diversity, results in a loss of (1.4, 1.4, 1.2) dB for $M = (2, 4, 8)$. Giving up the diversity, the transition from system (3) to (4) results in a further loss of (0.3, 1.5, 1.2) dB for $M = (2, 4, 8)$.

By comparing systems (4) and (5), we see that for a hard decision receiver, dual diversity in a system with no *J.S.I.* is approximately as helpful as *J.S.I.* in a system with no diversity for $M = 4$ and 8; however, for binary signaling the performance of system (4) is 1.4 dB better than that of system (5). System (4) does, however, permit twice the data rate of system (5) for a fixed signaling rate.

It is significant to note that of the four hard decision systems shown in the table, the first three performed within 3 dB of the ideal soft decision receiver with *J.S.I.* and dual diversity. Only the system with no diversity and no *J.S.I.* (system 6) performed more than 3 dB worse than the ideal receiver.

*Assumes either $r_0 = 1/2$ or 1/3 convolutional coding.

Table 7.5 — $J/S|_{max}$ Values for WCPB Noise Jamming; Noncoherent FH/MFSK; R = 2400 b/s; BER = 10^{-5}; W = 5 MHz; Constraint Length 7 Convolutional Coding.

System	Receiver	*J.S.I.*	Diversity	$J/S\|_{max}$		
				$M = 2$ ($r_0 = 1/2$)	$M = 4$ ($r_0 = 1/2$)	$M = 8$ ($r_0 = 1/3$)
1	SD	yes	dual	22.3	24.3	24.9
2	SD	yes	none	22.0	23.0	24.3
3	HD	yes	dual	20.9	22.9	23.7
4	HD	yes	none	20.6	21.4	22.5
5	HD	no	dual	19.2	21.5	22.7
6	HD	no	none	15.6	15.2	19.1

Notes: HD = hard decision
SD = soft decision
For R = 1200 b/s all values are increased by 3 dB
Trumpis rate 1 M-ary codes are used for $M = 4$ and 8;
r_0 = equivalent binary code rate
dual diversity: $m = 2$
no diversity: $m = 1$

In Table 7.6 we present similar results for a medium-quality data reception criterion. As noted earlier, the optimum diversity for a soft decision receiver with *J.S.I.* under a medium-quality data reception criterion is $m = 1$, i.e., no diversity. For a hard decision receiver with *J.S.I.* $m = 1$ is optimal for $M = 2$ and $M = 8$, but $m = 2$ is optimal for $M = 4$. For a hard decision receiver without *J.S.I.* we see that dual diversity results in an improvement of (1.6, 3.1, 0.1) dB for M = (2, 4, 8). As in the case of a high-quality data reception criterion, we again see that the performance of systems 3, 4, and 5 is within 3 dB of that of the soft decision receiver with *J.S.I.* The use of dual diversity is especially significant in 4-ary systems; however, there is virtually no difference in performance in the 8-ary signaling case.

Table 7.6 — $J/S|_{max}$ Values for WCPB Noise Jamming; Noncoherent FH/MFSK; R = 2400 b/s; BER = 10^{-3}; W = 5 MHz; Constraint Length 7 Convolutional Coding.

System	Receiver	*J.S.I.*	Diversity	$J/S\|_{max}$		
				$M = 2$ ($r_0 = 1/2$)	$M = 4$ ($r_0 = 1/2$)	$M = 8$ ($r_0 = 1/3$)
1	SD	yes	dual	23.2	25.5	26.1
2	SD	yes	none	23.8	25.5	26.6
3	HD	yes	dual	21.8	24.2	24.5
4	HD	yes	none	22.4	23.9	25.2
5	HD	no	dual	21.0	23.6	24.1
6	HD	no	none	19.4	20.5	24.0

Tables 7.5 and 7.6 permit us to compare the AJ performance that can be achieved by using difference receiver structures and diversity values. Results vary for different alphabet sizes and data quality criteria. However, we can observe that (except for the case of 8-ary signaling and BER = 10^{-3}), if *J.S.I.* is not available and no diversity is used, then performance is significantly worse than that of a soft decision receiver with *J.S.I.* However, it is encouraging to note that if either *J.S.I.* is available or if dual diversity is used, then performance is within 3 dB of that of the soft decision receiver with *J.S.I.* Of course, the use of dual diversity cuts the data rate in half. Also, the use of a soft decision receiver and the availability of *J.S.I.* imply a more sophisticated receiver structure.

The results of this section permit the evaluation of tradeoffs among data rate, performance quality, signal power and receiver complexity. It is expected that they will provide useful insight into desirable signal design and receiver structures.

7.4 Concluding Remarks

We have evaluated the AJ performance for FH-MFSK signaling and WCPB noise jamming for convolutional coding and three receiver structures, i.e., a soft decision receiver with *J.S.I.* and hard decision receivers with and without *J.S.I.* Although it is expected that one can do better than the performance of the less sophisticated receiver structures, these results do provide a useful pessimistic bound on AJ performance. After considering other factors that may degrade performance, such as maximum feasible hopping rates and available spread bandwidths at HF, the values of $J/S|_{\max}$ may have to be reduced appropriately. Furthermore, channel characteristics are never uniform, and the jamming threat is highly dependent on the particular scenario and is therefore difficult to quantify. Also the combined effects of other users and jammers must be considered to provide a more realistic channel model. These represent further continuing study areas in our ITF Network design. However, based on our current results, it certainly appears that it is possible to achieve some modest to good AJ capability for most links in the HF ITF Network. Coupled with the distributed control and alternate routing paths achievable through the ITF Linked Cluster Architecture, we feel that the system has strong AJ and survivability potential.

8.0 AJ PERFORMANCE OF NARROWBAND BINARY FSK SIGNALING OVER NONFADING CHANNELS

In this section we demonstrate that narrowband systems are extremely vulnerable to jamming, in contrast with the wideband FH systems that were discussed in Section 7. We consider only the case of nonfading channels and rate 1/2 convolutional coding of constraint length 7. This example is sufficient to demonstrate that the use of wideband signaling in military applications not only is advisable, but is, in fact, necessary in many cases. We restrict our discussion to the AJ capability that can be obtained through modulation and coding techniques. In Section 4 we noted that relying exclusively on adaptive arrays for jamming protection is inadvisable.

We assume a Gaussian noise jammer with constant noise density N_0 over the signal bandwidth. The basic narrowband AJ signaling parameters are as follows:

W = signal bandwidth

R = information data rate in b/s

S = received signal power

$J = N_0 W$

= received jammer power (in signal bandwidth W)

Note that W now refers to the narrowband signal bandwidth, rather than to the spread bandwidth of Section 7. Note also that we assume that the jammer knows which narrowband channel is being used, and can therefore concentrate all of his power J in this narrowband.

The resulting bit energy to noise density ratio is easily expressed as

$$E_b/N_0 = (W/R)\ (S/J).$$

The orthogonal tone spacing for noncoherent binary FSK is $1/T_s$, where T_s is the binary symbol duration. A total bandwidth of $W = 2/T_s$ is thus needed to contain the mark and space tones. The data rate can be expressed as $R = r/T_s$, where r is the code rate. For such a signaling scheme without diversity (beyond that inherently provided by the coding process) we thus have

$$W/R = 2/r.$$

For a rate 1/2 code, we therefore have $W/R = 4$.

Background noise is neglected in this phase of the analysis, but it will be introduced in Section 9, when we consider the achievable communication range in a jammed environment; such interference is generally much weaker than the jamming signal in regions where the jammer poses a significant threat. The maximum J/S ratio that can be tolerated by the rate 1/2 coded noncoherent narrowband communication system considered here is therefore

$$J/S|_{\max} = 4/(E_b/N_0),$$

where the value of E_b/N_0 used in this equation is that which is required for the desired BER. The value of E_b/N_0 required to achieve a BER of 10^{-5} for the coding modulation scheme considered here is approximately 10 dB, resulting in,

$$J/S|_{\max} = -4 \text{ dB.}^*$$

Thus, narrowband links are extremely vulnerable to jamming.

9.0 COMMUNICATION RANGE MODEL FOR A JAMMED ENVIRONMENT

In Sections 7 and 8 we demonstrated the AJ link capability, expressed in terms of tolerable J/S ratios, that can be achieved in wideband FH and narrowband systems, respectively. In this section we discuss the resulting communication ranges that can be achieved in a jammed environment. For the case in which all platforms (transmitting, receiving, and jamming) are surface vessels, we demonstrate the region in which a jammer will be disruptive, as a function of transmitter to receiver communication range, for realistic HF ITF signaling parameters. The use of wideband signaling permits the links to withstand considerably stronger jamming attacks than does the use of narrowband signaling.

The actual J/S ratio at the receiver is easily expressed in terms of transmitted power levels, propagation losses, and other system losses:

$$J/S = L_a \frac{P_{tj}/L_j}{P_{ts}/L_s}$$

where the following parameter values have been assumed:

P_{tj} = transmitted jammer power = 10 kW,

P_{ts} = transmitted signal power = 1 kW,

L_j = path loss of jammer,

L_s = path loss of desired signal,

L_a = additional system losses = 3 dB.

*This result is valid for any data rate. Note that the signal bandwidth is proportional to the data rate, and the jammer power is in all cases assumed to be uniformly distributed over the signal bandwidth.

These parameter values result in,

$$J/S = 20\,\frac{L_s}{L_j}.$$

In our examples the use of noncoherent binary FSK signaling with rate 1/2 convolutional coding of constraint length 7 and no diversity is assumed. Successful communication (e.g., BER $< 10^{-5}$) can take place as long as J/S does not exceed the tolerable level $J/S|_{\max}$ discussed in Sections 7 and 8.

Figure 9.1 illustrates the geometry of a transmitter, receiver, and jammer; d is the distance from the transmitter to the receiver, and d_j is the distance from the jammer to the receiver. Note that in this case of a single receiver there is circular symmetry about the receiver. We define $d_{\text{jmin}}(d)$ to be the minimum value of d_j that can be tolerated for a given d, or, in other words, the radius of the circle within which the jammer disrupts communication.

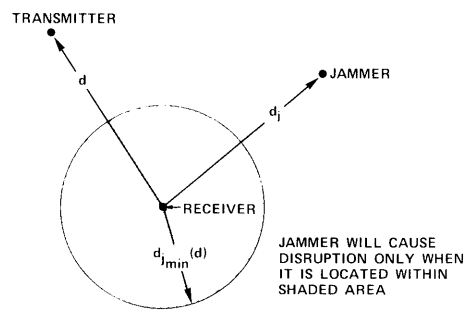


Fig. 9.1 — Transmitter-receiver-jammer geometry

The examples given in this report are based on the assumption that all platforms (transmitting, receiving, and jamming) are surface vessels, and that all propagation is via groundwaves. Therefore, Barrick's curves for groundwave propagation loss over seawater [12] can be used again.

The attenuation of HF groundwaves as a function of distance is considerably greater than inverse square-law; therefore, previously derived communication/jamming range relationships that illustrate the region in which the jammer is effective as a function of various system parameters (e.g., those developed by Cook [44,45]) do not apply here. We have, therefore, derived a communication range model for jammed HF channels that makes use of Barrick's propagation loss curves.

We have examined the case of an extremely rough sea (sea state 6), for which the propagation loss was illustrated in Fig. 6.2. The jamming threat is assumed to be WCPB noise jamming. The AJ capability that can be achieved by means of frequency hopping (one binary FSK symbol per hop) and rate 1/2 convolutional coding of constraint length 7 for sea state 6 is illustrated in Fig. 9.2. In this figure $d_{\text{jmin}}(d)$ is plotted as a function of d for the system parameters discussed above, for the case of an ideal soft decision receiver in which *J.S.I.* is available. As demonstrated in Section 7 the tolerable J/S ratio for BER $= 10^{-5}$ is,

$$J/S|_{\max} = \begin{cases} 22\ dB & R = 2400\ b/s \\ 37\ dB & R = 75\ b/s \end{cases}.$$

We note that $d_{\text{jmin}}(d)$ is less than d, despite the 10 dB power advantage that the jammer has over the transmitted signal. The curves of Fig. 9.2 are based on the simplifying assumption that attenuation is constant over a 5 MHz bandwidth centered at the frequencies shown. The degree of AJ capability is

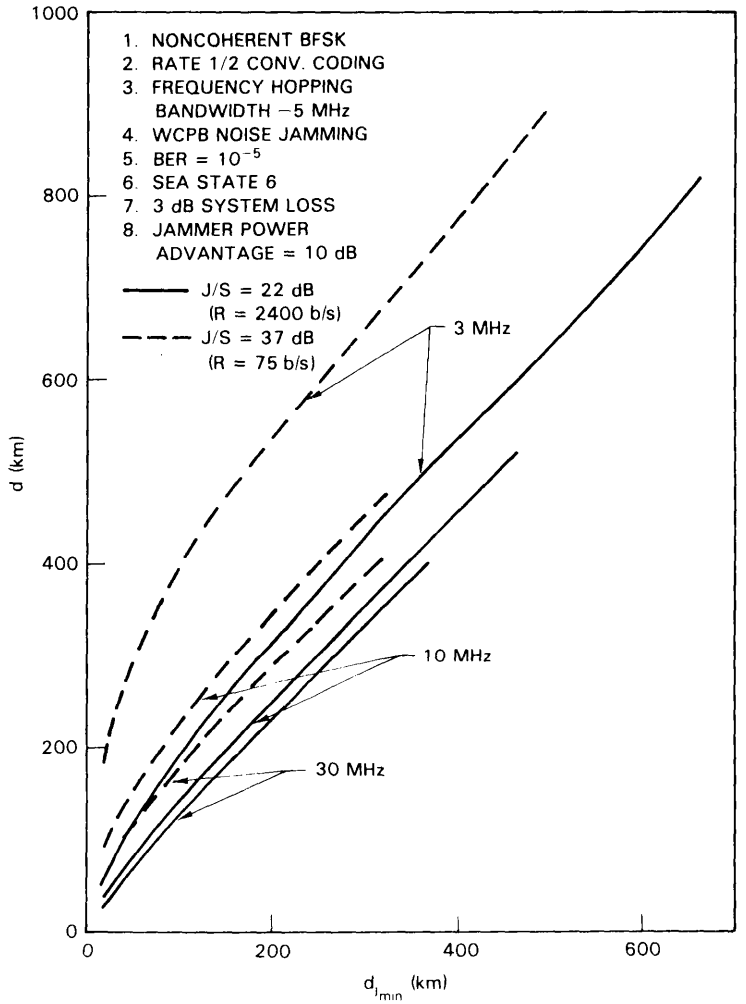


Fig. 9.2 — $d_{J\min}(d)$ vs d for HF groundwave model and wideband signaling

most pronounced at the lower end of the HF band for the examples shown. Lowering the data rate from 2400 b/s to 75 b/s improves AJ performance by increasing the tolerable J/S ratio $J/S|_{\max}$ by 15 dB; the size of the circle in which the jammer is disruptive is thereby reduced, an effect that is also most evident at the lower end of the HF band. Conversely, if $J/S|_{\max}$ is decreased (e.g., by raising the data rate or decreasing the hopping bandwidth), the disruption region is increased.

In our studies, we have exclusively dealt with WCPB noise jamming. Although it is anticipated that multitone jamming may result in a degradation of the tolerable J/S ratio by up to several decibels, the $d_{J\min}(d)$ vs d contours are not expected to change significantly. This insensitivity of performance with respect to small changes in $J/S|_{\max}$ is experienced because the attenuation rate of HF groundwave propagation is considerably greater than inverse square law. Thus, a relatively small change in range is expected to compensate for several dB in $J/S|_{\max}$.

Lowering the data rate by a factor of 32 may be implemented in two basic ways. In the first approach, the data rate reduction is achieved by lowering the hopping rate by this factor; no diversity is introduced. In this case the 15 dB improvement discussed above is a valid result. If, alternatively, we keep the hopping rate fixed at 4800 hops per second, then a lowering of the data rate by a factor of 32 results in the introduction of a diversity of $m = 32$ ($L = 64$). Use of this diversity value results in an

increase in the required value of E_b/N_0 by about 4 dB (see Fig. F1 in Appendix F) as a result of noncoherent combining losses. The corresponding increase in $J/S|_{max}$ is therefore only 11 dB. The curves we present in this section are based on the more optimistic no-diversity case; similar curves corresponding to the case of fixed hopping rate can be generated.

In the narrowband signaling case with rate 1/2 coding, performance is poor in a jammed environment because of the negative tolerable J/S ratio, i.e., $J/S|_{max} = -4$ dB. In Fig. 9.3 we illustrate $d_{jmin}(d)$ vs d for the same system and environmental parameters as considered in the wideband case, except that the signal is narrowband. These curves are valid at any data rate for the conditions discussed in Section 8. We now have $d_{jmin}(d) > d$; thus, the jammer can disrupt communication even when it is farther from the receiver than is the friendly transmitter. We also see that AJ capability is better at the higher end of the HF band for this *non-AJ* system, unlike the FH system discussed above. The propagation loss curves of Fig. 2.1 provide the reason for the improved AJ capabilities at the higher frequencies for non-AJ signaling schemes. These curves show that not only the attenuation, but also the *rate* of attenuation increases with increasing frequency. Consequently, if the jammer is at a range disadvantage, i.e., if the receiver is closer to the friendly transmitter than to the jammer, then the use of higher frequencies may be preferred to maximize the attenuation of the jammer's signal. Conversely, if the jammer has the range advantage, the use of low frequencies is preferred to minimize the attenuation of the desired signal. This point is clarified below.

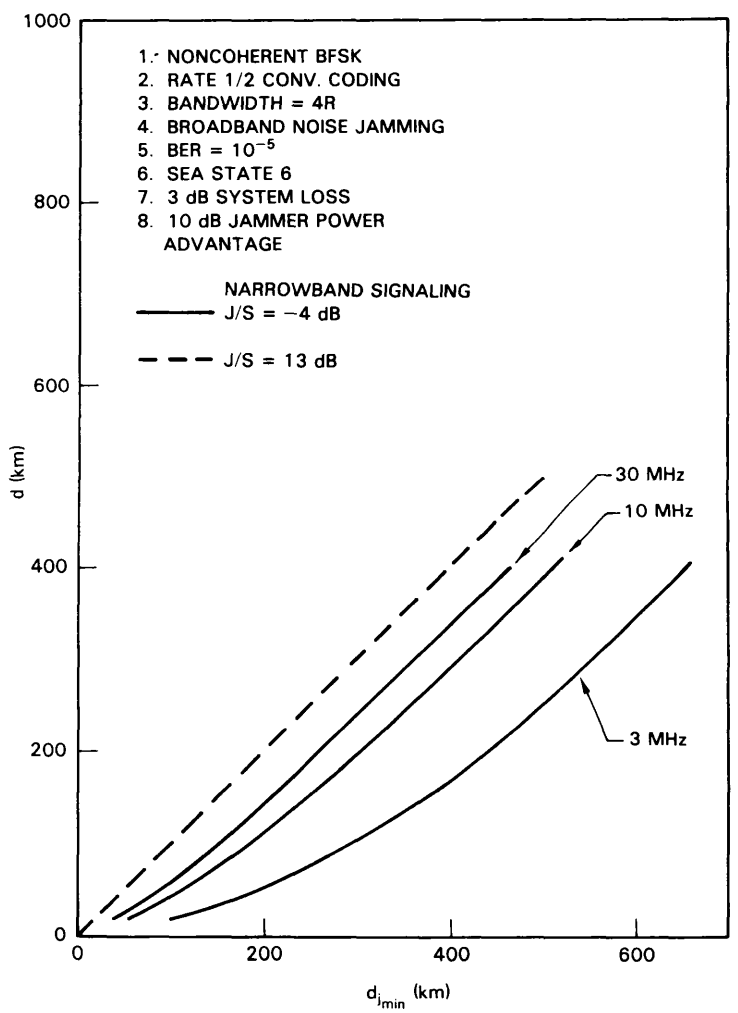


Fig. 9.3 — $d_{jmin}(d)$ vs d for HF groundwave model and narrowband signaling

We note that if system parameters are such that $\{J/S|_{\max} - L_a\}$ is equal to the jammer's transmitted power advantage (10 dB in our examples), then the curves for all frequencies are of unity slope, in which case $d_{\text{jmin}}(d) = d$. Thus, the jammer will disrupt communication if and only if it is closer to the receiver than is the friendly transmitter. We illustrate this case in Fig. 9.3, although we have not associated with it a particular signaling scheme.

In general, we have

$$d_{\text{jmin}}(d) < d, \quad \text{if } J/S|_{\max} - L_a > \text{JTPA},$$

and,

$$d_{\text{jmin}}(d) > d, \quad \text{if } J/S|_{\max} - L_a < \text{JTPA},$$

where JTPA is the jammer's transmitted power advantage. Furthermore,

whenever $d_{\text{jmin}}(d) < d$,

$d_{\text{jmin}}(d)$ increases as frequency increases;

thus we do better at the lower end of the HF band;

and conversely,

whenever $d_{\text{jmin}}(d) > d$,

$d_{\text{jmin}}(d)$ decreases as frequency increases;

thus we do better at the upper end of the HF band.

Background noise has been neglected in the evaluation of $d_{\text{jmin}}(d)$, and thus the curves of Figs. 9.2 and 9.3 depend only on the attenuation predicted by Barrick's model for groundwave propagation loss over seawater (for given waveform/coding design, transmitted power levels and sea state), and not on the background noise environment. These curves are therefore valid only for values of d that are less than the maximum achievable noise-limited values, such as those predicted by the curves of Fig. 6.3.

Under benign operating conditions the communication ranges presented in Fig. 6.3 provide a model for determining the network platform connectivities, and therefore for determining a network organizational structure by means of the Linked Cluster Algorithm (LCA) [3,8,9]. Under jammed conditions communication ranges are decreased, resulting in disrupted platform connectivities. Therefore, the network must reorganize itself, based on the new set of connectivities.

Connectivities in the jammed environment are determined as follows. For each pair of platforms that is connected in the benign environment, connectivity in the jammed environment is examined by computing the J/S ratio at the receiver, or equivalently by examining Fig. 9.2 or 9.3, for wideband and narrowband signaling, respectively. (We have neglected the situation in which the combined background noise and jamming causes loss of links, while neither is sufficient to do so alone.) Although all links that exist in the benign environment are two-way links (under the assumption of equal transmitter powers and uniform interference conditions throughout the network), in the presence of jammers some of them can become one-way links. This distinction is important because one-way links are equivalent to lost links in the execution of the LCA. Further studies are required to investigate possible schemes for the use of such one-way links, as well as asymmetrical links that can support a higher data rate in one direction than in the other.

In Fig. 9.4 we compare $d_{j\min}(d)$ vs d contours for wideband FH and narrowband signaling. These curves have been truncated at values of d corresponding to the communication ranges illustrated in Fig. 6.3. As noted earlier, in the wideband signaling case, performance is better at the lower end of the HF band. Conversely, in the narrowband signaling case, performance is better at the upper end of the HF band, provided that the achievable communication range in a benign environment is not exceeded. Therefore, one might be tempted to adopt the following strategy:

- use the lower end of the HF band whenever

$$J/S|_{\max} - L_a > \text{JTPA},$$

- use the upper end of the HF band whenever

$$J/S|_{\max} - L_a < \text{JTPA}.$$

From the standpoint of maximizing platform connectivity in the presence of severe jamming threats (neglecting issues such as other-user interference and background noise), this is in fact a valid conclusion. However, it must be used with caution.

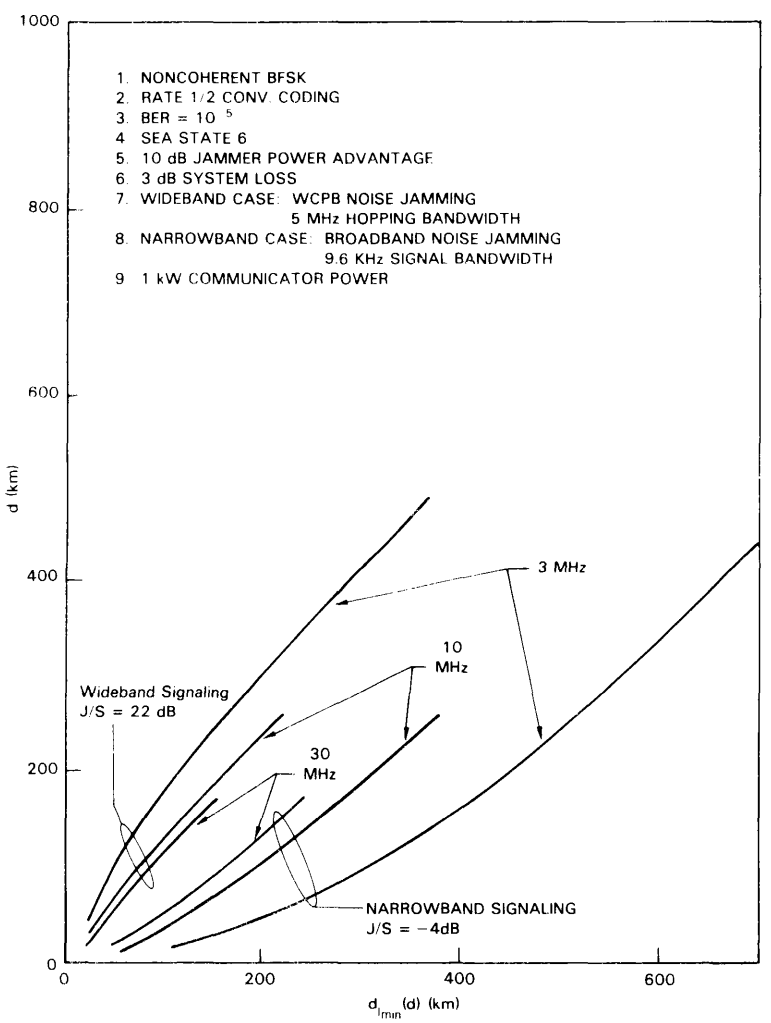


Fig. 9.4 — $d_{j\min}(d)$ vs d for HF groundwave model and R = 2400 b/s; wideband vs narrowband comparison

The problem of HF ITF Network reorganization in response to jamming has been studied by Baker et al [4,11], using the communication range model for the jammed environment that we have presented in this section. In that study the network reorganized itself periodically by executing the LCA. Under benign conditions the use of the lower end of the HF band results in a smaller number of larger clusters, while the use of the upper end of the HF band results in a larger number of smaller clusters. The presence of jammers reduces the communication range in both cases, but still results in a smaller number of clusters at the lower end of the band as long as $\{J/S|_{\max} - L_a\}$ is greater than JTPA. We note, however, that the jammer reduces communication range by a greater percentage at the lower end of the band where communication range is greater in the benign environment. A greater degree of network reorganization in response to jamming is therefore necessary at the lower end of the band than at the higher end of the band; i.e., there will be a greater number of changes in the nodes that are designated cluster heads and gateways. An effective jamming strategy in this case might be for the jammer to remain silent during the network organization period; the network would then organize itself based on connectivities that are characteristic of a benign scenario, but would have to operate in a severely jammed environment.

The point of this discussion has been to suggest that one should establish a BER constraint that is more stringent than is actually needed, especially when using the lower end of the HF band, in anticipation of the jamming threat. If one does not do so, then a network operating at the lower end of the HF band will experience greater disruption than one at the upper end, despite the fact that communication range is still greater at the lower end.

10.0 HF ITF MULTILINK (NETWORK) COMMUNICATION

In point-to-point (or single link) communication, we are primarily concerned with the power budgets that must be maintained in order to obtain acceptable BER performance. Also of crucial importance, especially for spread spectrum signaling, is the issue of acquiring and maintaining synchronization.

In the networking (or multilink) situation the signaling issues become considerably more complex, because now we must consider the coordination of the transmissions of many communication sources. In this section we are concerned primarily with interference caused by other task-force members, and therefore potentially under network control. We discuss the problem areas that arise as a result of such interference, as well as possible methods with which to solve them.

In Section 10.1, we address the issues that relate to networking, as compared to point-to-point communication. Although wideband (i.e., spread spectrum) signaling will be needed to provide AJ capability in the HF ITF Network, we discuss, in Section 10.2, narrowband networking considerations. The Linked Cluster Architecture can, in fact, be implemented using either narrowband or wideband signaling. This discussion serves as a basis for Section 10.3, in which we illustrate the networking implications of the use of wideband signaling as compared with narrowband signaling. Although a completely wideband network would be desirable from the AJ and LPI standpoint, it is clear that it will be many years before all Navy platforms are fully equipped with wideband systems. An evolutionary cycle is expected in which the current narrowband equipment is gradually replaced by the future wideband equipment. As an interim measure, a hybrid HF ITF Network will most likely be implemented, in which each platform will have the capability of supporting a small number of wideband links and a number of narrowband links. Also, it may be desirable to operate the network in a narrowband mode whenever possible, and switch to an AJ or LPI mode, on a link by link basis, only when necessary. Considerations relating to such hybrid networks are discussed in Section 10.4.

10.1 Network vs Point-to-Point Communication Issues at HF

The dependence of groundwave propagation range on frequency is the primary reason for our decision to define a separate linked cluster network for each frequency subband of approximately a few megahertz bandwidth. In the point-to-point case, where we do not explicitly consider the possibility of relaying, we would normally choose the operating frequency on the basis of factors such as the expected E_b/N_0 over the communication link, the possibility of skywave multipath interference, and the availability of a free channel. When there is no provision for relaying, the frequencies with better propagation characteristics (typically the lower end of the HF band which experiences less propagation loss than the higher end) are the most attractive, and might quickly become congested. In practice, relaying will often be necessary in a task force, even if the *best* frequency is used.

In the networking environment, where we can provide for relaying when necessary, there is more freedom to choose the operating frequency. We no longer have to choose the *best* frequency from the standpoint of propagation conditions, but can use any frequency band where there is a path (possibly via one or more intermediate relays) from the source to the destination. It is, in fact, not even necessary to use the same frequency band at each hop along the way. As an example, we consider the set of linked cluster networks that simultaneously operate in different frequency bands. At the lower end of the HF band there will be a smaller number of larger clusters, and at the higher end there will be a larger number of smaller clusters. As long as it is possible to form a connected network at a given frequency band (i.e., a network in which all platforms are connected directly to a cluster head and all cluster heads are connected via the backbone network), it will be possible for any two platforms in the task force to communicate with each other via that network. The communicator's decision of which network to use will be based on overall network performance criteria such as the expected probability of correct message delivery, expected delay time, speech quality for voice, throughput, and possibly other criteria. These network performance measures are related to link propagation loss, interference (mainly other-users and jammers), and congestion at the nodes (resulting in queueing delay or loss of messages). Intermediate nodes in a routing path may switch to a different network (i.e., frequency subband) based on these network conditions.

The use of relaying provides some degree of AJ and LPI capability by reducing the physical separation between transmitter and receiver (see e.g., [44-46]). Adaptive routing schemes can be used in some cases to route messages via platforms that, because of their physical location, are relatively free of jamming interference. The resulting paths may be longer than some more direct paths in terms of the number of relay hops, but the better link characteristics along the more circuitous path may result in an overall improvement in performance [47]. In the case of hierarchical network structures, it may be advisable to actually reorganize the network (i.e., in terms of redesignation of local controllers, etc.). Specifically, Baker et al. [11,4] have demonstrated the ability of the HF ITF Network to reconfigure itself in response to jamming by means of the execution of the Linked Cluster Algorithm.

The degree of AJ performance that can be obtained is related to the equivalent bit energy to noise ratio E_b/N_0 that is achieved over the links of the network. During the execution of the LCA, the platform pairwise connectivities are determined and summarized in the connectivity matrix as binary entries. Connectivity (or the lack of it) must be defined in terms of achieving the desired bit error probability, or alternatively the desired probability of correct packet reception. For a given data rate these probabilities can be related to the interference process for various coding schemes. It is essential that a network that is organized under benign conditions function well while being jammed. The threshold must be set high enough under benign conditions so that a sufficiently high E_b/N_0 is maintained for the highest jamming (as well as other-user interference) levels that are anticipated. Therefore, the threshold should ideally be based on knowledge of whether or not the network is being jammed when connectivity is tested, as well as on the jamming and noise levels expected under stressed conditions. Unfortunately, however, such information about the jammer is often not available.

10.2 Narrowband Network Communication

In a narrowband network a number of narrowband channels, or frequency bins, are available to be allocated among the network's users. The development of link management schemes that permit the efficient apportionment of these frequency bins to the users via either dedicated or shared links is a major network design issue.

Other-user interference will be troublesome only if two or more sources transmit simultaneously in the same narrowband frequency bin. Again we consider other-user interference to consist only of interference caused by other task force platforms; interference resulting from sources external to the task force is considered as part of the background noise level. The problem of other-user interference can therefore be avoided if the members of the network are somehow coordinated so that their transmitted signals are orthogonal, i.e., so that no two platforms transmit in the same frequency bin at the same time. (This condition can be relaxed in some cases to permit so-called frequency reuse as long as there is little or no skywave propagation, and if the two platforms are located sufficiently far apart so that their transmissions do not interfere.)

One possible approach for avoiding such interference is the assignment of a distinct narrowband frequency channel to each potential pair of communicating platforms in the network (neglecting for the moment the situation in which the relaying of messages by intermediate nodes may be necessary). The number of channels required in a fully connected network is $N(N-1)/2$, where N is the number of platforms. For the HF ITF Network we may have $N = 100$, resulting in 4950 channels, a number that would be prohibitive from the standpoint of bandwidth availability. Also, even if bandwidth were not a problem, we would still have the hardware requirement of a transmitter and receiver dedicated to each of these frequency channels at every platform. As a result of such bandwidth and equipment constraints, it is necessary to develop schemes that permit the efficient sharing of the available communication resources. Each narrowband channel can be shared among a number of users by applying *time-domain* multiple access protocols, such as those discussed in [1,2].

The number of narrowband frequency channels (or bins) that are available depends on the total available bandwidth and the bandwidth needed by each signal. In Section 5.2 we considered a quaternary FSK (4-ary FSK or QFSK) signal hopped at 2400 hops per second, with one symbol transmitted per hop. The bandwidth nominally required by such a signal is 9600 Hz. To reduce adjacent channel interference, pulse shaping may be advisable. This pulse shaping would increase the orthogonal tone spacing, thereby resulting in a channel bandwidth of approximately 20 kHz. Therefore, there would be approximately 50 narrowband channels per megahertz of spread bandwidth. In the HF ITF Network we are anticipating that the total bandwidth of each of the simultaneously operating networks under the Linked Cluster Architecture will be between 2 and 5 MHz. The total number of narrowband channels per network would therefore be between 100 and 250. In practice, the number of available channels would be somewhat lower, because certain parts of the HF band are dedicated to specific applications and are, therefore, not usable by the HF ITF Network.

As discussed in Section 1, the Linked Cluster Architecture provides a basis for reducing the total number of narrowband channels that are required, as compared with the totally distributed situation in which each platform maintains a link with each of its neighbors.

Normally, multiple access protocols are considered in the context of a single channel that is to be allocated among a number of users. In contrast, we are considering here the sharing of a number of channels among a number of users, each of which might be limited to the use of only a single channel or a small number of channels at any given time. This situation is repeated in each of the subbands that make up the HF band.

It is possible to consider an HF network that uses only a single narrowband channel, rather than one that uses the many channels that we are assuming to be available. For example, LINK 11 uses a central controller to query other net members when they are to transmit data; this is known in LINK 11 terminology as ROLL-CALL [48].

The use of a single channel for HF ITF communication has also been considered by NOSC [49-51]. Their network concept, like LINK 11, can handle only data; however, it has the advantage of not needing a central controller. Our efforts at NRL have emphasized the use of multichannel networks because of our desire to support greater traffic levels as well as integrated voice and data traffic, and to provide highly survivable communication at HF.

10.3 Wideband Network Communication

The use of spread spectrum signaling has had a great impact on the design of the HF ITF Network. The decision to use spread spectrum signaling is based on the need to provide AJ and LPI capability. Spread spectrum signaling leads naturally to the use of code division multiple access (CDMA) techniques, since under CDMA the dual purpose of providing multiple access capability as well as AJ or LPI performance can be achieved. We use the term CDMA to include all forms of spread spectrum multiple access; CDMA can be implemented using direct sequence (DS), frequency hopping (FH), or hybrid FH-DS signaling. We are interested in FH systems for the reasons discussed in Section 5.

To provide a sufficient degree of AJ and LPI capability, FH systems need a wide bandwidth channel over which to hop, perhaps the entire bandwidth of one of the linked cluster networks, i.e., about 2 to 5 MHz. One wideband channel would thus take the place of roughly 100 narrowband channels. In a *time-domain* system a *collision* occurs whenever two or more signals attempt to transmit simultaneously in the same channel; all signals involved in this collision are usually assumed to be destroyed (actually this is a pessimistic assumption). If a FH-CDMA channel were treated analogously to a time-domain channel, i.e., if only one signal were permitted to occupy it at any given time, the overall network throughput would be drastically and unacceptably reduced. By using CDMA techniques, however, the wideband FH channel can be shared among a number of simultaneous transmissions. We note that it would be possible, in many cases, to achieve greater throughput for a given bandwidth through the use of a number of narrowband channels that are coordinated to avoid contention. The need for AJ capability, however, makes the use of spread spectrum signaling imperative.

In FH-CDMA systems the code corresponds to the FH pattern. CDMA operation is usually asynchronous at the hop level (see Section 10.3.1), and therefore, it is possible for two or more users, using different hopping patterns, to transmit simultaneously in the same frequency bin, which may result in loss of data. The probability of such collisions (of fractions of packets in this case as compared with whole packets in the usual case of time-slotted multiple access schemes), known as frequency *hits,* increases as the number of users increases, thereby resulting in a gradual decrease in successful packet delivery. Error control coding is used to correct errors caused by these frequency hits.

The purpose of this section is to address the major issues associated with FH-CDMA systems and their networking implications. In particular, in this section we address the implications of using FH-CDMA in conjunction with the Linked Cluster Architecture. As in the case of narrowband signaling, there is a need to develop channel allocation schemes for the implementation of dedicated links in the backbone network and multiple access protocols for intracluster communication. These problem areas must be reinterpreted in the FH context, however. For example, we consider here the following problem areas:

a. Synchronization requirements and hopping rates
b. The generation of FH patterns
c. Interference in multiple user FH channels

d. The assignment and distribution of codes (i.e., FH patterns) to platforms
e. Contention among signals using the same code
f. Multiple access protocol considerations for FH networks
g. The coordination of different types of network traffic.

The actual design of multiple access protocols for the HF ITF Network is currently in progress, and a future report will cover this topic in detail.

10.3.1 Synchronization Requirements and Hopping Rates

In our discussion of narrowband networking considerations, we noted that other-user (i.e., other task force member) interference can be avoided if the users are coordinated so that at most one platform transmits in any narrowband channel at any given time. It is, in principle, possible to achieve such an orthogonality in FH systems by coordinating the FH patterns of all users so that no two are scheduled to transmit simultaneously in the same frequency bin, as long as the number of users that are permitted to transmit in the same time slot is not greater than the number of frequency bins. In such a case the FH patterns of each user might be related to a *master* FH pattern by means of either a constant or time-varying offset.

Figure 10.1 represents two signals using orthogonal frequency hopping patterns accurately synchronized at the hop level. In practice, the degree of synchronization required to achieve such coordination is generally not feasible, because of both timing uncertainties and security issues. In Fig. 10.2 we illustrate the use of the same two orthogonal FH patterns, but with a relative delay at the receiver of one hop dwell time. In this case there are frequency hits, thus resulting in loss of data.

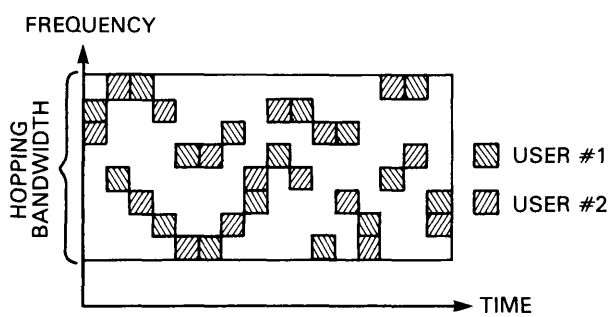


Fig. 10.1 — Two orthogonal FH patterns

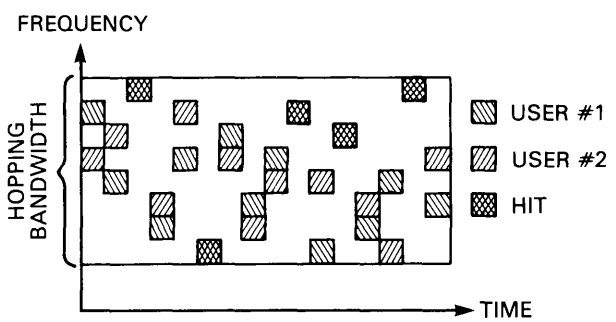


Fig. 10.2 — Two orthogonal FH patterns, one of which is delayed by one hop time

If network-wide synchronization at the hop level were to be maintained, it would be necessary to use guard time delays at each hop to account for timing uncertainties. In the HF ITF Network communication ranges from near zero to as great as 500 km may be expected, resulting in differential propagation time delays of up to 1.67 ms. Also, timing uncertainties at each platform are expected to be of the order of 1 ms. The total timing uncertainty, and therefore guard time requirement, is thus close to 3 ms.

For example, if one were to use a 50% duty cycle waveform (a 3 ms pulse followed by a 3 ms guard interval in each hop interval), the hopping rate would be 167 hops/s. Clearly, the reduced duty cycle cuts the data rate in half. Higher duty cycle systems could of course be used, thus permitting greater data rates, but at the expense of lowering hopping rates further.

An alternative approach may, however, permit hopping rates of about 300 hops/s while maintaining a 100% duty cycle waveform, and therefore experiencing no loss in data rate. The time axis could

be divided into hop intervals of 3 ms duration. Half of the frequency bins would be allocated to the even numbered slots and the other half to the odd numbered slots. As long as the hop duration is greater than the maximum relative timing uncertainty in the network, no two platforms would transmit in the same frequency bin simultaneously.

Although the lower hopping rates may in many cases be acceptable, fast hopping rates are desirable for improved AJ performance (including interference from repeat back jamming) and multipath (skywave interference) rejection. However, fast hopping rates make synchronization more difficult. Our goal is to have a network design that can function satisfactorily at all feasible hopping rates, and not be dependent on either especially fast or slow hopping rate.

Although network-wide synchronization at the hop level is not expected to be feasible, often it will be desirable to maintain synchronization at the packet level. By this we mean the establishment of a network-wide time slotting that is sufficiently accurate to ensure that packets transmitted in any one particular slot do not overlap with packets transmitted by another user in any other slot. The guard time requirement to achieve such synchronization is again approximately 3 ms. This is a small fraction of the anticipated packet length (tens of ms or greater), and therefore represents a small degree of added overhead. Although synchronization at the packet level does not reduce the occurrence of frequency hits, it does, however, facilitate the use of network management schemes that depend on the allocation of network resources on a time division basis.

10.3.2 The Generation of FH Patterns

There are a number of methods that can be used to generate hopping patterns for asynchronous FH-CDMA applications. For example, Sarwate and Pursley have proposed a class of FH patterns based on Reed-Solomon codes that have good cross-correlation properties (i.e., they experience few frequency hits [52]), and consequently good multiple access performance. However, these FH patterns are not suitable for secure military applications; they can easily be predicted by a jammer or interceptor, because they are periodic with period not greater than the number of frequency bins.

Another method that can be used to generate pseudorandom FH patterns is the use of linear feedback shift registers (LFSRs). The output of a set of m taps of the LFSR can be used to specify one of 2^m frequency bins. There are maximal length LFSRs that generate sequences of period 2^{n-1}, where n is the number of shift register stages. For $n = 64$, for example, the period is 1.89×10^{19} binary symbols, which at first glance suggests that they may be attractive for secure spread spectrum systems. However, knowledge of any $2n$ consecutive bits of the pseudorandom sequence is sufficient to predict the remainder of the sequence by means of the Berlekamp-Massey algorithm [53]. The sequences can be made considerably more difficult to predict via the use of nonlinear output logic, i.e, by using nonlinear combinations of the shift register taps [54].

These examples have been presented to illustrate two distinct approaches to the FH pattern design problem. The use of Reed-Solomon patterns is motivated by their multiple access properties, while the use of maximal length LFSRs with nonlinear output logic is motivated by security considerations. There is, in fact, little difference in multiple access performance between the Reed-Solomon and random FH patterns.

The methods actually used to generate secure FH patterns are of course classified, and the network designer will in practice have only limited control over the FH patterns. Therefore, it is not realistic to expect the availability of an orthogonal set of hopping patterns. The most important property of a FH pattern in a secure military application is that it appears to be random. Knowledge of the frequency at any set of hops should provide no information about the future hopping sequence. Secure FH patterns have been designed to satisfy this requirement. Therefore, we can assume in our mathematical analyses that random FH patterns are being used, and that network-wide synchronization

at the hop level is not possible. We can also assume that a very large family of such FH patterns exists, so that several distinct hopping patterns may be assigned to each platform. The intended receiver must, of course, know the hopping pattern of the transmitter and must be able to synchronize to it.

10.3.3 Interference in Multiple User FH Channels

We have noted that the Linked Cluster Architecture can be implemented using either narrowband or wideband signaling. As in the narrowband case, a wideband Linked Cluster network will use dedicated links for much of the backbone network communication. Although in the implementation of dedicated links there are similarities between the narrowband and wideband cases, there are also some major differences.

The most significant difference between FH and narrowband signaling is that in the FH case there is some degree of contention even when dedicated links are used. We define a dedicated link as one in which only a single transmitter can attempt to communicate with any particular receiver at any given time; therefore, there is no contention for access to the intended receiver. Thus, we include time division multiple access (TDMA) as well as other schemes that are *contention-free* in the time domain in the class of dedicated links. However, as a result of the lack of orthogonality among FH patterns, frequency hits may occur because of the interference from other signals (intended for other receivers) that share the same wideband channel. The question, therefore, arises of how many FH signals, that use random independently generated hopping patterns, can simultaneously share the same wideband channel without significant performance degradation. This problem is discussed in detail in Appendixes I and J. The model presented there is based on the work of Pursley and Geraniotis [55-59] who have approached this problem from the standpoint of bit and packet error probability derivations for nonfading and fading (nonselective and selective) multiuser channels, and Hajek [60] who has approached it from the standpoint of multiple access protocol techniques. The number of users within communication range of each other that can transmit over the channel simultaneously depends on factors such as the modulation/coding scheme, relative signal amplitudes, channel characteristics, ability (or lack of it) to detect frequency hits, and the definition of acceptable performance. Further work is needed to apply these results to the HF ITF Network, although a number of results can be presented now.

The most fundamental approach to this problem is in terms of the raw (prior to decoding) symbol error rate of a multiple user channel. Figure 10.3 illustrates an upper bound on the uncoded bit error probability for binary signaling with one symbol transmitted per hop, as a function of E_b/N_0, for a multiple user FH system characterized by a nonfading AWGN channel and a fixed number of simultaneous users. We emphasize that in these examples each of the K users transmits continuously using its own FH pattern. The randomness in the interference process thus arises from the pseudorandom nature of the FH patterns, rather than from variable user demands, although the case of a bursty traffic process can be considered [60]. It is clear from these curves that for E_b/N_0 greater than 12 dB, virtually all errors arise as a result of other-user interference rather than from channel noise. Similar curves for nonselective Rician and Rayleigh fading channels are shown in Figs. 10.4 and 10.5, respectively. In Fig. 10.4 the parameter $\gamma^2 = 0.1$ represents the ratio of the power in the scatter component to the power in the direct path component. For the Rayleigh fading channel we thus have γ^2 equal to infinity. For two or more channel users, the performance on the Rician channel is quite similar to that on the AWGN channel. However, considerable degradation in performance is seen on the Rayleigh fading channel. The system model used to generate these curves is discussed in Appendix J. In the future, curves of this type will be used to evaluate the BER performance of coded systems. We note that Pursley and Geraniotis [55,58,59] have also considered the case of more than one binary symbol per hop, as well as the case of selective fading models.

The curves of Figs. 10.3-10.5 are valid for any relative signal amplitudes, and are pessimistic because they assume that all hits result in a binary symbol error probability of 0.5. A tighter approximation has been obtained under the assumption of equal signal levels [55]. Further improvement is

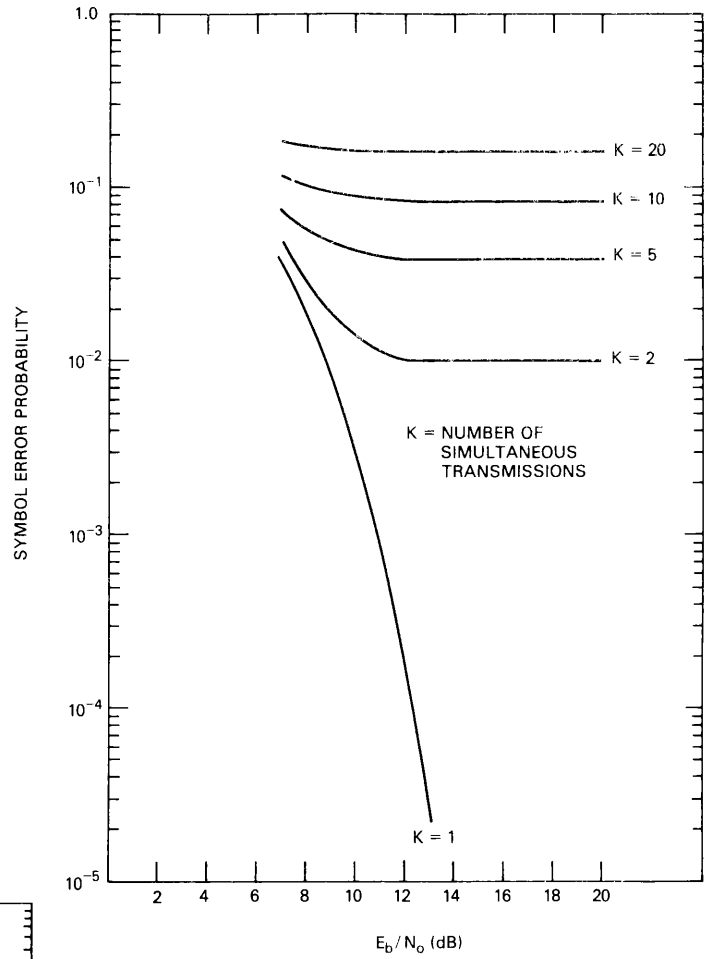


Fig. 10.3 — Raw symbol error probability for an asynchronous FH multiple-user channel; noncoherent binary signaling; q = 100 frequency bins; AWGN channel

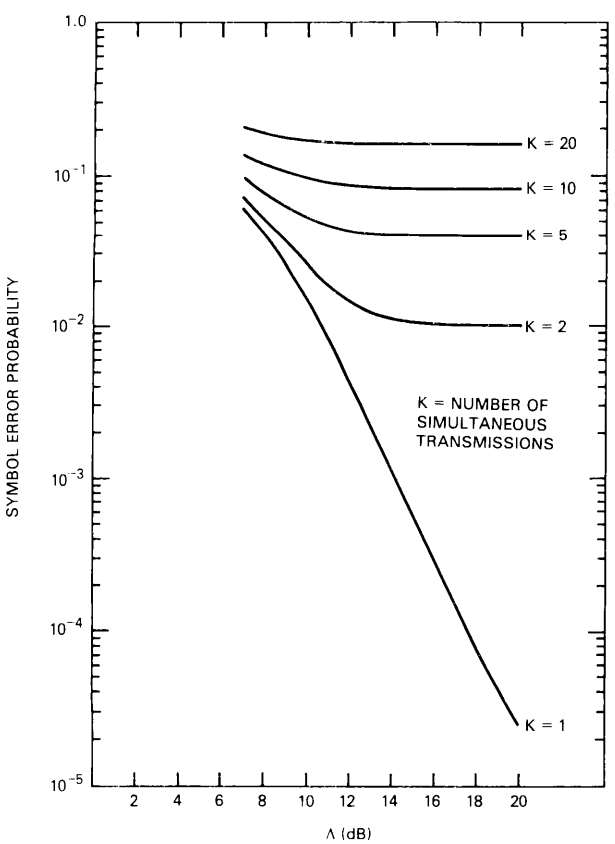


Fig. 10.4 — Raw symbol error probability for an asynchronous FH multiple-user channel; noncoherent binary signaling; q = 100 frequency bins; Rician fading channel $\gamma^2 = 0.1$

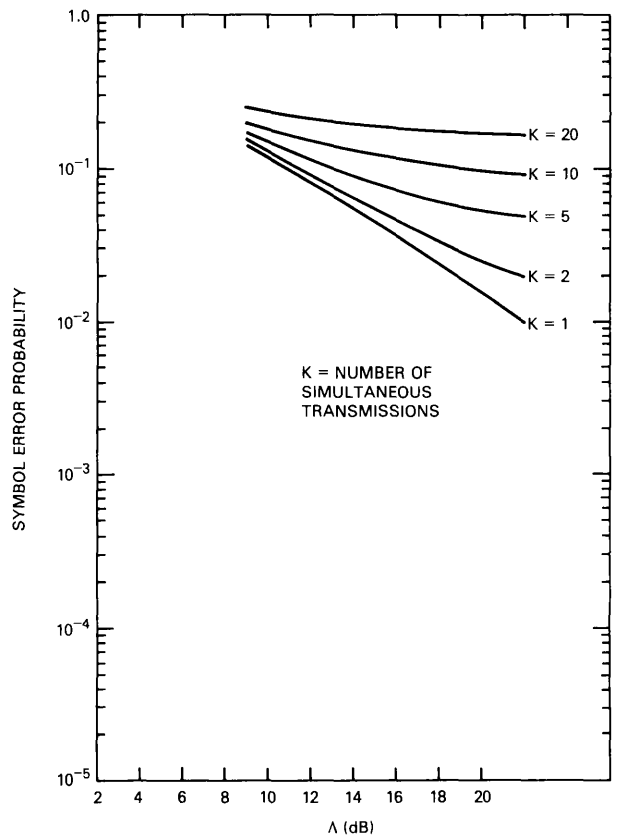


Fig. 10.5 — Raw symbol error probability for an asynchronous FH multiple-user channel; noncoherent binary signaling; q = 100 frequency bins; Rayleigh fading channel

obtained when the interfering signals are of lower amplitude than the desired signal. Frequency reuse at distant parts of the network will often result in negligible interference levels. However, we must also consider the fact that in frequency bands that support skywave propagation, the number of interfering signals will be increased. Thus, all network members can potentially interfere with each other, and some will produce both groundwave and skywave interfering signals. The effective number of other users may then be approximately twice the actual number of other users.

The problem of other-user interference in FH systems can also be approached in terms of packet error probability. Under the model developed by Pursley [55-57], and Hajek [60], and later extended by Wieselthier and Ephremides [61-64]. Reed-Solomon coding is used to correct the burst errors caused by frequency hits in a slow FH system (i.e., one in which several bits are transmitted per hop). This model is discussed in greater detail in Appendix I. Figure 10.6 illustrates the probability of packet error as a function of the number of users that are transmitting simultaneously over a channel with 100 frequency bins, for the case in which RS coding of rate approximately 1/2 is used. In this case the channel is assumed to be noiseless, and so the only source of errors is other-user interference. However, a noisy channel model can also be considered, as we discuss shortly. We consider two packet sizes chosen so that the data of a packet can be encoded as a single Reed-Solomon codeword. The RS-(31,15) code corresponds to a packet length of 155 bits; the packet is divided into 31 five-bit *bytes,* one of which is transmitted at each hop. This code is capable of correcting all codewords in which there are eight or fewer byte errors. The RS-(255,127) code corresponds to a packet length of 2040 bits, which are divided into 255 eight-bit *bytes.* This code is capable of correcting up to 64 byte errors. Virtually all packet errors are detectable, often permitting the subsequent retransmission of packets that are received with uncorrectable errors. The probability of an undetected packet error is less than 2×10^{-5} for the RS-(31,15) code, and less than 10^{-89} for the RS-(255,127) code, as discussed in Appendix B.

Figure 10.6 shows three curves. The two upper curves were generated under the assumption that frequency hits are not detectable and that they all result in byte errors. This is a pessimistic assumption, because relatively strong signals will certainly have lower error probability than weaker signals, and also because signals overlapping for a small fraction of the hop duration will not necessarily result in byte errors. The bottom curve was generated under the assumption that frequency hits are detectable and that the corresponding bytes can be erased. Knowledge of which symbols are affected by interference is often called *side information.* The packet error probability using the RS-(255,127) code with detectable hits is extremely low; it falls below the range of the plots. A similar set of curves, for the case with no side information, and in which the byte error probability in the absence of other-user interference (i.e., the byte error probability resulting from background noise, other users of the HF band external to the task force and jamming) is 0.1, is shown in Figure 10.7. The ability to detect frequency hits and to erase the corresponding bytes, therefore, results in a considerable increase in the number of simultaneous users that the FH channel can support. The tradeoffs between the increased complexity of such a system and its benefits must be addressed in the future.

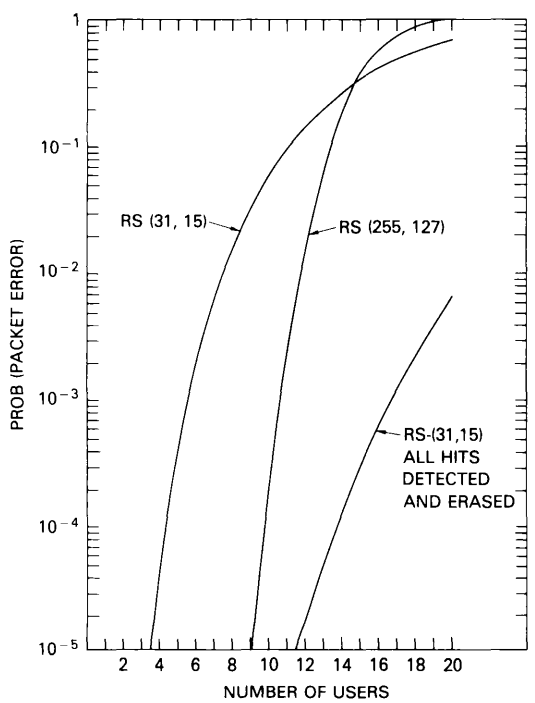


Fig. 10.6 — Packet error probability for a noiseless asynchronous multiple-user FH channel; rate $\approx$ 1/2 Reed-Solomon coding; q = 100 frequency bins

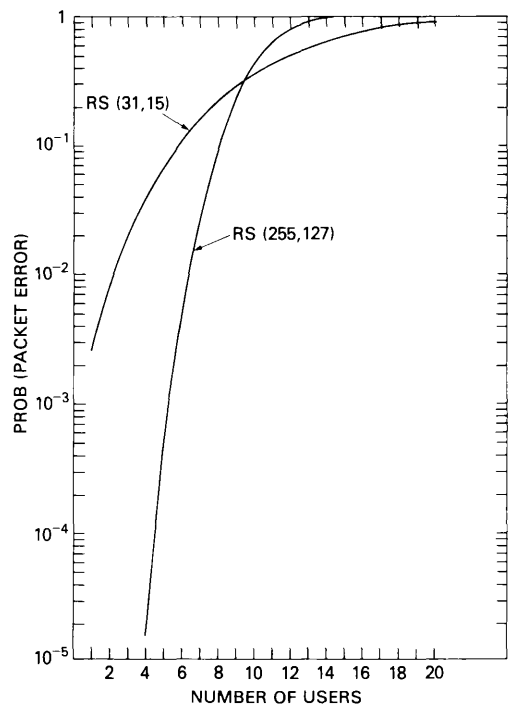


Fig. 10.7 — Packet error probability for a noisy asynchronous multiple-user FH channel with a byte error probability of $p_1 = 0.1$; rate $\approx$ 1/2 Reed-Solomon coding; q = 100 frequency bins

Wieselthier and Ephremides [61-64] have analyzed a signaling scheme that uses a more sophisticated receiver that can make correct byte decisions despite the occurrence of frequency hits, provided that the other-user interference is present for a sufficiently small fraction of the hop duration.

The throughput of a multiple user FH channel can be defined as the expected number of correct packets that can be delivered per time slot; a time slot is equal to a packet duration, and is therefore different for the different codes and packet sizes used. Thus, throughput is simply the product of the number of users and the probability of correct packet delivery for any given user. Under this model, packets that are incorrectly received are subsequently retransmitted. Thus, we have an ARQ system that is combined with Reed-Solomon coding. The ability to detect virtually all packet errors when using Reed-Solomon codes makes such a performance measure appropriate, provided that a mechanism for

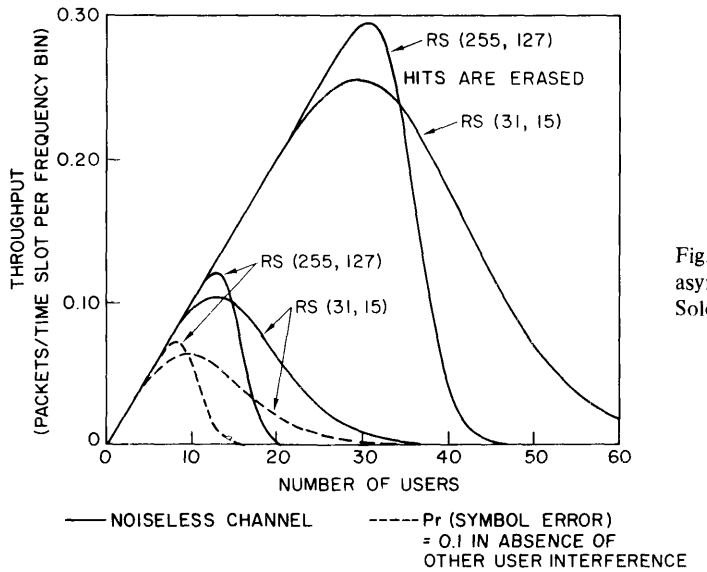


Fig. 10.8 — Throughput per frequency bin of an asynchronous multiple-user FH channel; rate ≈ 1/2 Reed-Solomon coding; q = 100 frequency bins

packet acknowledgment and retransmission can be implemented. We actually consider the throughput per frequency bin, expressed as packets/time slot per frequency bin. Figure 10.8 shows the throughput achievable on a channel with 100 frequency bins when using the RS codes discussed above, for the noiseless channel as well as for a channel in which the noise-induced byte error probability (in the absence of other-user interference) is 0.1. In the noiseless case, the maximum throughput is achieved for 13 users and is 0.121 packet per time slot per frequency bin for the longer code, and 0.103 for the shorter code.

When operating at the maximum throughput level, the resulting packet error probability (and therefore retransmission probability) might be too high for some applications, e.g., those in which retransmission delays cannot be tolerated or cases in which buffer space is limited at the terminals. In such cases it may be necessary to reduce the number of users that transmit over the wideband channel simultaneously. For example, for the RS-(255,127) code the retransmission probability is 0.069 when operating at the maximum throughput level in a noiseless environment. By reducing the number of users from 13 to 10 the retransmission probability is lowered to 1.9×10^{-4}, while reducing throughput from 0.121 to 0.1; this may be a favorable compromise in many cases. Further work is needed to assess the tradeoffs between retransmission probability and throughput that affect network operation.

Figure 10.8 also shows the noiseless case in which hits are recognized and erased. Throughputs of 0.294 and 0.254 packets/time slot per frequency bin are achievable by using the longer and shorter code, respectively. The ability to detect hits and to erase the affected bytes thus increases the achievable throughput by a factor of approximately 2.4.

Alternatively, it is possible to consider the bit error probabilities (rather than the packet error probabilities) that are associated with the use of these codes, although the transformation from packet error probability to bit error probability for Reed-Solomon codes is not straightforward. Bit error probabilities may be more appropriate than packet error probabilities when long data streams, consisting of many codewords, are transmitted.

The use of block codes in this discussion should not be interpreted as a recommendation to use block codes (with slow frequency hopping), rather than convolutional codes, in the HF ITF Network. Such a recommendation would be premature at this point. The use of packets that consist of a single block codeword appears attractive from the standpoint of performance, and it facilitates the type of

multiple access discussion presented here. Earlier, in Section 3, we discussed some of the tradeoffs between the use of convolutional and block codes.

10.3.4 The Assignment and Distribution of Codes

We have noted that the network designer will have only limited influence over the FH patterns used in the HF ITF Network, because the primary factor in the code generation process is security rather than multiple access capability. The most that can be assumed about the FH patterns is that they are random in nature, and that a large family of such patterns does in fact exist. The only form of network-wide synchronization of hopping patterns might be at the packet level, rather than at the hop level, as we have discussed earlier. The major issues relating to code selection are therefore related to the choice of codes for particular links and the distribution of a set of codes among the network's platforms, rather than the actual generation of these codes.

The basic code assignment considerations include the allocation of FH codes to users and the decision of whether to use a code associated with the transmitter or with the receiver, or in some cases a common code. The four basic types of links in the network will be point-to-point links, broadcast links, random access links, and common links.* Links of each of these types will share a wideband FH channel simultaneously. Considerations for these links differ; therefore, they will be considered individually.

Point-to-Point Links: A point-to-point link is characterized by contention-free operation in which a single source transmits to a single destination. When such links are used, the FH code of either the transmitting or receiving platform may be used, as long as consistency is maintained.

Broadcast Links: In a broadcast link a single transmitter sends a common message to two or more platforms. Therefore, it is appropriate to use a FH pattern associated with the transmitting platform, and monitored by all potential receiving platforms. Although a common code could be used in some cases, the use of transmitter-based code permits several broadcast links to operate simultaneously in the same neighborhood.

Random Access Links: In a random access link many platforms attempt to communicate with a single platform† on a contention basis. A FH pattern associated with the receiver is therefore essential, because the receiver will not know a priori which platform(s) are attempting to communicate with it in any particular slot.

Common Links: A common link, monitored by a number of network platforms, can be used for either contention-free broadcast or random access applications. A single FH pattern would be associated with such a link. Links of this type may either be established over a portion of the network (e.g., one might be associated with each cluster), or they may be network wide.

We now discuss the number of FH codes that may be used in the implementation of the HF ITF Network. To simplify the discussion, we assume at first that only a single transmitter and a single receiver are available at each platform in every subband. We shall demonstrate that it is sufficient in this case to associate one *transmitter-based FH code* and one *receiver-based FH code* with each of the N platforms in the network, and that two additional codes are needed. The total number needed in the network is therefore at most $2N + 2$. This situation is repeated in each of the Linked Cluster networks that are simultaneously operating in other frequency bands.

*We have used a rather broad definition of the term *link* in which we include multisource and/or multidestination traffic. Several links, which may be of different types, may share the wideband channel simultaneously using CDMA techniques.

†The case of random access to several destinations simultaneously is considered in the class of common links.

The types of links that are expected to be supported in the Linked Cluster Architecture are as follows:

a. A link, monitored by all platforms, to be used during the execution of the Linked Cluster Algorithm (LCA). Such a link may also be used during normal network operation.

b. A random access link that is monitored by all task force nodes.

c. A broadcast link associated with each cluster head.

d. A random access link associated with each cluster head.

e. Dedicated links to be used on the backbone network.

f. Auxiliary links.

We now discuss the FH codes that are required for the implementation of these links:

a. The use of a single network-wide common contention-free code would be adequate for the execution of the LCA. It would not be necessary to use a unique code associated with each platform (although unique codes could be used), since the transmissions of each platform are inherently noninterfering because of the TDMA frame structure of the LCA. Similarly, during normal network operation, a fraction of time slots may be allocated to the transmission of certain types of information on a network-wide TDMA basis. One application for such a mode of operation might be for communicating with platforms that do not know the transmission schedules throughout the network. The same code used during the execution of the LCA may be used.

b. The availability of a single network-wide random access link would allow a platform that is not part of the network formed by the LCA, to communicate with a node that does belong to this network. This might occur, for example, when a rapidly moving aircraft or a previously radio-silent submarine wishes to communicate with a network node. Another example of a need for a random access link would be to accommodate the merger of two task forces. Although equipment constraints would prohibit the continuous monitoring of such a channel, a fraction of each receiver's time slots may be used for this purpose.

c. A broadcast code would be used by a cluster head to transmit simultaneously to all members of its cluster, and each platform would monitor its cluster head's broadcast transmissions. One transmitter-based FH code should thus be associated with each cluster head to implement this kind of broadcast. In addition, since a node may be within range of several cluster heads, it might be worthwhile for it to monitor the broadcasts of other neighboring cluster heads in addition to its own. To do so would not require additional broadcast codes, but rather the ability to monitor additional codes. This would not necessarily require additional receivers, since it would be possible to monitor several cluster heads on a time division basis.

d. Intracluster communication, from cluster members to their head, can be implemented via either a contention-based or a contention-free mechanism. A transmitter-based FH code cannot be used when contention is permitted; either receiver-based or common codes are appropriate. There is more flexibility in the contention-free case; transmitter- or receiver-based, or common codes may be used. One reasonable approach, suitable for contention-based or contention-free operation, is to associate one receiver-based FH code with each cluster head. Note that each cluster member can use the receiver-based FH code of the cluster head to monitor the transmissions of the other cluster members that it can hear. The cluster head's receiver-based code may thus take on the role of a clusterwide common code.

e. The backbone network, defined in Section 1.3, consists of the cluster heads and gateways and the point-to-point links that connect them. These links can be implemented using the FH code of either the transmitter or receiver. The use of FH patterns associated with the transmitter offers the added flexibility of permitting several gateways (or other nodes) to monitor these transmissions.

f. The *primary network* consists of the backbone network plus those links that connect cluster members with their heads. Network links that are not part of the primary network are known as *auxiliary links.* Such links may be either point-to-point or broadcast, and may be either contention-free or contention-based. Clearly, the choice of codes used for auxiliary links follows the reasoning presented above. For example, it is appropriate to use transmitter-based codes when the channel access scheme is contention-free or if the transmission is of a broadcast nature. It is best to use receiver-based codes for contention-based schemes. Common codes may also be appropriate if, e.g., the source does not know who the destinations will be or vice versa. Actually, common codes could be used for all applications, but their use would prevent the simultaneous operation of multiple links.

We have assumed that each platform can transmit at most one packet at a time, and can receive at most one packet at a time; (simultaneous transmit/receive operation is expected to be feasible). To implement the types of links discussed above, the transmitting platform would choose the appropriate code from one of the following:

a. its own transmitter-based code,

b. the common code associated with its cluster (which could be the cluster head's receiver-based or transmitter-based code),

c. the receiver-based code of the destination platform,

d. the network-wide common broadcast code,

e. the network-wide common random access code.

The simplest approach to assign the frequency hopping patterns to the network platforms is to associate a priori a pair of unique codes with each platform in the network. It may be possible to disseminate this information to all task force members before deployment of the task force, so that each platform knows the codes to be used by every other platform. There would then be no need to develop a distributed scheme for the dissemination of this code information, although of course the need for channel allocation schemes to share resources on a time division basis still exists. If codes must be changed, then some mechanism will have to be developed to broadcast this information throughout the network. Security will be a major issue in the distribution of FH patterns, because it is essential that they are not known by an adversary.

The approach outlined here, of assigning one transmitter-based code and one receiver-based code to each network platform, plus the use of two additional network-wide common codes, permits the implementation of a Linked Cluster network. The total number of FH codes required in a network of N platforms is then $2N + 2$. This situation is repeated in each of the networks in different subbands that make up the HF ITF Network. Actually, a smaller number of codes could be used, because only a subset of network platforms will transmit in any slot. However, the use of distinct codes associated with each platform simplifies the choice of FH codes by eliminating the need to develop a distributed algorithm for code assignments.

Also, the availability of a distinct pair of codes at each network platform would greatly facilitate the implementation of quasi-orthogonal auxiliary links. The code assignment problem is therefore

transformed into a combined code and time slot assignment (channel allocation) problem. A distributed channel allocation scheme has in fact been developed for the HF ITF Network [65,66,8,9]. Such time division also helps to limit to acceptable levels the number of signals that are transmitted simultaneously. Hajek [67,68] has investigated alternative link allocation schemes that are designed to reduce the number of neighboring users that transmit simultaneously using different codes.

It is straightforward to generalize the above discussion to the case in which each platform has several transmitters and receivers for use in the same subband. It would be possible to associate a distinct code with each of them, and each platform would then be able to support additional communication links. It will be necessary to generalize the channel allocation algorithms to the case of multiple simultaneous channels per platform. The availability of several receivers will permit a platform to listen to several signals (transmitted on different codes) at once, and may result in considerable performance improvement. However, the availability of several transmitters is not expected to result in much improvement. Channel access protocols designed for CDMA networks must take into consideration the other-user interference levels. These schemes may not be able to take advantage of the additional transmitters, because the number of users transmitting simultaneously must not be too great if acceptable link performance is to be maintained. In future studies, we hope to address quantitatively the performance of schemes that use multiple transmitters and/or receivers.

10.3.5 Contention Among Platforms Using the Same Code

Another significant problem area is that of contention among signals that use the same code. Such a situation may arise when two or more platforms attempt to transmit to the same receiver simultaneously. We assume that no two receivers will use the same code, except to monitor a network-wide (or cluster-wide) random access channel or a common channel. The simplest case to consider is that of only two signals. The two basic situations that may arise are those in which the delay at the receiver between the two signals is:

a. less than the hop duration

or

b. greater than the hop duration.

In case *a* there is a collision that is characteristic of narrowband systems, and typically both signals are lost, unless one is considerably greater in amplitude than the other. In case *b* it will often be possible for the receiver to acquire and maintain synchronization with the signal that arrives first. Once a signal has captured the receiver, a delayed version of the same signal (generated by multipath propagation) or another signal that uses a delayed version of the same hopping pattern ideally will not be troublesome (although, in practice, synchronization can be disrupted by fading or jamming, especially when platform mobility results in varying propagation delays). The study of spread spectrum multiple access schemes that make use of the capture property is in its early stages [69-70]. A combinatorial technique for the analysis of framed contention-based multiple access protocols with a general model for capture has recently been developed by Wieselthier and Ephremides [71]. Further effort is needed to evaluate the performance of contention-based FH-CDMA systems that use a common code, and to assess the implications of their use in the HF ITF Network.

10.3.6 Multiple Access Protocol Considerations for FH Networks

The design of multiple access protocols for the HF ITF Network must take into consideration the use of FH signaling. It is not sufficient to design time-domain multiple access schemes independent of the signaling format, and then implement them using spread spectrum signals. The nature of FH multiple user channels in some ways facilitates the operation of multiple access protocols, while in other ways it complicates matters. A number of these properties have already been discussed. We now address these properties in the context of multiple access protocols.

The multiple user capability of FH channels (Section 10.3.3) provides flexibility in the design of multiple access protocols because it permits a number of users to share the same wideband channel simultaneously. In contrast, in purely time-domain schemes one must ensure that neighboring users do not transmit in the same time slot. A related property is the ability of a receiver to select one of several simultaneously transmitted signals by monitoring the appropriate code. This property was used by Wieselthier and Ephremides in the design of a distributed reservation scheme for spread spectrum channels [72].

Alternatively, one might use a number of narrowband links to achieve a multiple user channel capability. Successful communication would then require that only one user transmits in any narrowband channel at a time. It might be possible to achieve greater total throughput through the use of multiple narrowband channels in such a manner, but one would of course lose the AJ capability, which is the primary reason for using spread spectrum signaling.

The capture property also permits flexibility in the design of multiple access schemes. Let us consider the case of a number of users attempting to access a receiver by using the FH pattern of the receiver. In this case, not all packet collisions will result in the loss of all packets that are involved, but one packet may be able to capture the receiver if the other packets arrive following a relative delay of more than one hop time, as discussed in Section 10.3.5. The result is improved performance in terms of delay, throughput, and stability.

The use of FH signaling imposes synchronization constraints that can have a profound effect on the performance of multiple access protocols. The acquisition of synchronization, which is necessary for each packet, can be difficult if the interference level (from background noise, other users, or jammers) is high. The time required for synchronization results in overhead in each packet, and it puts a lower bound on the size of packets. The performance of reservation schemes that rely on minipackets to make reservations [73] is especially affected by such overhead requirements. Furthermore, multiple access schemes must be able to cope with the situation in which synchronization is not always achieved.

Many of the control schemes that have been proposed for time-domain multiple-access channels require that the users monitor the channel to determine whether there has been a single (and therefore successful) transmission, two or more transmissions (all of which are usually assumed to be unsuccessful), or no transmissions. The situation in a FH-CDMA channel is different because several signals can, in fact, be successfully transmitted simultaneously, as we have already discussed. Another characteristic of FH-CDMA is that it is not feasible to monitor the success or failure of each of the individual transmissions of all the other users, because to do so would require the use of a separate frequency hopped receiver corresponding to the FH pattern of every user in the population.

Instead of attempting to monitor individual transmissions, it may be better to observe the overall FH channel. During each time slot, every user can use a random hopping pattern to hop a receiver among the q frequency bins. To do so, however, requires the use of a receiver that one might prefer to use for other purposes. An estimate of overall channel activity can be obtained by counting the number of dwell times that the channel is in use in the monitored frequency bins. Hajek, [60] has demonstrated that feedback information of this type can be used in conjunction with the techniques of [74] which he orignally developed for time-domain channels, to implement a control policy suitable for FH random access systems.

Hajek has also developed the Acknowledgment Based Retransmission Control (ABRC) policies, a class of policies that require a minimum of feedback channel information [75]. Under these schemes a user needs to know only whether or not his own transmissions are successful; no additional feedback information (e.g., number of time slots in which collisions occur) is required. The minimal feedback requirements of the ABRC policies make them attractive for use in FH systems, where feedback information is difficult to obtain.

Carrier Sense Multiple Access (CSMA) schemes are often used in random access applications in which the propagation delay is short compared to the packet length [76], a condition that is expected to be satisfied in the HF ITF Network. Under such schemes a user that has a packet to transmit will do so only if he senses the channel to be silent. In a FH multiple user channel, however, the user would have to determine whether or not another user were communicating with the particular destination to which he wanted to transmit. To sense that someone else was transmitting to that receiver would require synchronization with the transmitted signal of the other user. The implementation of such a scheme might be feasible, but considerably more complex than for a standard time-domain CSMA scheme. Furthermore, it would require considerably more delay (enough hops to synchronize to the other user's signal), thus resulting in performance degradation. (The performance of CSMA systems is strongly dependent on the speed with which a busy channel can be sensed.) However, a quantitative performance analysis of FH-CSMA has not yet been performed.

Similar to CSMA is Busy Tone Multiple Access (BTMA) [77], in which the destination broadcasts a narrowband tone to inform the other users that it is busy. Such a scheme is useful when a central station is within range of a number of platforms, but not all platforms are within range of each other. This is known as the *hidden terminal* problem. Such a situation can arise within the individual clusters that make up the HF ITF Network. A narrowband tone would not be appropriate for most military applications, however, because a spoofer would be able to inject spurious busy tones into the system, and thus could disrupt communication by making the channel appear busy at all times. The use of a FH busy signal would require synchronization delay as in CSMA (in this case the delay would occur at the central station as well as at the user when trying to acquire synchronization with the busy signal), and thus would result in degraded performance as compared with the narrowband case.

In this subsection we have illustrated some of the major issues that are associated with the implementation of multiple access protocols with FH signaling. The field of FH multiple access protocols is a new area that requires much basic research. For example, further research is needed to translate multiple access protocols originally designed for use in the time domain into multiple access schemes that function well in the FH code domain. Furthermore, additional research is needed to develop schemes that take full advantage of the properties of FH multiple user channels. The design of multiple access protocols for the HF ITF Network is currently in progress. In a future report we will discuss FH multiple access issues in greater detail than they have been presented here, and we will also present proposed designs for multiple access protocols.

10.3.7 The Coordination of Different Types of Network Traffic

The other-user interference environment in the HF ITF Network is very complex, because backbone network signals, intracluster signals, and auxiliary link signals will use the same jammed wideband channel simultaneously. Methods are needed to ensure that acceptable interference levels, and therefore performance levels, are maintained. The situation becomes considerably worse at frequencies that support substantial skywave propagation, since the number of interfering signals is greatly increased. Since there is no way to coordinate the FH patterns at the hop level on a network-wide basis, the only way to reduce the degree of contention and thereby reduce error rates is to reduce the number of signals that are simultaneously transmitted within any frequency band or subband. To do so is one of the goals of the channel allocation algorithm developed by Baker et al. [65,66,8,9] and the multiple access protocols discussed by Wieselthier et al. [1,2].

It is especially important to ensure that a robust backbone network is maintained, because the control structure of the network will normally make extensive use of the backbone network's capabilities. It may be possible to maintain the integrity of the backbone network by ensuring that backbone network signals and nonbackbone network signals do not contend for the same part of the wideband channel simultaneously. One way to do so would be to structure the FH patterns so that a certain portion of the hopping bandwidth is allocated to the backbone network (or, in general, to high priority

traffic), and is thus protected from interference caused by lower priority traffic. This portion of the hopping bandwidth should be varied every few hops to prevent the jammer from concentrating his energy on the control structure of the network. Although, as discussed in Section 10.3.2, the network designer is expected to have only limited control over the choice of hopping patterns, it would be desirable to be able to partition the subband on a time-varying basis. By protecting a portion of the hopping bandwidth from the interference by lower priority traffic, one could maintain the integrity of the higher priority traffic without unduly structuring access to the channel by the lower priority traffic. Thus, one would still be able to use a large variety of multiple access schemes for lower priority traffic, including highly nonstructured contention-based schemes, without adversely impacting on the operation of the higher priority traffic.

10.4 Hybird Network Communication

We have noted that, although wideband communication is required to provide adequate AJ performance, an evolutionary cycle is expected in which the current narrowband equipment is gradually replaced by the future wideband equipment. We must therefore consider, as an interim measure, a hybrid HF ITF Network in which each platform can support a small number of wideband links plus a number of narrowband links. Hybrid networks may also be used even when wideband equipment is available, because it may be desirable to operate the network in a narrowband mode whenever possible. The network might switch to an AJ or LPI mode, perhaps on a link by link basis, only when necessary. Such mode switching is certainly a nontrivial operation, because it requires the coordination of all platforms that are involved. Such coordination may be difficult to achieve in a jammed environment. Furthermore, we note that it is undesirable for the network to exhibit obvious mode changes, because this would provide the jammer with an assessment of the success of the jamming operation.

We are primarily concerned in this subsection with the interference environment in a hybrid network, rather than with the protocols that may be used in such a network. It is possible to consider a network in which the set of frequency bins allocated to wideband communication and the set allocated to narrowband communication are mutually exclusive. More interesting, however, and probably more realistic is the case in which narrowband signals and wideband FH signals share the same wideband channel.

We first consider the interference problem from the viewpoint of wideband signals. The presence of narrowband signals in a wideband FH channel simply results in a frequency hit whenever the FH signal enters an occupied frequency bin. As in the case of a purely FH network, forward error correction coding must be used to handle loss of data resulting from hits. It makes no difference to the wideband signal whether the hit results from interference with a wideband or narrowband signal.

Next, we consider a narrowband signal that suffers hits from one or more FH signals. Usually, only a small fraction of the data will be affected, as long as the number of FH signals is not too large. Forward error correction coding can be used to correct the resulting errors. There may, in fact, be no loss of data if the hop rate is considerably faster than the data rate of the narrowband signal. For example, a 400 μs hop dwell can probably be tolerated by a 13.6 ms Link 11 signal, if the amplitude of the hit is not much greater than that of the signal.

The main point of this discussion is that FH signals can share a frequency band with narrowband signals, while maintaining acceptable performance levels among both classes of signals, as long as the number of simultaneous users is not too great. Further work is needed to quantify this performance.

11.0 SYNCHRONIZATION ISSUES AND THEIR IMPACT ON THE HF ITF NETWORK

Throughout this report, in our discussions of FH systems we have assumed that the receiver is perfectly synchronized to the received signal, i.e., we have assumed that the receiver's bandpass filter is hopping in perfect step with the received signal. In making this assumption we have taken for granted what is in fact the most critical problem in the design of spread spectrum systems. In this section we qualitatively examine some of the most important issues relating to the synchronization of FH spread spectrum signals, and the impact of these synchronization issues on the overall network performance.

The synchronization process would be considerably easier if the receiver were able to examine the entire hopping bandwidth simultaneously, as is done in some systems. In that case, one could simply store all of the received signals across the hopping bandwidth, and then synchronize/detect/decode after receiving a sufficient amount of data. In fact one would be able to monitor a number of signals simultaneously if such a receiver were available (see e.g., [78]). At HF it is quite difficult, however, to build a receiver that is capable of monitoring such a large bandwidth. A major difficulty results from the fact that the HF band is characterized by large interference levels that may overload the receiver, thus preventing the correct detection even of signals in other parts of the hopping bandwidth. Also, we note that the hopping bandwidth at HF (of 1 to 5 MHz) is a significant fraction of the center frequency; therefore, such a receiver would be difficult to implement. We have assumed in our studies that the receiver's bandwidth is equal to that of a single frequency bin, permitting only one frequency bin to be examined per hop interval. Therefore, synchronization must be acquired before detection and decoding can take place. Any data received before synchronization has been acquired is lost.

The receiver must, first of all, know the pseudonoise (PN) sequence (the hopping pattern in FH systems) that was used to spread the signal. It must then be able to synchronize the hopping pattern of its narrowband filter with that of the received signal. This resolution must be achieved to within a sufficiently small fraction of the dwell time. The relative difficulty of the synchronization process is related to the timing uncertainty, expressed as a multiple of the dwell time. Thus, higher hopping rates result in a greater uncertainty, because it may be necessary to attempt synchronization at all possible values of relative time delay. The difficulty of synchronization also depends upon the channel characteristics, most notably noise, jamming, other-user interference, and fading, each of which can disrupt the acquisition and maintenance of synchronization.

In direct sequence (DS), systems one must synchronize to within a chip interval. In practical applications, the chip duration in DS systems is considerably smaller than the hopping dwell time in FH systems. For a given timing uncertainty (e.g., in milliseconds) the timing uncertainty in DS systems (measured as a multiple of the chip duration) is thus considerably greater than that in FH systems (measured as a multiple of the dwell time).

Performance measures used to evaluate synchronization schemes include the expected time required to achieve synchronization, the probability of failure to synchronize with the desired signal, and the probability of false alarms (i.e., the probability of falsely synchronizing to something other than the desired signal).

Rapid reliable synchronization is especially important in the packet communication network environment because individual transmissions are of short duration. An inefficient synchronization scheme would therefore result in having a large fraction of the transmitted bit stream devoted to the synchronization process.

The two phases of the synchronization process are acquisition and tracking. In this section we are more concerned with acquisition, which is the resolution of time uncertainty to less than the dwell time. Tracking is the maintenance of synchronization to within a sufficiently small fraction of the dwell time once acquisition has been accomplished.

11.1 Synchronization Schemes

There are a number of schemes for the acquisition of synchronization in FH systems. The simplest is a *move ahead and wait* scheme in which the receiver advances its frequency bin to a state corresponding to the time of day plus the maximum possible relative timing uncertainty between the transmitting platform and itself. The receiver then waits until the energy received in this frequency bin exceeds a threshold, at which point the hopping process at the receiver is initiated.

The main disadvantage of such a scheme is that false alarms can occur as a result of noise (including jamming or spoofing) or other-user interference. If, after initiating the hopping process, the receiver decides that a false alarm has occurred, the receiver stops hopping. The entire acquisition process could then be repeated by advancing the receiver's frequency bin to a state corresponding to the new time of day plus maximum relative timing uncertainty. However, it is clear that the timing uncertainty region has already been reduced during this first, albeit unsuccessful, attempt at acquisition. Thus, it is sufficient to sit and wait in the last frequency bin visited until a signal is detected, at which point hopping is resumed. In using such a strategy one must recognize the possibility that the hop that is being waited for might not be detected, as a result of fading, antenna characteristics, or other factors. Therefore, if synchronization is not acquired after examining the entire uncertainty region, the complete acquisition process is in fact started anew by advancing the receiver to the new time of day plus maximum relative timing uncertainty as discussed above.

We now discuss a modified move ahead and wait scheme that is more resistant to false alarms. As above, the search is started at the frequency bin corresponding to the maximum time uncertainty; however, the receiver dwells at this frequency throughout the entire timing uncertainty window. There are several possible situations:

- *No signal is detected:* the search is then repeated based on the new time of day.
- *One signal is detected:* the receiver adjusts its clock based on the time of arrival of the signal and then starts hopping. If it is later determined that this was a false alarm, the acquisition process is repeated.
- *Two or more signals are detected:* at least one of these is a false alarm. Therefore, the acquisition process must be repeated, perhaps several times. Even if two or more signals are detected in every search interval, it may be possible to acquire synchronization on the basis of the observations made during the collection of search intervals, since the false alarms will normally be uncorrelated with the hopping pattern of the desired signal.*

A similar technique for acquisition is the *stepped serial search* scheme [79,80]. Unlike the move ahead and wait scheme, the receiver is in a hopping mode at all times. Several (M) consecutive hop intervals are examined. If all of them are occupied, the decision is made that the desired signal is present and the receiver continues hopping. If not, the relative time delay at the receiver's hopper is slipped by some increment (perhaps half of the dwell time) at each attempt at synchronization throughout the uncertainty region until acquisition has been achieved.

As in the move ahead and wait scheme, it is still possible for false alarms to occur as a result of either jamming, noise, or other-user interference. The false-alarm probability decreases as the number of consecutive hops examined (M) increases. However, the time required for synchronization increases as M increases. Fading, on the other hand, can result in missed hops. In a fading environment rather than to require energy in all M hop positions, one might set a threshold at some lower number of hops being occupied. We thus have a tradeoff between false alarms and missed acquisition.

*In some applications there may actually be two or more signals simultaneously using the same hopping pattern. If their relative delay is greater than one hop interval, then it may be possible for the receiver to selectively choose which of these it wishes to monitor; this is the capture property that is discussed in Section 10.3.5.

A third type of acquisiton scheme is the *matched filter* [79,80], a passive correlator that searches the received sequence in real time. As in the stepped serial acquisition scheme, M consecutive frequencies are searched. However, in the matched filter they are searched in parallel through the use of M band pass filters and delay lines. The use of a matched filter acquisition scheme can result in greatly improved performance in many applications; however, the requirement of M band pass filters is not consistent with our operational constraint of being able to monitor only a single frequency bin in each hop.

Of the schemes we have considered, the move ahead and wait (both the basic and modified schemes) and the stepped serial search techniques are possible candidates for the HF ITF Network. As compared with the stepped serial search scheme, the basic move ahead and wait scheme is simpler to implement, and it permits a faster search through the uncertainty region. However, it is more easily disrupted by interference and fading. Like the stepped serial search, the modified move ahead and wait scheme bases its initial decision that synchronization has been acquired on several hops. Both of these schemes are therefore more robust to interference than the basic move ahead and wait scheme, under which hopping begins as soon as energy is detected in one particular frequency bin. It is still too early to determine the best acquisition scheme for the HF ITF Network. A detailed analysis of the relative performance of these schemes, or possibly hybrids of them, in the HF ITF environment must be performed before a definitive choice of synchronization scheme can be established.

11.2 Timing Uncertainty and its Impact on Synchronization

We can consider three classes of acquisition problems that have been defined in terms of the degree of timing uncertainty, as discussed in [79]:

- initial acquisition
- *coarse* reacquisition
- *fine* reacquisition.

Initial acquisition is the establishment of timing synchronization from total timing uncertainty. It is thus the most difficult and time-consuming type of acquisition. Fine reacquisition, on the other hand, is performed where timing uncertainty is small (perhaps up to several hops). Typical examples characteristic of fine reacquisition include TDMA operation (fine reacquisition may be needed as the receiver switches from one signal to another) as well as reacquisition after synchronization is disrupted by noise, jamming, or other-user interference. The intermediate case of coarse reacquisition would apply when platforms have not communicated with each other for some period of time, but do have some knowledge of their relative timing delays. A typical example of coarse reacquisition might be that which is associated with the establishment of network structure by means of the execution of the Linked Cluster Algorithm.

The timing uncertainty of the received signal arises from two basic sources:

- absolute time uncertainty,
- propagation delay uncertainty between platforms.

Highly accurate clocks are needed on each platform to provide a stable timing reference. Their use is expected to maintain absolute time accuracy to within a millisecond on each platform throughout the duration of any military operation. (Although most naval platforms are not presently equipped with such accurate clocks, it is reasonable to assume that they would be by the time the HF ITF Network is implemented.) The relative time uncertainty between the clocks on two platforms is thus expected to be no greater than 2 ms. The propagation paths in the HF ITF Network will vary from near zero to as great as 500 km, resulting in propagation delays between 0 and 1.67 ms. The total time uncertainty between two network platforms at network initialization should thus be less than about 4 ms. Initial

acquisition, as defined above, would therefore be unnecessary. At worst, one would have to achieve coarse reacquisition.

As an example let us consider a hopping rate of 1000 hops/s, a timing uncertainty of 4 ms (perhaps corresponding to network initialization, as we have just discussed), and a cell duration of 0.5 ms (half the dwell time). The timing uncertainty is thus 8 cells. Increasing the hopping rate to 2400 hops/s, while reducing the cell duration proportionately to 0.2 ms, would raise the uncertainty to about 20 cells. Decreasing the hopping rate to 120 hops per second, while increasing the cell duration to 4 ms, would decrease the uncertainty to 1 cell.

The establishment of synchronization between a pair of users reduces the timing uncertainty between them to less than a dwell time. If both platforms have sufficiently accurate clocks (which is a reasonable assumption) and if they remain stationary, then they can remain synchronized even after periods of silence. In general, however, platform motion results in changing propagation delays; therefore, timing uncertainty is increased and synchronization may have to be reacquired. The timing uncertainty that results from an interval of silence is proportional to the change in propagation path length during that interval. For the platform velocities characteristic of ITF surface platforms and the hopping rates anticipated in future HF systems it is expected that the timing uncertainty accrued between two consecutive channel accesses in most TDMA applications will be less than a cell duration. In such cases fine reacquisition would appear to be not very troublesome.

Not all HF ITF Network traffic will fall in this category, however. Greater time uncertainties can occur if the time interval between successive transmissions is excessively long, particularly when airborne platforms are involved (i.e., as a result of their high velocity). Furthermore, interference resulting from background noise, other-user interference, or jamming can severely impact on the synchronization process, as we discuss in the next subsection.

11.3 Synchronization in an Adverse Environment

As we have noted throughout this report, the HF ITF environment is characterized by high levels of noise, other-user interference, and jamming. Each of these forms of interference can impact severely on the synchronization process.

We have already noted that background noise, jamming, spoofing, or other-user interference can result in false alarms not only in move ahead and wait schemes but also in the more robust stepped serial search schemes. Similarly, fading can result in failure to acquire synchronization.

Synchronization preambles, during which no data are sent, are often used. The time required to acquire synchronization, and in fact the probability of being able to do so, are variable and depend on the actual value of timing delay, the search strategy, and the channel environment. It is possible in some cases to acquire synchronization on data alone, provided that clocks are sufficiently accurate and timing uncertainties are sufficiently small, as we shall soon discuss. The performance of a stepped serial acquisition scheme in a fading jammed multiple user environment is analyzed in [80].

Systems are especially vulnerable to jamming during the acquisition phase, because failure to acquire synchronization can result in unacceptable delays or even loss of entire messages. An effective jamming strategy in many cases would thus be for the jammer to concentrate its energy by pulse jamming during the synchronization preamble, where it can do the most damage.

11.4 Synchronization in a Networking Environment

The impact of synchronization issues in a networking environment can best be examined if we consider first the *nonnetworking* case of a single source that transmits, perhaps intermittently, large

amounts of data to a single destination. In such cases, even a relatively long synchronization delay will have little impact on performance if it represents a small overhead in terms of the total bit stream that is transmitted. The probability that synchronization is eventually acquired can be a more significant performance measure than is the expected time needed to acquire synchronization.

In a packet-switched network environment, however, synchronization must be achieved for each packet. As in the single-source single-destination case, any symbol that is transmitted before synchronization has been acquired, is lost. Synchronization preambles during which no data are sent are often used, as we have discussed earlier. If packet lengths are small, the synchronization preamble can represent considerable overhead, and therefore significant reduction in the achievable data throughput. As we have already noted, the time required to acquire synchronization is variable. The choice of the length of the synchronization preamble thus represents a tradeoff between the probability of acquiring synchronization and the overhead involved in the synchronization process.

In a network a large number of codes may be used. In Section 10.3.4 we discussed the use of FH patterns associated with the transmitter vs those associated with the receiver, as well as network-wide (or cluster-wide) common codes. We assume that each platform knows all of the codes in use in the network. The time delay (perhaps up to tens of milliseconds) that may be incurred as a transmitting or receiving platform changes from one code to another is related to the synchronization problem. The code changing process may involve the loading of a new sequence into a shift register. A brute force approach to this problem would be to simply add a guard time at each slot. By doing so, a considerable overhead might result if the time required to change codes is a significant fraction of the packet length; however, there are better ways to solve this problem. For example, in a TDMA application in which all users monitor all transmissions, a single common (e.g., cluster-wide) code can be used, thus eliminating the need to change codes. In more general applications the use of more than one code can usually not be avoided, but in some cases it may be possible to establish transmission schedules that minimize the need to use different codes in consecutive time slots. Future efforts to develop multiple access protocols for the HF ITF Network will take into consideration the delays that can result when switching from one code to another.

A very desirable feature in future HF receivers will be the ability to rapidly change from one code to another (e.g., in less than one hop duration time) while maintaining timing synchronization. Such a capability would permit the flexible use of a variety of code assignment schemes without being burdened by the constraints we have just discussed.

11.4.1 Network Reorganization

Network reorganization by executing the Linked Cluster Algorithm (LCA) is a form of network initialization. As discussed in Section 1, the LCA consists of two TDMA frames during which the network platforms determine who their neighbors are, exchange connectivity information, and establish clusters. Although timing uncertainty among certain pairs of users may be small, new communication links among previously unconnected platforms may have to be established. The maximum timing uncertainty thus corresponds roughly to that of maximum clock uncertainty plus maximum propagation delay, i.e., the 4 ms value discussed earlier.

The information exchanged during the network reorganization phase is of critical importance, because it impacts directly on the structure of the network until the next reorganization. Failure of platforms (within communication range of each other) to acquire synchronization during the reorganization phase is thus usually a much more serious problem than failure to acquire synchronization during normal network communication. In the latter case it may be possible to retransmit the affected packet at a later time, perhaps upon failure to receive an acknowledgment following some timeout interval. Acquisition with very high probability requires either the use of a sufficiently long synchronization preamble or sufficient redundancy in the transmitted data, both of which can result in high overhead.

Future studies will address quantitatively the probability of acquisition as a function of preamble size and interference levels, with special attention given to the network reorganization phase. We note that the use of a special packet size for the reorganization phase may be unadvisable, because it would inform an adversary that a reorganization is taking place. The appearance of normal communication should be presented during the reorganization process to avoid a jamming attack designed specifically to disrupt the reorganization. Knowledge of when the reorganization takes place may be used to develop a jamming strategy that disrupts synchronization by concentrating the jamming energy in the synchronization preamble portion of each time slot.

We note that although the TDMA operation during the execution of the LCA is markedly different from normal network operation, when several platforms normally transmit simultaneously, it would be possible for platforms to transmit dummy messages to mask the reorganization operation by giving the appearance of normal network operation. The hopping patterns used for these messages could be chosen so that they do not interfere with that of the desired signals. Note that we assume that network platforms are capable of simultaneous transmission and reception (as long as they do so in frequency bins that are sufficiently far apart) and so such a scheme would, in fact, be feasible.

It may be advisable to run one or more TDMA synchronization frames prior to the actual execution of the LCA to increase the probability that synchronization is acquired. During each such frame the timing uncertainty between a pair of users is reduced. Alternatively, it may be possible to do some sort of probing operation during normal network operation to permit the discovery of new connectivities as they arise, and then to establish synchronization. It would be useful for platforms to maintain connectivity (which in a FH system requires maintaining synchronization) with as many other platforms as possible, even if such links are not part of the primary network (i.e., the subset of links connecting platforms to their cluster heads as well as cluster heads to each other via gateways). To do so facilitates not only future network reorganization, but the establishment of auxiliary links as well.

11.4.2 TDMA and Other Contention-Free Applications

The use of synchronization preambles is not always necessary in packet switched systems. In TDMA applications the timing uncertainty accrued between successive channel accesses will often be less than a dwell time. In these cases data can be transmitted throughout the packet duration. TDMA operation, however, assumes that the participating platforms have already acquired synchronization with respect to each other, and are able to maintain synchronization during silent periods between transmissions. These silent periods, as we have noted, must be sufficiently short so that the timing uncertainty remains less than a dwell time. Typically, the receiver will store a list of the relative timing delays associated with each of the transmitting platforms, and will delay or advance its hopping pattern depending upon whom it is listening to, without having to execute a complete search procedure. In general, the entire population of users can monitor the transmissions of each of its neighbors in this manner. If the receiving platforms do not store the relative time delay of each transmitting platform, then some sort of acquisition scheme will be necessary. In such cases the use of either a preamble or increased redundancy is needed.

Similarly, the timing uncertainty may also be small when contention-free links are used, i.e., links in which the receiver knows the identity of the transmitter, and can therefore know its relative timing delay sufficiently accurately. However, there is still other-user interference in the form of frequency hits when using contention-free FH signaling, as we have discussed in Section 10.3.3.

As a result of the lack of need for a synchronization preamble, the packet is no more vulnerable to jamming in any one portion than in another. Furthermore, the lack of need for a synchronization preamble allows the use of relatively short packets, and possibly the use of minipackets as are used in some reservation-based multiple access protocols [73].

The use of TDMA and other contention-free channel access schemes can therefore greatly simplify the synchronization process, provided that at each network platform knowledge is maintained of the relative delays with respect to the other platforms within communication range.

11.4.3 Contention-Based Channel Access Schemes

Two major features distinguish contention-based channel access schemes from those that are contention-free. The first is the possibility for two or more users to transmit to the same destination(s) simultaneously. The second is that since the destination does not know who is transmitting, it cannot adjust its own relative timing to take advantage of previously acquired synchronization. Two distinct situations may arise. In the broadcast case, packets are intended to be received by multiple destinations. Therefore, the timing uncertainty is at best equal to the propagation delay uncertainty; it is normally greater because of the difficulty in synchronizing accurately the absolute time measured by the clocks on each platform. In fact, the timing uncertainties at each channel access may be comparable to those experienced at network initialization. The use of synchronization preambles and the execution of an acquisition search procedure are therefore necessary in this case.

In the second case we consider a single destination. If a transmitting platform knows the relative timing delay between itself and the destination, it can advance or delay its transmission accordingly to compensate for relative time differences. In such cases a synchronization preamble may not be absolutely necessary.

Synchronization in a contention-based multiple-user environment is especially difficult when interference levels, which result from other users, jammers, and background noise, are high. Unlike the case of network reorganization, during which a single user transmits in each TDMA slot, a number of task force members may be transmitting simultaneously. Regardless of whether they are attempting to communicate with the same destination, they share the same hopping bandwidth and thus can interfere with each other.

The difficulty of acquiring synchronization is an argument for using contention-free channel access schemes whenever possible, to thereby reduce the difficulty of synchronizing with small packets in high interference environments. The use of contention-free channel access schemes does not limit one to ordinary TDMA, since adaptive TDMA-based schemes or reservation schemes (which are more suitable for bursty traffic) that use a TDMA reservation channel can also be used.

One can not completely escape the need for contention-based channel access schemes, however, because of the changing communication requirements throughout the network. For example, platform motion results in changing connectivities, and thus the need to establish new links. Also, some bursty traffic may be too time critical for a reservation scheme. Furthermore, certain platforms may transition from a radio-silent state to a transmitting state, thereby necessitating the use of some procedure (including a synchronization mechanism) to accommodate their entry into the network.

11.5 Cryptographic Considerations

Security considerations fall into two major areas, i.e., transmission security (TRANSEC) and communication security (COMSEC). Thus far in this report we have considered only TRANSEC, which refers to the security issues related to waveform design, but not to the protection of the specific contents of the waveform. For example, TRANSEC includes the use of spread spectrum signaling to avoid interception. TRANSEC also includes the use of techniques that make it difficult for an adversary to gain knowledge about a system by observing traffic patterns.

COMSEC, on the other hand, refers to the use of cryptographic techniques to make a message unintelligible to an interceptor, as well as to verify the identity of the originator of a message. In this subsection we discuss conceptually the impact of cryptographic considerations on the synchronization process, although we do not discuss specific cryptographic schemes currently in use in military communication systems. Our study of cryptographic issues is still in its early stages.

Basically, encryption is an operation in which half of the binary symbols in a data stream are reversed in some unpredictable manner. This operation is accomplished by adding a pseudorandom sequence of 1's and 0's to the data stream on a symbol-by-symbol basis. Such a sequence may be generated by using a maximal length shift register with nonlinear output logic, similar to that used for the generation of secure FH patterns, as we have discussed in Section 10.3.2. Decryption recreates the original data stream by repeating the encryption operation on the received data stream. Thus, the receiver must know the sequence that was used to encrypt the data.

In many applications a crypto-synchronization operation is necessary to permit the receiver to recreate the sequence. For example, the receiver may know the mechanism used to generate the sequence (e.g., the shift register taps and nonlinear output logic), but not the initial contents of the shift register. A crypto-synchronization preamble is therefore needed in which this information is transferred to the receiver. An interceptor without knowledge of the shift register taps and nonlinear output logic will not be able to recreate the sequence on the basis of the preamble alone.

A detailed description of the crypto-synchronization process would be beyond the scope of this report. We do want to point out, however, the inadvisability of permitting more than one platform to use the same crypto sequence at any given time. The availability to an interceptor of two messages encrypted using the same pseudorandom sequence could in some cases provide enough information to compromise the encryption process. Thus, what is often done is to associate a different (time-varying) crypto-key variable with each user. For example, different users may be using different segments of a very long sequence. Crypto-synchronization is acquired when all ambiguity, including relative time delay, about the crypto sequence is removed, thus permitting its generation at the receiver. Some cryptographic equipment currently in use require a preamble of 50 to 60 error-free bits to acquire synchronization, a large number in the case of the HF channel. Delays of hundreds of milliseconds or greater may result, thereby putting a lower limit on packet size. Such overhead could not be tolerated in a packet-switched HF ITF Network. It is therefore necessary to develop HF communication equipment that can permit the implementation of more efficient crypto-synchronization schemes that are in fact consistent with the demands of a packet-switched network. We now discuss some of the features that would facilitate crypto-synchronization.

Much of the delay associated with crypto-synchronization could be eliminated if the crypto sequence were timed concurrently with the frequency hopping process. The acquisition of FH synchronization would then facilitate the acquisition of crypto-synchronization. In applications in which only a single user transmits at a time (e.g., TDMA), only a single crypto sequence would have to be stored, and acquisition of FH synchronization would in fact remove all ambiguity. A highly accurate clock would be needed, but we have already assumed its availability for the FH pattern generator in our discussion of FH synchronization.

In cases involving simultaneous transmissions by more than one platform, different crypto sequences (or possibly different segments of the same very long sequence) would have to be used. It should be possible to develop schemes in which the crypto key is fully determined by knowledge of the transmitter's identity plus time of day; the latter is determined accurately during the FH synchronization process. In applications involving contention transmission, the identity of the transmitter might be transmitted unencrypted. This identity information would still be protected by the FH operation and error control coding. The disadvantage of using a preamble (i.e., having a short and therefore vulnerable segment) does not apply, because the identity information can be located anywhere in the packet.

Unlike FH synchronization, where data cannot be received until synchronization has been acquired, the decryption operation does not have to be performed until the entire packet is received.

In this subsection we have noted that the requirements of crypto-synchronization can impose severe constraints on packet-switched network operation. It is expected, however, that the design of HF crypto equipment that takes advantage of the use of highly accurate clocks and the development of crypto schemes that are coordinated with the frequency hopping pattern can greatly reduce the impact of cryptographic requirements on network operation, thereby facilitating the use of the packet switching schemes that are being developed for the HF ITF Network.

11.6 Conclusions on Synchronization in the HF ITF Network

The acquisition and maintenance of synchronization is one of the most critical problems in FH systems. It is especially critical in packet-switched networking applications, because the time required for synchronization can result in considerable overhead, and therefore lowered throughput. Furthermore, failure to acquire synchronization with sufficient reliability can disrupt the control structure of the network.

The degree of the difficulty of synchronization depends on the timing uncertainty as well as on the channel's interference, and possibly fading, characteristics. The use of low hopping rates makes synchronization easier, but results in reduced multipath protection, and, more importantly, reduced AJ capability against repeater jammers. Highly accurate clocks are needed on each platform to provide a stable timing reference. It is expected that such clocks will be available on naval platforms by the time the HF ITF Network is implemented. Their availability will greatly facilitate the synchronization process, primarily during the network initialization (reorganization) phase.

We have noted the advisability of using contention-free channel access schemes whenever possible. Doing so results in faster and more reliable acquisition with lower overhead, especially if each network platform maintains knowledge of its relative timing delay with respect to every other platform within communication range. A quantitative study is needed, however, to better assess the tradeoffs between different types of acquisition schemes and their impact on the performance of multiple access schemes for the HF ITF Network. The effects of other-user interference, noise, jamming, and fading will be investigated. Important considerations include the dependence of performance on hopping rate, packet size coding scheme, and whether or not relative timing information is available.

The impact of crypto-synchronization on network operation will be investigated further. We have noted that the difficulty and overhead associated with crypto-synchronization can be greatly reduced through the use of highly accurate clocks and crypto sequences that are synchronized with the hopping patterns. Further development of HF communication equipment is needed, however, to permit the implementation of such schemes.

12.0 CONCLUSIONS AND FUTURE PLANS

12.1 Summary and Conclusions

The three major areas that are being investigated in the design of the HF ITF Network are:

a. The architectural organization of the network;
b. The signaling issues (i.e., waveform and coding considerations); and
c. The multiple access protocols for the sharing of communication resources.

An understanding of each of these areas is vital to the design of the network. Furthermore, as our studies have progressed, we have developed a better understanding of the relationships among these areas.

In a preliminary report [1] the problems in the design of an HF ITF Network were identified and reviewed. In a later report [2] a preliminary system concept for the HF ITF Network was presented; this preliminary system concept addressed each of the above areas. Separate reports have discussed in detail the architectural structure and organization process of the network [3] and the ability of the network to reconfigure itself in response to jamming [4]. A future report will be devoted to multiple access protocols.

In the present report we have addressed the signaling issues that arise in the design of the HF ITF Network in greater detail than they were presented in [2]. In particular, we have demonstrated the relationships among signaling, network organization, and multiple access.

The ITF Network will consist of various platforms with markedly different characteristics, and it is expected to support the traffic requirements of many diverse scenarios. The use of the HF groundwave medium is dictated by its natural survivability properties in postnuclear detonation environments and by its ELOS communication range. At the same time HF suffers from time-variable and unpredictable fading and self-interference, and as a result it is extremely difficult to obtain an accurate model for the HF channel. Consequently, the ITF Network must exhibit a high degree of robustness with respect to channel behavior uncertainties both in terms of its organization and in terms of the signaling used. Furthermore, equipment limitations place additional important constraints on the network.

It is necessary to use spread spectrum signaling to provide protection from jamming. We have recommended frequency hopping (FH) as the spectrum spreading mechanism and noncoherent frequency shift keying (FSK) as the modulation technique.

Bit error rate (BER) performance under worst case partial band (WCPB) noise jamming has been evaluated as a function of the equivalent signal to noise ratio for FH-MFSK signaling. Performance tradeoffs were developed for several alphabet sizes, many values of diversity, several coding schemes, both hard and soft decision receivers, known or unknown jammer state, and for nonfading as well as Rayleigh fading channels. From these results we have concluded that coding and/or diversity are essential for jamming resistance in the HF ITF Network, and that modest to good AJ capability is in fact achievable for most links in the network through the use of convolutional coding with FH-MFSK signaling. A communication range model for jammed ITF environments has been developed by using a model for HF groundwave propagation loss in conjunction with the results for the jammer to signal power ratio that can be tolerated at the receiver.

In considering multipoint or network communication the signaling issues become considerably more complex than in the single link (or point-to-point) case because of the potential interference among the users. In a wideband architecture, such as that envisioned for the HF ITF Network, the use of spread spectrum signaling provides an inherent natural means of multiplexing the different users with acceptable levels of interference. This is achieved via code division multiple access (CDMA) techniques in which, for the proposed choice of frequency hopping, each FH pattern corresponds to a distinct code. The use of CDMA techniques alone, however, is not sufficient to solve the network's multiple access problems. It is necessary to develop techniques to share the network's communication resources jointly in the time domain and code domain. In this report we have examined the issues relating to the sharing of a wideband FH channel on a CDMA basis, including factors relating to the development of multiple access protocols for FH-CDMA channels.

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GLOSSARY

ABRC	Acknowledgment Based Retransmission Control
AJ	Antijam
ARQ	Automatic Repeat Request
AWGN	Additive White Gaussian Noise
BER	Bit Error Rate
BFSK	Binary Frequency Shift Keying
BLOS	Beyond Line of Sight
BTMA	Busy Tone Multiple Access
CDMA	Code Division Multiple Access
CSMA	Carrier Sense Multiple Access
DPSK	Differential Phase Shift Keying
DQPSK	Differential Quadriphase Shift Keying
DS	Direct Sequence
ELOS	Extended Line of Sight
EMI	Electromagnetic Interference
FEC	Forward Error Control
FH	Frequency Hopping
FOT	Frequency of Optimum Transmission
FSK	Frequency Shift Keying
HF	High Frequency
HFIP	High Frequency Improvement Program
HF ITF	High-Frequency Intratask Force
J.S.I.	Jammer State Information
LCA	Linked Cluster Algorithm
LFSR	Linear Feedback Shift Register
LPI	Low Probability of Intercept
LRI	Limited Range of Intercept
LUF	Lowest Usable Frequency
MFSK	M-ary Frequency Shift Keying
MUF	Maximum Usable Frequency
PN	Pseudonoise
RS	Reed-Solomon
TDMA	Time Division Multiple Access
WCPB	Worst Case Partial Band

Appendix A

RELATIVE PROPAGATION DELAYS OF GROUNDWAVE AND SKYWAVE SIGNALS

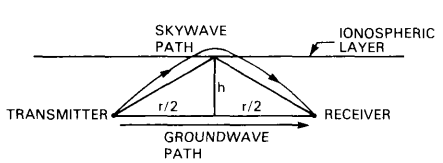
HF skywave propagation, which relies on the refraction of radio waves by the ionosphere, is normally used for Beyond Line of Sight (BLOS) communication at ranges of up to thousands of kilometers. We have noted in Section 2 that groundwave propagation will be used for most of the HF ITF Network communication traffic. However, skywave signals are often generated along with the desired groundwave signals, and in such cases they will appear at the receiver as interference in the form of multipath signals. In this appendix we discuss the role of the ionospheric layers in the reflection of HF skywave signals and the resulting relative propagation delays that may be expected between the desired groundwave signals and the interfering skywave multipath signals.

A1 THE IONOSPHERIC LAYERS

The ionosphere is a region in the upper atmosphere where free electrons are produced by the ionization effect of ultraviolet light and soft X-rays from the sun. There are four significant concentrations of free electrons, which are known as the D, E, F1, and F2 layers. The D layer (which, at an altitude of about 60 km, is the lowest of the layers) is mainly a daytime phenomenon. It acts primarily as an absorptive medium, rather than a reflector, in which lower frequencies are attenuated more than the higher frequencies. The other layers serve as reflectors. The lowest of these is the E layer, at an altitude between 90 and 110 km, which is present primarily during the daytime. The F1 layer, at about 150 km, and the F2 layer, at about 250 to 300 km, are responsible for the propagation at the longest ranges. At night the F1 and F2 layers merge into a single F layer. HF skywave signals consist of the superposition of reflections from one or more of these layers. The physical process of reflection is actually refraction, in which the signal is continuously refracted as it passes through the layer. An overview of HF skywave propagation is given in [A1], and more exhaustive treatments are found in [A2], and Chapter 9 of [A3].

A2 RELATIVE PROPAGATION DELAYS

The relative propagation delays between groundwave and skywave signals may be estimated by comparing the propagation path lengths of the two signals. We model single hop propagation paths for the ionospheric layers as shown in Fig. A1. The resulting relative propagation delays are then easily evaluated for communication ranges that are characteristic of a task force. In particular, we are interested in direct path groundwave communication ranges that can vary from near 0 to 500 km.



d = ADDITIONAL LENGTH OF SKYWAVE PATH
$= 2\sqrt{h^2 + (r/2)^2} - r$

Fig. A1 — Model for groundwave and skywave propagation paths

We note that we have neglected the complex nature of HF skywave propagation, e.g., the existence of *high* and *low* rays, as well as *ordinary* and *extraordinary* rays. The simplified model we use is sufficiently accurate to provide an estimate of relative propagation delays that will permit a qualitative evaluation of the effects of skywave multipath interference.

We define

h = height of the layer, and

r = source to destination groundwave path length.

The difference in the lengths of the skywave and groundwave propagation paths is therefore

$$d = 2\sqrt{h^2 + (r/2)^2} - r.$$

The resulting propagation delay difference is simply

$$t = d/300,$$

where t is expressed in milliseconds and d is expressed in kilometers.

In Fig. A2 the relative propagation delay (t) is plotted as a function of groundwave range for values of h corresponding to the ionospheric layers. Figure A3 shows the hopping rate that is required to ensure that the skywave component does not arrive until after the groundwave signal has hopped out of the frequency bin of interest. This hopping rate is simply the reciprocal of the relative propagation delay.

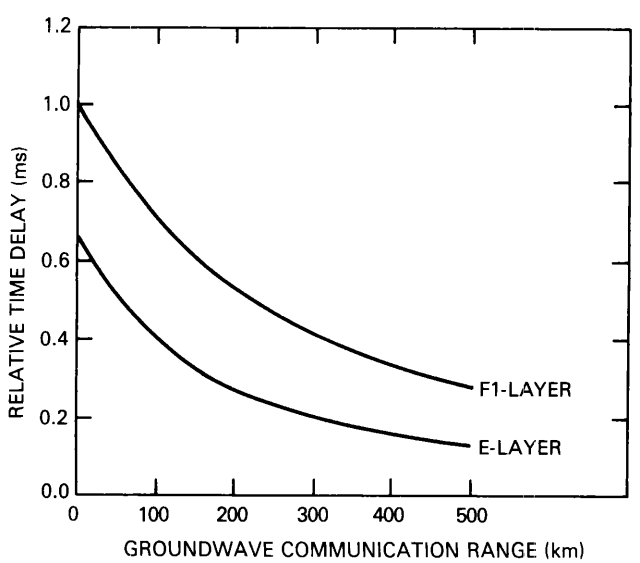


Fig. A2 — Relative time delay between groundwave and skywave signals

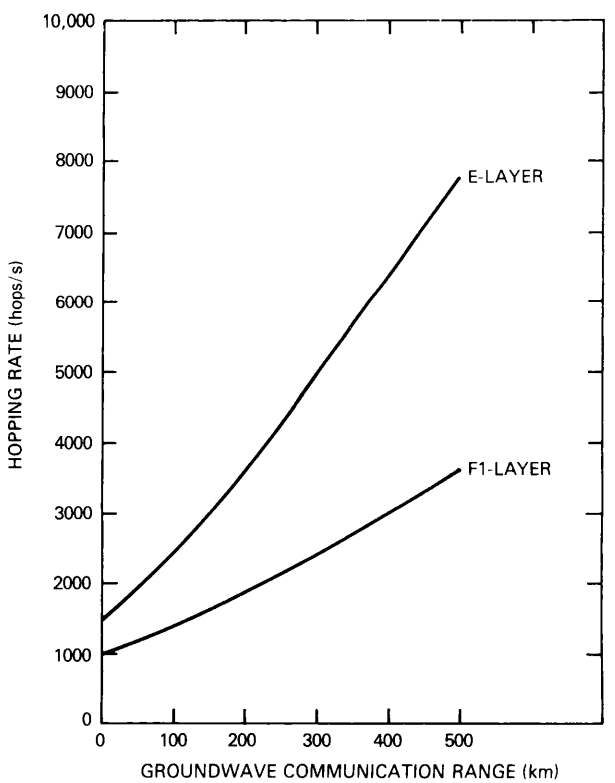


Fig. A3 — Hopping rate needed to avoid all skywave multipath interference

For example, a hopping rate of 2400 hops/s is sufficient to avoid all F1 layer (h = 150 km) reflections if the groundwave communication path length is 300 km or less. For the case of E layer (h = 100 km) reflections a hopping rate of 2400 hops/s is sufficient only if the communication range is less than 100 km. We note that the amplitude of E layer reflections is often small enough so that it does not pose serious problems; however, it can sometimes be quite significant and must therefore be considered. Thus, we can conclude that skywave multipath interference will often be present (provided of course that the transmission frequency is between the LUF and the MUF) when using practical frequency hopping rates, e.g., up to 2400 hops/s. We note, however, that robust modulation techniques (e.g., the use of noncoherent FSK, with either one symbol transmitted per hop or several symbols transmitted in parallel in each hop) used in conjunction with effective error control coding can often mitigate against the effects of multipath interference, even when the hopping rate is not fast enough to outhop such interference.

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Appendix B

CODING SCHEMES AND THEIR APPLICATION TO THE HF ITF NETWORK

Several classes of coding approaches may be considered for use in the HF ITF Network:

- Automatic Repeat Request (ARQ)
- Block codes
- Convolutional codes
- Concatenated codes

The selection of one or more of these coding schemes depends on our understanding of the channel disturbances. Each of these classes of schemes has its advantages and disadvantages, as we shall demonstrate in this appendix.

Excellent references on error control coding include Lin and Costello's recent textbook [B1], a more practical, less theoretical book by Clark and Cain [B2], and a survey article by Bhargava [B3].

B1 AUTOMATIC REPEAT REQUEST (ARQ) SCHEMES

Of the classes of coding schemes that we have listed, perhaps the simplest technique for burst error channels is to use error detection and retransmission. In this scheme the information is encoded by a parity check code. At the receiver, if the parity code is accepted, then the information bits in that block are also accepted as correct. If the parity code is rejected, the receiver instructs the transmitter to repeat the message. A variety of ARQ schemes have been investigated [B4,B1].

Although basic ARQ schemes are relatively easy to implement, they have several drawbacks. First, there is a need for a reliable transmitter/receiver feedback link. Secondly, a variety of buffering problems are introduced at the transmitter and receiver by the need to retain messages for retransmission. Also, under noisy channel conditions there will be many repeated messages leading to low throughput on the forward channel. Finally, the delay introduced by the retransmission process may not be tolerable in certain time-critical applications, including voice. We also note that retransmission would not be appropriate for certain classes of data, e.g., track reports that are frequently updated.

An ARQ scheme by itself would not be acceptable for use in the HF ITF Network. To achieve low BER, and thus high throughput under jammed or fading conditions, it would be necessary to maintain unrealistically high signal levels. However, the use of ARQ in combination with some forward error control (FEC) scheme is a possibility for some network services.

B2 BLOCK CODES

Block coding, or algebraic coding as it is sometimes referred to, is a technique that groups the incoming information bits into equal length sequences of k binary digits each. There are $M = 2^k$ different binary sequences of length k and the encoder denotes a code word for each. (We note that M-ary signaling can be considered as a simple form of block encoding without using any redundancy.) The block encoder chooses a code word of a given sequence length, n, which is called the block length.

Within the block length is a specific algebraic structure of k information bits and $n-k$ check (or parity) bits. Hence, a block code is sometimes referred to as an (n,k) code where the code rate is defined as $r = k/n$.

We are especially interested in the Reed-Solomon (RS) codes, which are a class of powerful burst error correcting block codes. Reed-Solomon codes are M-ary in nature, rather than binary, with a block length of M-1 M-ary symbols (although there are also extended RS codes with block lengths of M and $M+1$). We will sometimes refer to an M-ary Reed-Solomon symbol, which is the equivalent of $K = \log_2 M$ binary symbols, as a *byte.*

A Reed-Solomon $(M-1,v)$ code can correct *any* pattern of up to $t = (M-1-v)/2$ byte errors in a codeword. It is thus a *maximum distance separable code,* and as such it is considerably more powerful than binary (e.g., BCH) codes [B5]. It is important to note that, from the standpoint of codeword error probability, all byte errors are equivalent, regardless of the number of binary symbols within the byte that are in error. For this reason Reed-Solomon codes do not perform as well as BCH codes on random error channels, i.e., because a single binary symbol error in a Reed-Solomon codeword destroys the entire byte.

Virtually all codeword errors are detectable with Reed-Solomon codes, often permitting the subsequent retransmission of codewords that are received with uncorrectable errors. Such a system is an example of forward error control coding combined with ARQ. When such retransmission is possible, we can interpret the codeword error probability as a *codeword erasure probability.* The probability of an undetected codeword error is less than $1/t!$ [B5], where t is the number of correctable errors as defined above. This probability is less than 2×10^{-5} for the RS (31,15) code and less than 10^{-89} for the RS (255,127) code. Therefore, a system using Reed-Solomon coding with ARQ can have a resulting codeword error probability that is extremely low for sufficiently large values of t, because virtually all codewords that are originally unsuccessful will eventually be transmitted successfully. A useful performance measure is thus the channel throughput (i.e., the expected number of codewords that are correctly delivered per time unit), rather than the BER. In cases in which retransmission is not possible, one can evaluate not only the packet error probability, but also the BER; in addition one can detect, with high probability, which packets have errors and which are error free. In cases where channel traffic consists of streams of many codewords, it is sometimes desirable, depending on the characteristics of the disturbance process, to interleave the data among several codewords to randomize the occurrence of byte errors.

The use of Reed-Solomon coding is well suited for slow FH systems, i.e., systems in which several binary symbols, or one or more M-ary symbols, are transmitted in each hop. For example, one or more Reed-Solomon symbols can be transmitted in each hop. Packet sizes can be chosen so that the data of a packet can be encoded as a single Reed-Solomon codeword. Different Reed-Solomon M-ary alphabet sizes therefore result in different codeword sizes. Each packet consists of M-1 M-ary symbols. The use of quasi-orthogonal FH patterns permits the sharing of a wideband channel by a number of users on a CDMA basis, as is discussed in Section 10.3.3. FH multiple user systems are often asynchronous at the hop level, and so frequency hits (i.e., the transmission by two or more users in the same frequency bin simultaneously) can occur, thus destroying data. Errors caused by other-user interference will tend to occur in bunches, because hits will usually affect more than one binary symbol in a hop. Reed-Solomon codes, which are powerful burst error correcting codes, as we have just discussed, are thus a good choice to correct the errors caused by frequency hits.

It is often assumed that hits destroy each of the affected bytes. Considerable improvement can be obtained in a system that is capable of detecting which of the received symbols have been affected by hits, and of erasing the corresponding bytes, as is discussed in Section 10.3.3 and Appendix I.

Knowledge of which hops are corrupted by other-user interference or jamming is known as *side information.* In the general case in which both errors and erasures occur, a correct codeword decision can be made as long as the number of byte erasures plus twice the number of byte errors is not greater than $2t$.

We note that binary block codes, such as BCH codes, can also be considered for FH systems in which a single binary symbol is transmitted in each hop. Binary block codes do not, however, have the maximum distance separability property of Reed-Solomon codes, and thus will not be able to tolerate as large a percentage of symbol errors.

One could use Reed-Solomon codes in the case of a single binary symbol per hop, by grouping together several hops to form an M-ary RS symbol. However, such an approach would be appropriate only if disturbances normally affect a large part of the bandwidth simultaneously and for a duration of several hops, e.g., as with impulsive noise produced by lightning or other sources, or by pulsed broadband jamming, or possibly fading. For the case of other-user interference, there is no reason to group together symbols that experience independent disturbances. The grouping together of binary symbols thus artificially creates RS symbol errors (hence bursts of binary symbol errors) in place of only a single binary symbol error, and it would thus typically result in degraded performance.

In the past, a major drawback of the use of block codes, such as Reed-Solomon codes, was the complexity of the decoder. We note that, as a result of the availability of very large scale integration (VLSI) devices, a Reed-Solomon decoder would require little more than a single chip. Therefore, there is no significant difference in implementation complexity between block decoders and convolutional decoders.

B3 CONVOLUTIONAL CODES

Convolutional codes are different from block codes in that there is no separation of information bits into blocks of coded symbols. Also, the decoding algorithms are probabilistic in nature rather than algebraic. Encoding takes place by advancing the information bits, k at a time, through one or more shift registers. For every k input bits to the encoder, n channel symbols are created. The rate of the code, r, is simply the same as before ($r = k/n$ uncoded information bits per generated channel symbol). A variety of algorithms exists for decoding convolutional codes. Sequential and Viterbi decoding are particularly well suited to random error channels, whereas threshold decoding might be more useful on burst channels.

In Appendix F we discuss the performance that is achievable through the use of convolutional coding on broadband and worst case partial band (WCPB) noise jammed channels. The signaling scheme considered there is one in which one M-ary FSK symbol (where $M = 2$, 4, or 8) is transmitted per hop. The use of M-ary convolutional codes, which we discuss shortly in the subsection on concatenated coding, is assumed. As a result of the transmission of a single M-ary symbol per hop, M-ary symbol errors tend to be random rather than bursty. Burst errors can arise, however, as a result of disturbances that affect the entire hopping bandwidth, as we have discussed in the case of block codes. If two or more M-ary symbols are transmitted per hop, as is necessary in systems in which the hopping rate is less than the M-ary signaling rate, burst errors will also arise. There are a number of convolutional codes that are suitable for burst error correction, as discussed in [B1]; however, these codes cannot take advantage of the use of Viterbi or sequential decoding.

An alternative approach for handling burst errors is the use of interleaving, which spreads out bursts of errors so that they appear to be random (see e.g., [B1]). Standard decoding techniques for convolutional codes can then be used. Fortunately, the use of interleaving, while improving performance for burst error channels, does not normally degrade the performance in the presence of random errors. Some specific interleaving schemes are discussed in [B6]. In Appendixes E and F we implicitly

assume that there is ideal interleaving. Without such a scheme to randomize errors, a jammer might be able to induce a bursty error process, and thus higher BER, by using some form of pulse jamming. The degree of interleaving that is required depends on the statistics of the burst error process. To sufficiently randomize errors, symbols transmitted in the same hop must be separated by several constraint lengths by the deinterleaver. The degree of interleaving that can actually be achieved in packet switched systems is limited by the short packet length. Further study is needed to assess the affects of nonideal interleaving on the decoding of convolutional codes in the HF environment.

An advantage of convolutional coding is the ability to use soft decision receivers with these coding schemes. In contrast, block codes are usually limited to hard decisions, although side information can sometimes be introduced to erase unreliable symbols. In a soft decision decoder, greater weight can be given to symbols of greater amplitude (or in more sophisticated systems to symbols that are closer in amplitude to their expected value), whereas in a hard decision decoder all received symbols are treated equally and a binary decision is made on a symbol by symbol basis. The soft decision structure normally enhances the performance of convolutional codes over block codes. However, soft decision detection requires a maximum likelihood estimate of the channel characteristics, which is very difficult to obtain for the HF channel, particularly when it is jammed. In contrast, hard decision receivers do not require such knowledge of channel characteristics to function satisfactorily.

In the discussion of block codes, we noted that side information can be very useful in the decoding process. The side information was the knowledge of whether or not individual symbols are corrupted by other-user interference; all of the affected bytes were then erased. In the discussion of convolutional codes in Section 7, the type of side information that is assumed is *Jammer State Information (J.S.I.)*, i.e., knowledge of whether the individual symbols are affected by jamming on a partial band jammed channel. One does not lose all of the jammed symbols, but instead uses knowledge of whether or not the symbols are jammed to make a maximum likelihood estimate of the symbol. It is expected that practical receiver structures can come close to the performance of this ideal receiver. In Appendix H we outline a procedure, developed by Omura [B7], to evaluate the AJ performance of hard decision receivers, with or without *J.S.I.*, and thereby to demonstrate the degradation in performance that results from the use of nonideal receiver structures.

Convolutional codes are inherently designed for long data streams, rather than relatively short fixed length packets. They can be used in packet-switched applications, however; e.g., the encoder and decoder can be initialized (i.e., set to the zero state) for each new packet. To do so several bits of overhead are normally needed to set the system to the zero state; however, *tail-biting* schemes have recently been developed that do not require this overhead [B8].

With convolutional codes the bit error probability decreases as the code's constraint length increases. Although extremely low bit error rates can be achieved, provided that sufficient signal levels are present, it is not possible to verify that a block of data is error-free with a probability as high as that for Reed-Solomon codes. The accumulation of large path weights in the Viterbi decoding process can indicate that a block of data is likely to contain errors, however.

B4 CONCATENATED CODES

The basic concept of concatenated coding (for two levels) is to cascade two separate codes together. This combination of inner and outer codes with appropriate decoders can be thought of as a superchannel. Typically, the outer code might be a block code of length n, of which k are information bits. These output symbols are then further encoded by an inner encoder. As an example, the symbols may be fed into the inner coder K at a time, and then be appended by $N-K$ parity symbols, yielding an inner block length of N symbols. The resulting overall code has length Nn symbols with Kk information bits per concatenated code word, and the composite code rate is equivalent to:

$$r = r_i r_0 = kK/nN.$$

Many decoding algorithms can be envisioned for both inner and outer codes. For example, an outer (n,k) block code, or even a convolutional code, with M-ary signaling (inner code with no redundancy) can be viewed as a simplified concatenated code.

B4.1 M-ary Convolutional Coding

When M-ary signaling is used, a single M-ary symbol is generated for each $K = \log_2 M$ binary symbols that are input to the modulator.* M-ary convolutional coding can be viewed as a form of concatenated coding in which the inner code is simply M-ary signaling (with no redundancy) and the outer code is a binary convolutional code, as shown in Fig. B1. In this case, $r_i = K$ is the inner code rate. We therefore have,

$$r = (\log_2 M)\ r_0.$$

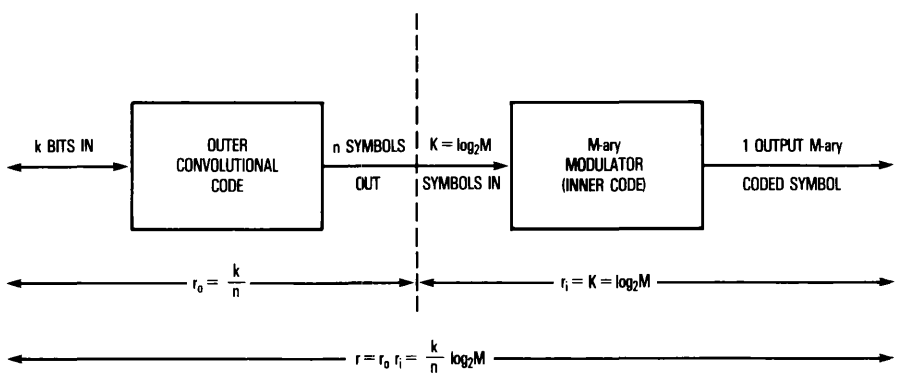

Fig. B1 — M-ary convolutional coding (concatenated coding model)

For example, for 8-ary signaling the generation of a single 8-ary symbol for every 3 binary input symbols corresponds to an inner code rate of $r_i = 3$. Since we are using convolutional codes, the outer code rate, r_0, is simply that of a binary convolutional encoder. In the special case of $r_0 = 1/K$, we will always have a composite code rate equal to $r = 1$, which corresponds to one M-ary symbol generated per uncoded data bit. In the examples we present in Section 7 for alphabet sizes of $M = 4$ and 8, this will indeed be the case, where our convolutional (outer) code rates will be $r_0 = 1/2$ and 1/3 respectively. Table B1 illustrates the outer convolutional code rate r_0, the inner M-ary code rate r_i, and composite code rate r for $M = 2$, 4, and 8.

Table B1 — Convolutional, M-ary, and Composite Code Rates

M	r_0	r_i	r
2	1/2	1	1/2
4	1/2	2	1
8	1/3	3	1

*This is *not* the same K used above in the definition of inner code rate.

In our examples for binary signaling we have assumed the use of Odenwalder's optimum rate 1/2 convolutional codes of constraint length 7 [B9]. In our M-ary signaling examples we have assumed the use of Trumpis's rate $r = 1$ codes that are optimized for the alphabet size [B10]. We note that the use of Odenwalder's binary codes on M-ary channels can result in better performance than the use of Trumpis's codes (i.e., lower required E_b/N_0 values for a specified BER). However, the use of binary codes on an M-ary channel necessitates the use of interleaving, which may be unnecessary if an M-ary code is used (provided that a single M-ary symbol is transmitted in each hop, the channel is memoryless, and the statistics of the jamming process are time-invariant). Furthermore, although a binary encoder and decoder may be easier to implement, it is harder to use soft decision information when the alphabet size is not matched to the modulation format.

B4.2 Other Concatenated Coding Schemes

The results that we present in Section 7 apply to the use of convolutional coding (outer code) and M-ary signaling (inner code). If desired, more sophisticated concatenated coding designs can be configured. In the past, a natural choice has been to concatenate Reed-Solomon outer codes with inner convolutional codes. The inner code cleans up the channel very well during random error events by using sequential or Viterbi decoding, but it has difficulty when burst errors are encountered. The RS outer code, having a burst error correcting capability, then attempts to further improve the error rate out of the inner decoder, which is characterized by a burst error process. Since both random and bursty error processes are expected in the HF ITF Network, such a scheme may provide improved performance over that of either Reed-Solomon or M-ary convolutional coding alone. Also, the Reed-Solomon outer code can help to verify the correct reception of a packet, as we have discussed in Section B.2.

As an example, for the AWGN channel, the coding gain at BER $= 10^{-5}$ is 4.4 dB (M=8) to 7.3 dB ($M = 128$) when using M-ary signaling and noncoherent detection. Incorporating further Reed-Solomon coding will improve performance by approximately another 2 to 3 dB [B2]. This suggests that M-ary signaling is quite effective on the AWGN channel with noncoherent detection, and that Reed-Solomon outer codes may be quite useful.

One could consider a concatenated coding scheme in which the inner code is designed to provide a *medium-quality* grade of service (e.g., BER $= 10^{-3}$), and the outer code improves performance to provide a *high-quality* grade of service (e.g., BER $= 10^{-5}$). In such cases one might use the outer code only when high-quality service is required. Alternatively, one might use the outer code only when it is needed, as dictated by channel characteristics. Such an approach is similar in principle to the adaptive use of increased diversity coupled with lower code rates in response to deteriorating channel conditions.

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Appendix C

WIDEBAND COMMUNICATION IN THE HF ITF ENVIRONMENT

Wideband signaling will be needed in the HF ITF Network to provide adequate protection from jamming. In this appendix we present the fundamentals of spread spectrum signaling and discuss specific considerations relating to the HF band, and in particular to the ITF environment. We provide the reasons for our recommendation of noncoherent frequency hopping (FH) as the spreading mechanism.

C1 FUNDAMENTALS OF SPREAD SPECTRUM SIGNALING

A spread spectrum system is one in which the transmitted signals are spread over a frequency band that is much wider than the minimum bandwidth required to transmit the information. For example, a baseband signal with a bandwidth of only a few kilohertz may be distributed over a band that is many megahertz wide. At the receiver, the wideband spread spectrum signals are remapped into the original information bandwidth and the signal is recovered.

There are four basic types of spread spectrum modulation. In the direct sequence (DS) method, the data to be transmitted modulates a higher rate pseudonoise (PN) code (or chip) sequence. The modulated PN sequence is then transmitted by a radio frequency (RF) carrier. In the frequency hopping (FH) mode the carrier shifts its frequency rapidly from one frequency bin to another using a pattern determined by a pseudorandom code sequence. In the pulsed-FM or "chirp" mode, a carrier is swept over a wide bandwidth during a given pulse interval. In the time hopping mode, time slots are assigned by means of a code sequence in the same manner that frequency hopping systems are assigned frequency bins. Hybrid systems that combine two or more of these techniques are often used.

The advantages of spread spectrum signaling, as enumerated in [C1,C2] are:

a. Selective addressing capability

b. Code division multiplexing for multiple access

c. Inherent message privacy/security

d. Low density power spectra for signal hiding

e. High resolution ranging

f. Interference rejection (protection from jamming and other-user interference as well as multipath propagation).

All of the properties listed above can be used advantageously in the HF ITF Network. The first three properties can be achieved through the use of quasi-orthogonal PN sequences, as we discuss shortly. The interception of spread spectrum messages is difficult because the power spectral density over any narrow frequency band is extremely low. An interceptor must therefore either integrate energy over a long time interval to detect the presence of a signal, or focus on some other signal characteristic that is more amenable to detection. Jamming resistance is achieved because it is difficult for a jammer to occupy the entire spread bandwidth at power levels that are sufficiently high to prevent signal detection.

Multipath signal components can be separated in DS systems by using correlation or matched filter techniques, and in FH systems by hopping at a sufficiently high rate. Alternatively, the use of robust signaling schemes (e.g., noncoherent FSK energy detectors with appropriate coding) permits correct signal reception despite multipath interference even at relatively slow hopping rates. The use of spread spectrum techniques in a packet radio (although not HF) environment is discussed in Kahn [C3]. We now discuss the basic features of DS and FH systems.

C1.1 Direct Sequence (DS) Systems

In a DS system, each receiver is equipped with a bank of matched filters, one for each particular PN sequence it is designated to receive. The PN sequences must have small cross-correlation. As a result, signals to which a filter is not matched appear as low power level noise-like interference. (If the family of PN sequences could be chosen to have zero cross-correlation, then the sequences would be orthogonal and this interference problem would be eliminated; however, some nonzero cross-correlation is always present in practice, and so only quasi-orthogonality is achievable.) We note that although in some applications short or periodic sequences may be used, security constraints in many military applications raise the need for *nonrepeating* sequences, that may in fact have periods of days, months, or years. The correlation we have discussed above thus refers to a partial correlation operation in most cases of interest.

Selective addressing is achieved by using a PN sequence that is associated with a particular receiver. Similarly, protection from interception is achieved by using sequences that are unknown to potential eavesdroppers. Code Division Multiple Access (CDMA) is of particular importance in the network environment. It takes advantage of the fact that a number of spread spectrum signals (generated using different PN sequences) can share the same wideband channel as long as the PN sequences used to generate them have good cross-correlation properties. The PN sequences then take on the role of codes. As the number of simultaneous transmissions (using different codes) increases, the interference level typically increases gradually. As a result graceful degradation, rather than the catastrophic collision that is characteristic of *time-domain* contention-based multiple access systems, is experienced. When all received signals are of equal power, the number of simultaneous signals that can be tolerated is approximately one tenth of the bandwidth expansion factor [C4]. The actual number varies and depends on factors such as signal to noise ratio, modulation scheme, coding, and acceptable bit error rate. When the signals are not of equal power, the number will generally be considerably less, as we discuss in Section C.2.4.

Resistance against jamming is achieved because the jammer's waveform is poorly correlated with the spreading waveform. The AJ capability of a spread spectrum system is directly proportional to its *processing gain,* which is the ratio between the bandwidth of the transmitted signal and the data rate in the information baseband channel. A DS system is, however, vulnerable to very large jamming signals, especially when they are near the center of the spread bandwidth. Consequently, realistic DS systems often include some technique for filtering out the largest interfering signals prior to, or as a consequence of, the correlation process. Such filtering is especially important at HF, because of the wide dynamic range of channel disturbances.

C1.2 Frequency Hopping (FH) Systems

In FH systems the transmitter hops from one frequency bin or subband to another, transmitting a narrow band signal at each hop. It is assumed that there is no *spectral splatter,* i.e., that the spectrum of the transmitted signal is limited to its narrow band. Such an assumption is reasonable, provided that appropriate pulse shaping techniques and frequency guard bands are used. (The effects of spectral splatter are discussed in [C5].) The hopping pattern is determined by a pseudorandom code that is analogous to the PN sequence used to achieve spectrum spreading in the DS case. The hopping can be performed over the continuum of the available spread bandwidth, or alternatively distinct frequency

bins can be established. In this report a frequency-slotted model is assumed, because it facilitates the formulation of the FH multiple access problem. It may be possible to achieve improved antijam and antiintercept performance by hopping over a continuum, however.

Again, selective addressing is achieved by the transmitter by its use of the intended receiver's unique FH pattern, and interception is made much more difficult when secure (unpredictable) FH patterns are used. Resistance against jamming is achieved because the jammer cannot put large amounts of energy throughout the entire frequency band, and because it cannot predict the FH pattern. (Either the hopping rate must be high enough so that repeat-back jamming is not feasible, or the signal design must be chosen so that it is not vulnerable to this jamming strategy.) A FH system will be able to tolerate the loss of a number of hops (which would correspond to fractions of packets) caused by jamming, if appropriate forward error correction coding and diversity are used. We should note that, although DS systems are vulnerable to jamming by a sufficiently large signal (especially when it is near the center of the spread frequency band), FH systems can be disrupted only if jamming energy is placed into the frequency bin at which the receiver is dwelling.

Code Division Multiple Access (CDMA) can be implemented in FH systems. The code now corresponds to the FH pattern. Systems employing CDMA operation are often asynchronous; therefore, it is possible for two or more users, even when using different hopping patterns, to transmit simultaneously in the same frequency bin, and thus to experience loss of data. The probability of such collisions (of fractions of packets in this case as compared with whole packets in the usual case of time-slotted multiple access schemes), known as *frequency hits,* increases as the number of users increases. The loss of data resulting from such hits can be handled in the same manner as the loss of data caused by jamming or fading, i.e., by using coding and/or diversity. The BER of a system with coding increases gradually as the number of users increases; this is another example of graceful degradation. The number of users that can simultaneously transmit in a FH channel while maintaining acceptable performance levels and related coding considerations are discussed in Section 10.3.3. Note that other-user interference occurs only when two or more signals simultaneously occupy the same frequency bin. The presence of strong signals in other frequency bins is not a problem. The concept of processing gain, therefore, is not useful when comparing narrowband interference power with signal power.

C1.3 Hybrid FH-DS Systems

Hybrid FH-DS signaling is often used to combine the best features of pure FH and DS. In such systems the signal transmitted at each hop is no longer narrowband, but instead it is DS spread (although by a much smaller factor than it would be in a pure DS system). In Section C2.7 we discuss some of the advantages and disadvantages of the use of hybrid FH-DS signaling.

C2 ON THE CHOICE OF SPECTRUM SPREADING TECHNIQUE

The choice of the spectrum spreading mechanism is one of the fundamental considerations in the design of a spread spectrum system. The three forms of spread spectrum signaling that we have considered are direct sequence (DS), frequency hopping (FH), and hybrid FH-DS.

The primary criteria by which an AJ spread spectrum system should be evaluated are its ability to provide AJ capability at acceptable data rates, its ability to function in the particular environment for which it has been designed (in our case the HF ITF environment), and the practicality of its implementation. We have also considered factors relating to Low Probability of Intercept (LPI), Limited Range of Intercept (LRI), and Limited Probability of Exploitation (LPE) modes of communication. By considering performance under each criterion we have concluded that, although each of the spreading mechanisms has its advantages, a pure FH system is the most practical choice for use in the HF ITF Network.

C2.1 Processing Gain — Bandwidth Considerations

As mentioned earlier, the AJ capability of a spread spectrum system is directly proportional to its *processing gain,* which is the ratio between the bandwidth of the transmitted spread spectrum signal and the data rate in the information baseband channel. In DS systems we are concerned with the instantaneous bandwidth, which is proportional to the chip rate of the pseudorandom noise (PN) code generator. In FH systems the bandwidth refers to the entire bandwidth over which the instantaneously narrowband signals are hopped; the processing gain is therefore proportional to the number of narrowband channels, or frequency bins, into which the signal may hop.* For reasons related to the Linked Cluster Architecture, as discussed in Section 1, we are considering the division of the HF channel into subbands of perhaps 2 to 5 MHz, over each of which the groundwave communication range is relatively constant (taking into account only the propagation characteristics of the HF medium, and not effects such as other-user interference or jamming). Although some signals could be spread over the entire HF band under some alternative network architecture, here we consider spreading over these subbands only. The instantaneous bandwidth of the DS system is limited by the coherence properties of the HF medium. The groundwave channel can support signal bandwidths that may be as great as 1 MHz; however, the maximum signal bandwidth suitable for skywave paths is normally at most 100 kHz, which, in many applications, would not provide adequate processing gain. (The use of channel equalizers could permit the use of higher instantaneous bandwidths [C6]; however, the resulting receiver would be considerably more complex, and therefore expensive.) Although we are assuming that the ITF Network will use primarily groundwave propagation, it would be inadvisable to employ a waveform that could not be used for skywave paths, since to do so would limit the flexibility of the network, and would ignore the consequences of mixed path propagation. Another significant advantage of FH is that it does not require a contiguous frequency band. Unlike DS, a FH signal can avoid portions of the channel that are corrupted by excessive noise (e.g., atmospheric noise, other-user interference, or jamming), that are characterized by poor propagation conditions, or that are dedicated to other users.

The difference in AJ performance between DS and FH systems is so small, under either additive white Gaussian noise or worst case partial band interference, that the choice between the two approaches must be based on other factors [C7-C9]. Based on the bandwidth considerations discussed here, FH systems have a distinct advantage over DS systems. Most of the other considerations that we address later also favor FH systems.

C2.2 Multipath Considerations

Since skywave signals are a source of multipath interference, which often results in the fading of the desired signal as well as in intersymbol interference, the spectrum spreading method should be chosen to minimize these effects. Spread spectrum signals have an inherent antimultipath capability. Direct Sequence (DS) receivers perform cross-correlation, thereby resolving the multipath components as distinct signals if the chip rate is higher than the reciprocal of the multipath time delay difference (provided that synchronization has been achieved and is not disrupted). The resolution normally achievable is roughly the chip duration, which is a fraction of the data bit duration. However, the frequency selective nature of the HF channel is apparent at the large instantaneous bandwidths characteristic of DS systems, often resulting in severe degradation of performance. The RAKE receiver concept [C10, Section 11-5] has been proposed to handle such distortions; however, its implementation is much more complex than that of a FH system. This is an additional reason for using FH rather than DS systems at HF.

*Alternatively, the processing gain is sometimes defined as being equal to the number of frequency bins (e.g., [C1]); either definition may be used as long as consistency is maintained in the analysis. In Appendix G we discuss in detail the interpretation of processing gain in FH systems.

Frequency Hopping (FH) systems can reduce multipath effects by using energy detectors (e.g., noncoherent FSK), and can in fact avoid such effects completely by hopping at a sufficiently high rate; multipath signals that arrive after the receiver's hopper has left a particular frequency bin cannot interfere with the desired signal. Since the FH signals are instantaneously narrowband, they suffer relatively little distortion caused by frequency selective fading as compared with DS signals. In fact, frequency selective fading can actually help the performance of FH systems that employ coding and/or diversity by assuring that the fading at each hop is independent.

We have noted that intersymbol interference and fading are the two harmful effects of multipath. The narrowband HF systems discussed in Section 4 avoid intersymbol interference through the use of guard times of approximately 4 ms during which no data are sent, thus resulting in reduced data rate. In a FH system, intersymbol interference can be avoided if the transmitted symbol(s) remain constant throughout the hop duration. No guard times are needed, although settling times of a fraction of a millisecond are to be expected at each hop. The requirement of constant symbol(s) does not necessarily limit us to one bit per hop, because symbol alphabets of higher order than binary can be used (e.g., M-ary FSK), and because multiple tones and therefore multiple M-ary symbols can be transmitted as in the narrowband case. Since we are assuming the use of energy detectors, the reception of a signal and its delayed replica within a single dwell time interval can result in either reinforcement or fading of the latter part of the received signal, depending on the relative phases of the skywave and groundwave signals. The relative delay between groundwave and skywave signals varies with frequency. Although some hops will experience fading, normally a majority will experience reinforcement. Error control coding (perhaps of rate 1/2) can adequately correct errors caused by such fading.

To guarantee a completely avoidance of the fading that results from skywave multipath interference, the receiver must hop out of the frequency bin before the skywave signal arrives. The hop rate necessary to do so depends on the relative delay between the signals propagating via groundwave and skywave modes. For multipath protection, it is best to hop at the fastest possible hop rate. Appendix A presents a discussion of the expected delays between groundwaves and skywaves in the HF ITF environment. It is shown that a hop rate of 2400 hops/s will be sufficient to avoid all F-layer skywave reflections at ranges of less than 300 km. Although E-layer reflections may still be present, they will usually be considerably smaller than the groundwave or F-layer skywave signals; therefore they will often not be troublesome, although fading may result in some cases. Coding and/or diversity will be required to achieve acceptable performance. If hopping rates are limited, because of implementation considerations, to at most a few hundred hops per second, then no further multipath protection would be provided by the FH operation beyond that which is achievable through the use of energy detectors and guard times, and we would have to rely more on coding and diversity to counter the effects of fading.

A possible alternative to a pure FH system is a hybrid FH-DS system, which can provide improved multipath rejection at relatively low hop rates, although the chip rate must be sufficiently high to combat multipath in this case. Other considerations, however, which we shall address later in this section, favor the pure FH waveform.

C2.3 Synchronization of PN Codes

Rapid acquisition of time synchronization is the most critical problem in the operation of spread spectrum systems. First, the receiver must know the PN sequence that was used to spread the signal. This replica of the spreading signal is then used to despread the waveform into its original baseband format. Time resolution must be achieved to within a sufficiently small fraction of the basic timing unit of the system. In DS systems the basic unit is the chip duration, whereas in FH systems it is the dwell time. For a given time uncertainty (e.g., in milliseconds) we can measure the synchronization uncertainty in terms of the basic timing unit. The chip duration in DS systems is considerably smaller

than the dwell time in FH systems. Therefore, the uncertainty encountered in DS systems is considerably more than that of FH systems, since an attempt at synchronization might be made for all possible values of time uncertainty. The acquisition problem for DS systems is therefore generally much more difficult than for FH systems. Furthermore, it is harder to maintain the synchronization of a DS system (e.g., because of jamming, platform motion, or natural causes such as signal fading) because of the finer time resolution that is involved. Maintenance of synchronization in the presence of jamming is essential.

Rapid synchronization is especially important in the packet communication network environment, since individual transmissions are of short duration. An inefficient synchronization scheme would therefore result in a large fraction of the transmitted bit stream being devoted to the synchronization process (i.e., high overhead). Synchronization issues relating specifically to FH systems are discussed in more detail in Section 11.

C2.4 Multiple Access Capability

The improved AJ performance of spread spectrum signals is gained at the expense of greatly increased bandwidth. Since the HF ITF Network will have to support communication among a large number of users, each wideband channel will have to be shared via CDMA. Although CDMA techniques can be implemented for either DS or FH systems, the former are highly susceptible to differences in received signal power levels (the *near-far* problem) while the latter are not. The CDMA capability of DS systems is based on the assumption that all signals are of (roughly) equal power. In such cases the response of the receiver to signals transmitted on codes other than its own is a relatively low noise-like output. However, when an undesired signal, is considerably greater in magnitude than the desired signal, the interference resulting from the undesired signal may make correct reception impossible [C11]. The deterioration of the performance of DS-CDMA as a result of the near-far problem is one of the greatest drawbacks to its use in the HF ITF Network. In contrast, in a FH system an undesired signal will result in interference only if it occupies simultaneously the same frequency bin as the desired signal. Relative power levels are immaterial as long as there is no spectral splatter.

C2.5 Robustness

By robustness we mean the ability of the system to operate satisfactorily in a variety of hostile environments. In general, FH systems are more robust than DS systems. The following five items have been addressed earlier in this section, but they merit repetition here as well:

a. FH systems are virtually immume to the *near-far* problem, and are therefore better suited than DS systems for multiple access applications.
b. PN code synchronization is easier to be acquired and more difficult to be disrupted in FH systems;
c. FH systems are less sensitive to the dispersive nature of HF channels;
d. FH systems do not require a contiguous bandwidth;
e. Wider bandwidths (and therefore greater processing gain) can be achieved with FH systems.

Also, when noncoherent signaling is used (e.g., noncoherent FSK, which is proposed for the HF ITF Network, as discussed in Appendix D), there are no carrier phase acquisition problems.

C2.6 LPI, LRI, and LPE Considerations

In our HF ITF Network studies, we have emphasized AJ performance rather than three other performance criteria that are also important in many military communication systems, i.e., Low Probability of Intercept (LPI), Limited Range of Intercept (LRI), and Limited Probability of Exploitation (LPE).

The use of LPI and LRI modes, which are used when covert communication is necessary, presents some constraints that are antagonistic to those of AJ systems. The primary purpose of the LPI and LRI modes is to hide signals (and often the communicating platforms themselves) in relatively quiet environments, rather than the successful delivery of a signal in a jammed environment. DS signals are normally considered better for LPI and LRI operation because the power spectral density of DS signals is typically buried in the channel noise, thereby making it difficult for an interceptor to detect the presence of signals. (We assume in both the DS and FH cases that the interceptor knows the basic signaling format, but not the PN spreading sequence.) In contrast, the presence of FH signals is more easily detected, because at every hop there is a significant signal spectral density in one frequency bin.

The use of secure FH patterns, a sufficient number of frequency bins, and sufficiently high hopping rates can normally prevent the eavesdropper from intercepting more than a small fraction of the hops. However, it is easier to detect the presence of FH signals than that of DS signals, a distinct disadvantage if platforms (e.g., submarines) wish to remain hidden.

In Section 2.2 we noted that use of frequency subbands that do not support skywave propagation is also an important consideration for LPI and LRI communication.

AJ performance is more important in the HF ITF Network than LPI or LRI considerations are, because the primary purpose of the HF ITF Network is to provide a communication capability that will enable the task force to survive attack, both physical and via jamming. Of primary importance is the reliable delivery of tracking data (which constitutes a large percentage of network traffic) to network platforms while operating in an environment characterized by high levels of hostile jamming. Also, surface platforms may have little need to remain hidden. Usually the location of the task force will be known anyway.

The use of LPI and LRI communication modes may not be necessary if LPE can be achieved. An LPE capability is the ability (with high probability) to prevent an interceptor from exploiting a received signal in any way (e.g., direction finding, signal identification, spoofing, traffic analysis), even when the signal strength at the interceptor is relatively large. The use of secure FH patterns, a sufficiently large number of frequency bins, and sufficiently high hopping rate contributes to LPE operation by permitting the interceptor to detect only a small fraction of the data stream. The use of cryptography will prevent the interceptor's exploitation of the signals, even if he can detect a significant part of the message. Furthermore, it is difficult to distinguish among the signals of the platforms that use the FH channel simultaneously, thus creating confusion and adding to the difficulty of signal exploitation.

The use of secure FH patterns also provides resistance against spoofing, i.e., the introduction of false messages into the system by an adversary, because knowledge of the FH pattern being monitored by the receiver is necessary for successful communication. Further resistance against spoofing is achievable through the use of cryptographic techniques that can permit the verification of the identity of the transmitter.

We have concluded that, although DS systems can perform better than FH systems in LPI and LRI modes, this criterion is not as critical as that of AJ performance in terms of overall network operation. Furthermore, some degree of LPE capability can in fact be achieved in FH systems, providing a measure of robustness with respect to threats of signal exploitation.

C2.7 FH vs DS and Hybrid FH-DS

The discussion in this section so far, especially as summarized under *Robustness* above, has indicated that FH is clearly preferable to DS as the spreading mechanism in the HF ITF Network. Torrieri [C12] has, in fact, concluded that FH is preferable in most military networks. The primary disadvantages of DS systems, which were noted to be near-far power differentials, synchronization difficulties,

and limited processing gain, far outweigh the advantages of DS in the areas of LPI and LRI communication modes. These disadvantages make the DS systems extremely inappropriate for the ITF environment.

The remaining candidate spreading mechanism is the hybrid FH-DS system. One of its advantages over pure DS is that much shorter PN sequences can be used in the hybrid system, since part of the interference rejection capability is provided by the frequency hopping. The use of shorter PN sequences results in easier signal acquisition than that for DS systems. Also, like FH systems, FH-DS systems are less dependent on the relative power levels of signals at the receiver than are DS systems. Furthermore, like FH systems, FH-DS systems take advantage of the entire hopping bandwidth and can therefore provide better AJ performance than can DS systems, whose signal bandwidth may be limited, particularly at HF.

Primarily we are interested in the potential advantages of FH-DS over pure FH systems, since pure DS has already been excluded from consideration. The only disadvantage of FH systems, that we noted earlier, is their vulnerability to the multipath signals that are generated when the hopping rate is too slow. The FH-DS system has a greater antimultipath capability than a pure FH system has, since it is able to resolve time essentially to the chip duration, rather than to the hop duration. The antimultipath capability of a FH-DS system is thus equivalent to that of a DS system with the same chip rate. Protection from multipath propagation can therefore be provided in FH-DS systems, even at low hopping rates. However, we have noted in Section C2.2 that acceptable performance can often be maintained in FH systems despite the presence of multipath signals when robust signaling techniques are used (e.g., the use of noncoherent FSK and signals that remain constant throughout the hop duration). Thus, the improved multipath capability that can be achieved in hybrid FH-DS systems may not be essential.

However, there are a number of disadvantages, in the use of the hybrid FH-DS system. For example, the equipment required for the hybrid system is more complex and less likely to be available. Also, the use of *somewhat spread* (e.g., 50 to 100 kHz) signals clearly reduces the total number of frequency bins available for CDMA operation, thereby increasing the occurrence of frequency hits. Although a FH-DS system is more tolerant of hits than a pure FH system is, the short length of the PN sequences used for DS spreading can only provide limited code separation, and therefore the near-far power differential problem may be significant when hits do occur.

As a result of the use of short PN sequences for the DS spreading, the synchronization process for hybrid FH-DS systems is considerably easier than it is for pure DS systems, and, in fact may not be much more difficult than it is for a pure FH system. A correlation operation could be performed at each hop to determine which symbol of a limited alphabet size was transmitted, thereby eliminating the need for a lengthy search procedure, because the DS spreading is timed concurrently with the frequency hopping process. The synchronization operation is aggravated, however, by the increased probability of other-user interference, again because of the reduced number of available frequency bins.

C2.7.1 Concluding Remarks

We acknowledge that we have not performed a definitive quantitative performance comparison between FH and hybrid FH-DS. The major advantages of FH-DS are that it provides the antimultipath capability of DS, while permitting the spread bandwidth that is characteristic of FH. The major disadvantages are that the implementation of FH-DS is considerably more complex than that of pure FH, and that the near-far problem is expected to be significant when frequency hits occur, because only limited separation can be provided among the DS codes. We have noted that the improved antimultipath capability may not really be needed, however, because robust signaling schemes that are tolerant to multipath interference can be used. Thus, although some benefits may arise from the use of hybrid FH-DS signaling, we have concluded that the use of pure FH in these studies represents a reasonable

choice based on both performance and practical implementation considerations. Furthermore, the choice of FH facilitates the analysis of AJ and multiple user channel performance, both of which are discussed elsewhere in this report.

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Appendix D

DATA MODULATION METHOD: NONCOHERENT FSK

In this appendix we address considerations relating to the choice of the data modulation method, and we present the basis for our recommendation of the use of noncoherent frequency shift keying (FSK). Binary and M-ary signaling are considered.

The first choice to be made regarding the modulation method is between coherent and noncoherent schemes for the modulation and detection of the FH signal. A FH system is said to be coherent if the dehopper at the receiver maintains phase coherence with the received signal. It is noncoherent if such coherence cannot be maintained. Coherent FH systems are much more difficult to implement than noncoherent systems. In the case of the HF skywave channel, the dispersive nature of the medium would, in fact, preclude the use of coherent FH systems because of frequency selective fading; i.e., it would not be possible to maintain the relative phases among the different frequencies. It may be possible to implement a coherent system for HF groundwave signals; however, it would be inadvisable to use a signaling scheme that could not also be used for skywaves.

Another choice to be made is that of the hopping rate. We have noted earlier that it is generally advisable to hop as fast as possible (with respect to channel and equipment limitations) to obtain multipath rejection and to avoid repeater jammers. A hopping rate of 2400 hops/s is assumed in many of the examples in Section 7. Fast FH systems (i.e., more than one hop per symbol) permit the use of diversity (i.e., multiple transmissions of the same symbols) and/or the use of lower rate codes. In either case, improved interference rejection is achieved. In the case of skywave signals, hopping rates will probably be limited to at most several hundred hops per second because of the dependence of propagation path, and therefore propagation delay, on frequency. However, with groundwave signals it is possible to use considerably higher hopping rates that are limited primarily by equipment and bandwidth availability constraints, rather than the properties of the medium.

The most practical data modulation scheme for noncoherent FH systems that hop one or more times per symbol is FSK. The simplest form of FSK is binary FSK (BFSK), under which each encoded bit is transmitted as one of two frequency tones; one frequency corresponds to a *mark* (a logical 1) and the other to a *space* (a logical 0). Figure D1 illustrates a noncoherent receiver for FH-BFSK. The receiver bases its decisions on the relative energy contained in the two frequency tone positions. No phase information is required, and each bit is independently detected; FSK is therefore highly robust in terms of fading and interference. In narrowband FSK systems, the two frequency tone positions remain constant. In a FH-BFSK system, the narrowband containing the two frequencies is hopped according to the FH pattern.

To avoid spectral overlap, the two tones are usually separated by the *orthogonal spacing,* which is $1/T_b$ Hz for rectangular pulses (and also raised cosine pulses), where T_b is the binary symbol duration as shown in Fig. D2. The bandwidth required to contain both the mark and space frequency tones is therefore $2/T_b$. For a bit rate of 2400 b/s, this bandwidth is then 4.8 kHz. It is often advisable to use pulse shaping to reduce spectral splatter (at the expense of increasing the orthogonal tone spacing), and/or to allow frequency guard bands between channels, especially when there are large power differentials among signals in the network. For example, a channel bandwidth of $4/T_b$ might be used for binary signaling.

When two or more symbols are transmitted per hop, differentially encoded modulation schemes such as differential phase shift keying (DPSK), differential quadriphase shift keying (DQPSK), or

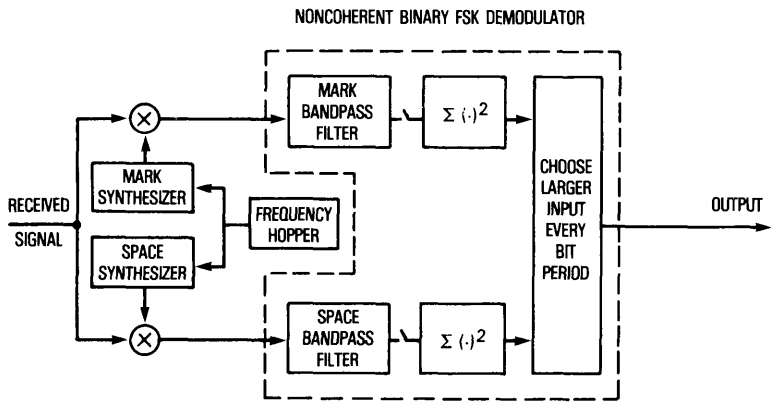

Fig. D1 — Frequency-hopping receiver

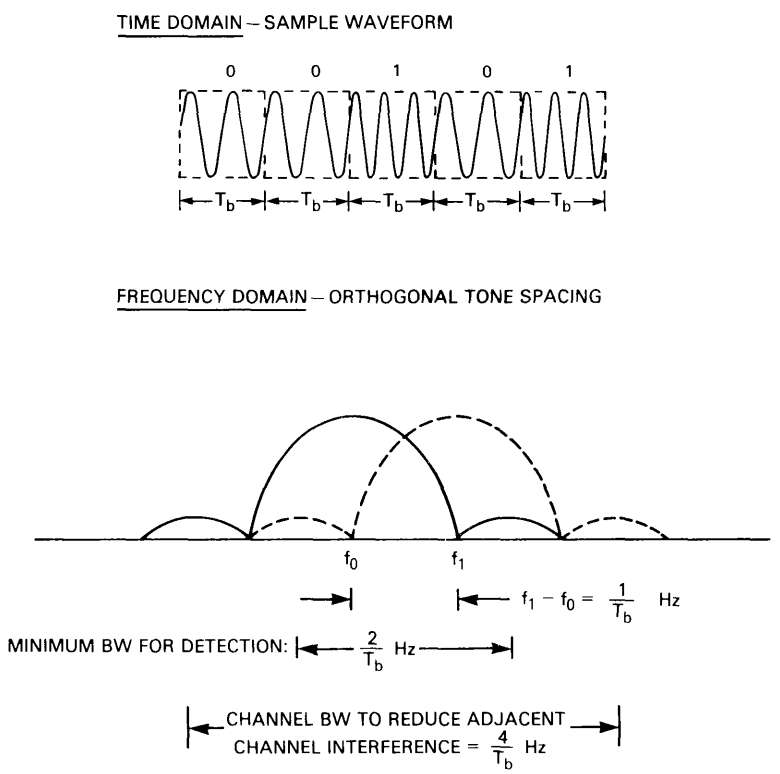

Fig. D2 — Binary frequency shift keying (BFSK)

minimum shift keying (MSK) are possible alternatives to FSK [D1]. Under these schemes a relative phase is established in each hop via the transmission of an extra phase reference. The overhead resulting from this additional symbol will be significant only if the number of symbols transmitted per hop is small. In schemes such as these, a stable phase reference must be maintained throughout the hop duration, but it may be difficult to do so in a fading environment. The advantages of these schemes are a lower BER for given values of E_b/N_0 as well as reduced bandwidth requirements. The implementation of these schemes is more complex, however, than that of noncoherent FSK.

Despite the possible advantages of some of these modulation schemes at lower hopping rates, it is felt that noncoherent FSK is the most practical scheme for the HF ITF Network for the following reasons:

a. Feasibility of implementation;
b. Robustness with respect to channel properties (such as fading and interference);
c. Compatibility with hopping schemes in which either several symbols are transmitted per hop, one symbol is transmitted per hop, or several hops are transmitted per symbol.

D1 GENERALIZATION TO M-ARY FSK

In our discussion of FSK, we have thus far considered only the binary case, i.e., the transmission of one of two possible frequency tones to represent a binary symbol. BFSK can be generalized to M-ary FSK (MFSK), in which one of $M = 2^K$ frequency tones of duration T_s is transmitted to represent one K-bit symbol. We again assume the orthogonal tone spacing of $1/T_s$. It is clear that, for a fixed signaling rate of one MFSK symbol per time unit, the bandwidth of MFSK signals increases exponentially with K, whereas the bit rate increases only linearly with K. Since bandwidth is a scarce resource in the HF ITF environment, BFSK and quaternary FSK (QFSK, i.e., $M = 4$, $K = 2$) would be most appropriate from a bandwidth per bit point of view. (Under BFSK a one-bit symbol of duration T_s requires a bandwidth of $2/T_s$, while under QFSK a two-bit symbol of the same duration requires a bandwidth of $4/T_s$; the bandwidth required per transmitted bit is therefore the same. Under 8-ary FSK, however, a three-bit symbol of the same duration requires a bandwidth of $8/T_s$, or $8/3T_s$ Hz per bit as compared to $2/T_s$ Hz per bit in the other two cases).

Quaternary Frequency Shift Keying (QFSK) is well suited to our maximum data rate requirement of 2400 b/s. For example, if a hopping rate of 2400 hops/s is used, then 2400 quaternary symbols per second would be transmitted, resulting in a transmitted symbol rate equivalent to 4800 binary symbols per second. If rate 1/2 coding (e.g., rate 1/2 convolutional code of constraint length 7) is used, the resulting data rate is the desired 2400 b/s. If the orthogonal frequency spacing of $1/T_s$ is used, the required frequency bin bandwidth is then $4/T_s$, or 9600 Hz. In practice, the use of pulse shaping and/or guard bands will require a frequency bin bandwidth of approximately 20 kHz (about twice the ideal bandwidth) to adequately suppress interference.

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Appendix E

AJ LINK PERFORMANCE OF UNCODED FH-MFSK SIGNALS

In this appendix we derive the AJ link performance that can be achieved through the use of uncoded FH-MFSK signals. The performance measure we use is the tolerable J/S ratio:

$$J/S|_{\max} = PG - E_b/N_0,$$

where all quantities are expressed in dB. In this equation the processing gain PG = W/R is simply the ratio of total spread hopping bandwidth W to the data rate. In Section E1 we discuss the evaluation of the bit error rate (BER) as a function of equivalent E_b/N_0 for uncoded systems. We consider two jamming threats (broadband noise jamming and worst case partial band (WCPB) noise jamming), and two types of channels (nonfading and Rayleigh fading). Knowledge of the value of E_b/N_0 required for a specified signaling scheme at a specified BER permits the evaluation of the tolerable J/S ratio, which we discuss in Section E2.

E1 BER PERFORMANCE

In this appendix and in Appendix F we present curves that illustrate the dependence of BER on modulation coding, and diversity, as a function of E_b/N_0. With these curves, which are based on [E1], we can compare various schemes on the basis of their performance as a function of this fundamental quantity. The performance curves illustrate the values of E_b/N_0 that are required to ensure a specified BER.*

E1.1 BER Performance on Nonfading Channels; No Coding

For the case of uncoded signaling and nonfading channels we consider three situations:

a. broadband noise jamming,
b. WCPB noise jamming: no diversity,
c. WCPB noise jamming: diversity.

We demonstrate the tremendous performance degradation caused by WCPB noise jamming to a system without diversity, as compared with broadband jamming. We also demonstrate that, by using diversity, the communicator can nullify virtually all of the advantage originally gained by the WCPB jammer.

Broadband Jamming of FH-MFSK Signals

We are assuming the use of noncoherent MFSK signaling; each M-ary symbol corresponds to K binary symbols, where $M = 2^K$. Continuous wave (CW) jamming at the correct frequency would place the most energy into a noncoherent MFSK detector, since CW signals are tones and therefore look like MFSK signals. However, multitone jamming, which is also a realistic and possibly somewhat greater threat, is not expected to have a significant effect on achievable communication ranges as compared with noise jamming, as is discussed in Section 9. For ease of analysis we shall assume that the jammer uses Gaussian noise waveforms.

*Note that few exact results for BER are obtainable, and most results are given in terms of upper bounds. The bounds we present are believed to be accurate to within 1 dB for BER less than 10^{-3} (i.e., the actual value of E_b/N_0 required to achieve a specified BER is within 1 dB of the value shown on the curve).

As shown in Fig. E1, we assume that the jamming signal can be modeled as an additive white Gaussian noise (AWGN) process with single sided spectral density

$$N_0 = J/W,$$

which is constant over the entire bandwidth W. In this case of broadband jamming, the equivalent E_b/N_0 defined earlier is thus identical to the actual E_b/N_0 throughout the entire bandwidth W.

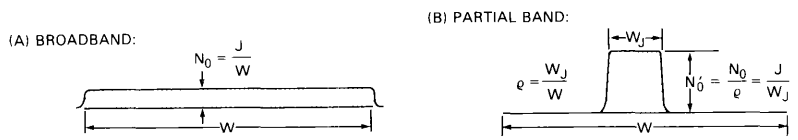

Fig. E1 — Noise jammer

The probability of M-ary symbol error for uncoded MFSK in AWGN is [E1, note 2]:

$$P_s = \sum_{n=1}^{M-1} \binom{M-1}{n} (-1)^{n+1} \frac{1}{n+1} e^{-\frac{n}{n+1}\left(K \frac{E_b}{N_0}\right)}.$$

The resulting bit error probability is

$$P_b = \frac{M}{2(M-1)} P_s.$$

The probability of bit error as a function of E_b/N_0 is illustrated in Fig. E2 for various values of M. It is clear from this figure that as K increases the probability of bit error decreases for a given E_b/N_0. The reason for the improved performance is that the probability of symbol error is directly related to the energy per M-ary symbol to noise density ratio E_s/N_0 (where $E_s = KE_b$ and $M = 2^K$) rather than directly to E_b/N_0. The decrease in symbol error probability with increasing K is considerably faster than 1/K, whereas the number of bit errors per symbol error is equal to M/2(M-1); the net result is therefore a decrease in bit error probability as K increases. The improved performance with respect to E_b/N_0 for higher values of K is obtained at the expense of bandwidth. The data rate increases linearly with K, but the required bandwidth at a fixed signaling rate of MFSK symbols per second increases exponentially with K.*

Partial Band Jamming: No Diversity

A partial band noise jammer spreads noise of total power J evenly over some bandwidth W_j, as shown in Fig. E1. We define the fraction of the spread spectrum that is jammed as

$$\rho = W_j/W.$$

The noise density in the jammed fraction of the band is therefore,

$$N_0' = J/W_j = N_0/\rho$$

The jammer would like to choose ρ to maximize the probability of bit error; this is called worst case partial band (WCPB) jamming. Any other choice of ρ by the jammer would result in better communication performance. Therefore, the results we present are pessimistic from the communicator's viewpoint since the jammer will usually not be able to choose exactly the optimum (i.e., worst case) value of ρ, which, as will be demonstrated, depends on M-ary alphabet size, received signal energy per bit, coding/diversity scheme, and jamming noise density at the receiver. (The results are also pessimistic because we are using upper bounds on bit error probabilities rather than exact expressions; the

*Note that binary and 4-ary signaling are equally efficient in terms of bandwidth required at a given binary symbol rate; the inefficiency results for $M \geqslant 8$.

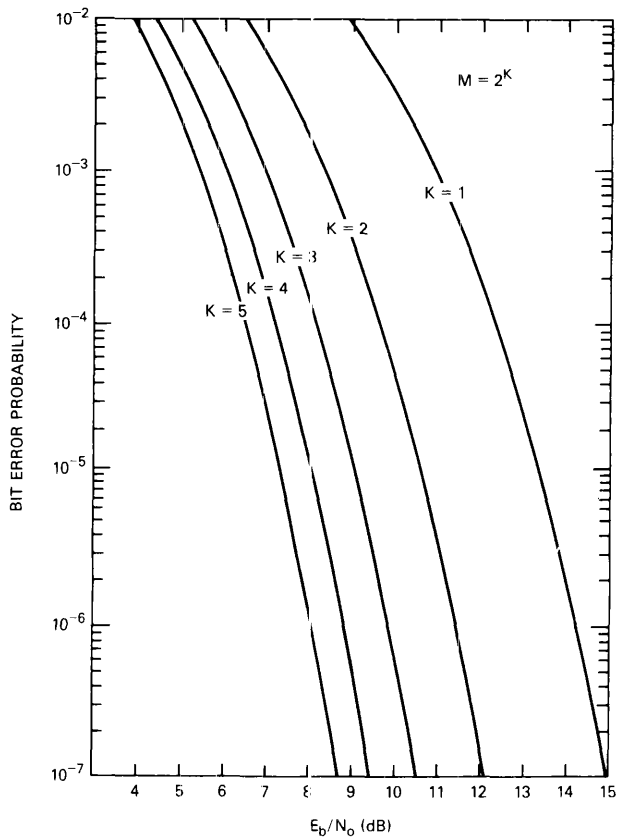

Fig. E2 — Performance of noncoherent FH-MFSK in broadband noise jamming; no coding or diversity (from [E1])

bounds are believed to be accurate to within 1 dB for BER less than 10^{-3}.) Normally, the jammed band will be frequency hopped, although not necessarily at a fast rate, to prevent the communicator from simply choosing a quiet part of the subband in which to operate.

The bit error probability of FH-MFSK signaling with no diversity or coding is shown in Fig. E3 for both broadband and WCPB noise jamming. Note that the broadband curve is simply the bit error probability in Gaussian white noise, and it is applicable to narrowband signaling as well. Although, for broadband jamming, both narrowband and wideband FH signaling schemes have the same BER performance as a function of E_b/N_0, in the FH case the jammer must spread its energy over a wide bandwidth. If the total jamming energy is fixed, then N_0 is inversely proportional to the number of frequency bins over which the signal is hopped. The tolerable J/S ratio, defined earlier, which takes into consideration the signaling scheme and the processing gain, is thus a more significant quantity in terms of ultimate system performance, as will be demonstrated in Section E2.

In Fig. E3 we see that WCPB noise jamming degrades performance tremendously. For example, we observe a 32 dB degradation for binary FSK at a bit error probability of 10^{-5} (i.e., the value of E_b/N_0 required is 32 dB greater than that for broadband jamming).

We note that the curves of Fig. E3 are exact. They are not based on either the union bound or union-Chernoff bound that have been used to evaluate the AJ performance of systems with coding and/or diversity. We also note that the use of the union bound to evaluate AJ performance of a system without coding or diversity in WCPB Gaussian noise jamming, leads to the misleading conclusion that performance degrades as alphabet size M increases. It has been demonstrated by Crepeau and

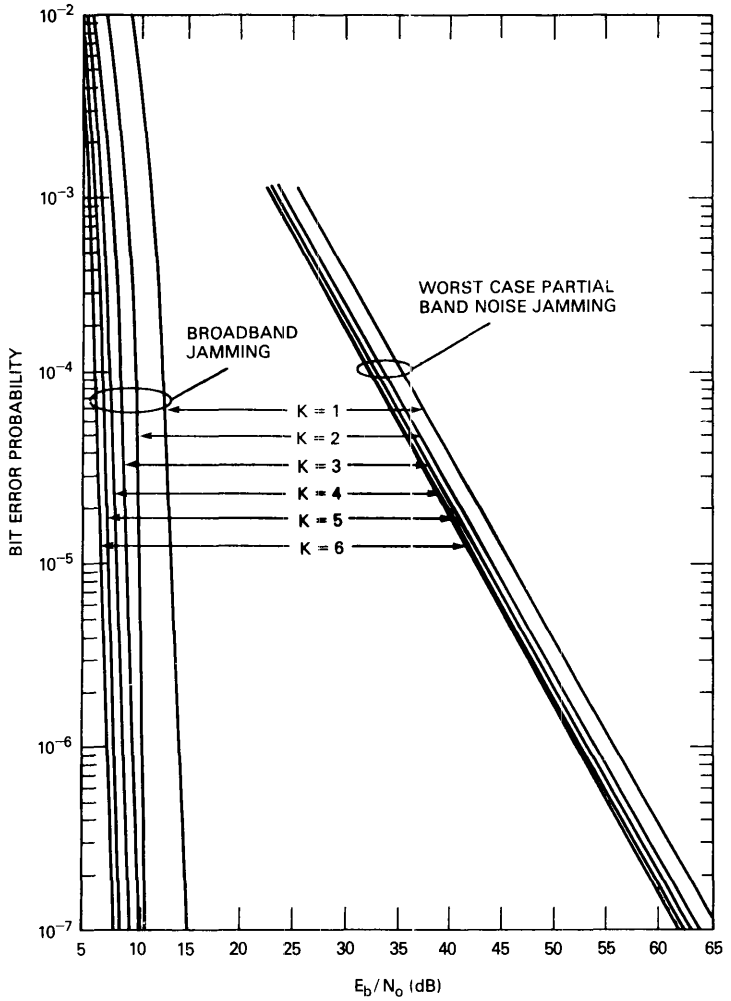

Fig. E3 — Performance of noncoherent FH-MFSK in broadband and WCPB noise jamming; no coding or diversity (from [E1])

McGregor [E2] that performance does in fact improve as M increases, as it did in the broadband jamming case. Such misleading behavior of the bounds occurs only when bit error probability is a linearly decreasing function of E_b/N_0. (The bounds are not incorrect; they simply become increasingly loose as M increases.) In cases with coding and/or diversity, however, bit error probability is an exponentially decreasing function of E_b/N_0, and the bounds demonstrate the improvement with increasing M, which is, in fact, correctly indicative of system performance.

Partial Band Jamming: Diversity

The deterioration of performance that occurs when WCPB noise jamming is used, as compared with the broadband noise jamming case, can be almost totally eliminated by the use of diversity. Diversity, in its simplest form, is the division of each MFSK symbol into m MFSK chips which are independently hopped.* The receiver noncoherently combines the outputs of the m chip energy detectors for each of the M baseband tones. The received energy per chip can be expressed as

$$E_c = E_s/m = KE_b/m.$$

*Diversity is also inherently provided by forward error correction coding, as will be discussed in Appendix F. Note that E_b is the energy received per information bit (i.e., the received power divided by the data rate R), rather than the energy received per M-ary chip. E_s is the total received energy of all m chips corresponding to a single M-ary symbol.

The use of diversity, for a fixed alphabet size and data rate, clearly results in an increased hopping rate (and therefore bandwidth per frequency bin) by a factor of m. Alternatively, if the hopping rate (and therefore bandwidth per frequency bin) is kept constant, the data rate can be lowered to provide diversity.

The bit error probability curves that we shall present have been generated under the assumption that a soft decision receiver is used, and that Jammer State Information *(J.S.I.)* is known. Under this latter assumption:

We have an ideal receiver that knows which of its received chips are jammed and which are not.

It is expected that the performance of practical receiver structures can approach that of such an ideal receiver, although some small processing losses can be expected. In Appendix H we indicate how the results presented in this section can be extended to the case of less intelligent receiver structures in which hard decision decoding is used and *J.S.I.* either is or is not available.

If *J.S.I.* is available, then all m chips of a symbol must be jammed in order for a symbol error to occur (although the case in which all chips are jammed does not necessarily result in an error). Although it might at first appear desirable to use as much diversity (i.e., as large a value of m) as possible, there is a point of diminishing return, beyond which the noncoherent combining losses outweigh the benefits of having a greater number of chips upon which to make a symbol decision.

A minimax approach is used to determine the WCPB jamming fraction ρ^* and the optimum diversity parameter m^* (see [E1 note #2]). First we consider this problem from the standpoint of the jammer who wants to choose the most disruptive value of ρ. ρ^* is a function of m (the number of chips per symbol) as well as of K and E_b/N_0:

$$\rho^* = \begin{cases} \dfrac{3m}{K(E_b/N_0)}, & \dfrac{E_b}{N_0} > \dfrac{3m}{K} \\ 1, & \dfrac{E_b}{N_0} \leqslant \dfrac{3m}{K} \end{cases}.$$

An upper bound on bit error probability, based on the Chernoff bound, is:

$$P_b \leqslant 2^{K-2} \left[\frac{4m}{eK(E_b/N_0)} \right]^m.$$

We now consider this problem from the viewpoint of the communicator. This bound can be minimized by choosing m so that

$$m^* = KE_b/4N_0.$$

(Although the analysis assumed real values of m, the use of the closest integer to m^* will not affect results appreciably.)

Use of optimal diversity results in a corresponding WCPB noise jamming fraction of

$$\rho^* = 3/4.$$

We have thus determined the minimax (ρ^*, m^*) pair.

Note that if the communicator uses a different value of m, his performance may be either better or worse than that corresponding to m^*, depending on the value of ρ used by the jammer. If the value of m is different from m^* and is known by the jammer, the latter can vary ρ accordingly to result in

more severely degraded performance. The use of m^* thus represents a relatively *safe* choice for the communicator. Similarly, if the jammer uses a value of ρ different from ρ^* and the communicator knows this value, then he can adjust m to improve his performance. The effects of using different values of m and ρ in partial band noise jamming applications are discussed in [E3].

The performance of FH-MFSK with optimum diversity, for $K = 1$ and 5, is compared with the no diversity case in Fig. E4. We can see that when optimum diversity is used, the performance is only about 3 dB worse than in the broadband jammer case of Fig. E2. By using diversity the communicator has therefore been able to nullify virtually all of the partial band advantage originally gained by the jammer. This has been accomplished by forcing the jammer to spread his power over most (3/4) of the band.

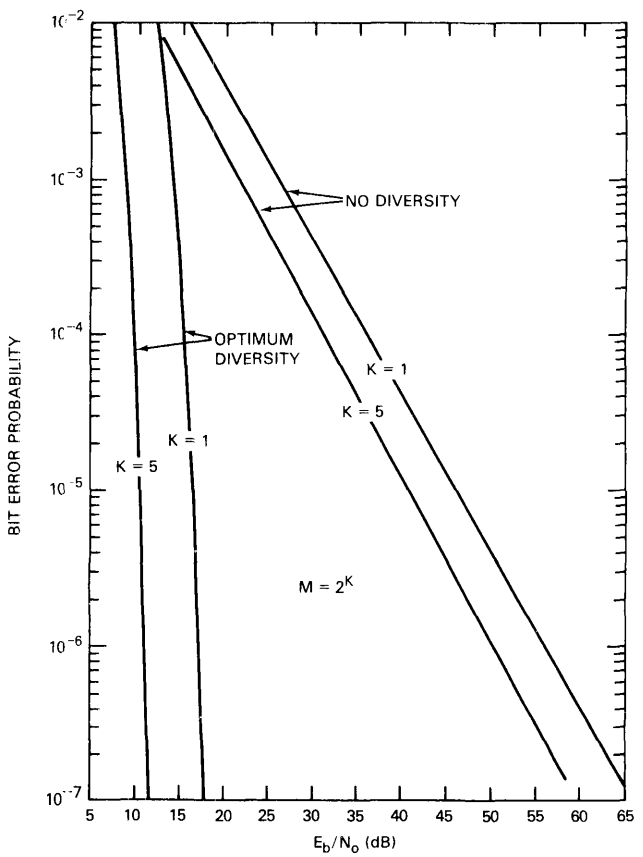

Fig. E4 — Performance of noncoherent FH-MFSK in WCPB noise jamming; no coding; no diversity and optimum diversity; soft decision receiver; J.S.I. known (from [E1])

In practice, the value of diversity that can be implemented is limited by the hopping rate of the system. If sufficiently high values of diversity cannot be obtained at a given data rate, then diversity can be increased by lowering the data rate (thereby increasing the number of chips per symbol for a given hop rate). It is unlikely that the HF ITF Network will always use the optimum value of m, because data rates and hop rates are generally determined by criteria other than that of obtaining optimum diversity. For example, it is anticipated that the hopping rate would be fixed and that sources at several data rates will be using this same hopping rate. The optimum diversity value m^* is, as we have noted, a function of received bit energy and jammer average noise spectral density. Often, there will be tradeoffs between maintaining specified data rates and BER performance.

We now define a new diversity parameter:

$$L = R_h/R = m/r,$$

where R_h is the hopping rate, and R is the information data rate in bits per second. The composite code rate r, as defined in Section B4.1, is the number of information bits per M-ary symbol (for the *no diversity*, $m = 1$ case). In words, L is the diversity per uncoded data bit, i.e., the number of M-ary symbols or chips transmitted over the channel for each source bit. L can in fact assume fractional values; e.g., if 8-ary FSK is used with no coding (corresponding to $r = 3$, as discussed in Section B4.1), and $m = 1$ (no diversity) then $L = 1/3$. To clarify further the relationship between L and other system parameters we note that

$$E_b = LE_c,$$

where E_c is the received chip energy, as defined earlier. As a reminder, m is the number of times that each M-ary symbol* is transmitted over the channel. The use of L rather than m permits the more direct performance comparison of signaling schemes that use different code rates.

Figures E5 through E7 show bounds on the bit error probability for uncoded systems as a function of E_b/N_0 under WCPB noise jamming for several values of diversity, for binary, quaternary (i.e., M = 4), and 8-ary FSK respectively. Also, the performance of a system with no diversity in the presense of broadband jamming is shown for reference. These figures clearly show the benefits of diversity, as well as the fact that the optimum value of diversity depends on K and E_b/N_0. Excessively large values of L result in inefficient use of transmitted energy; however, if the received power levels are low it may, in fact, be necessary to use large values of L to achieve sufficiently high values of E_b.

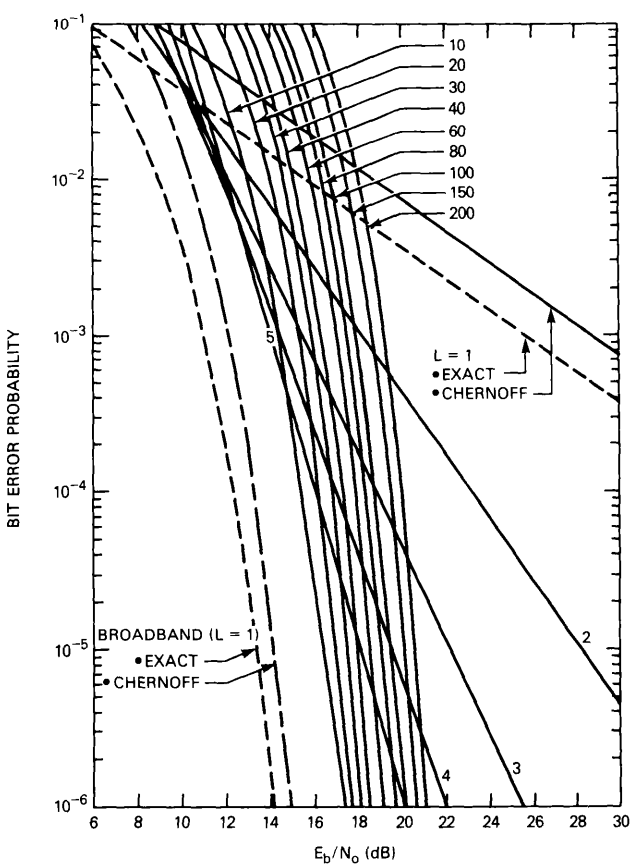

Fig. E5 — Performance of noncoherent FH Binary FSK in WCPB noise jamming; nonfading channel; no coding; several values of diversity; soft decision receiver; J.S.I. known (from [E1])

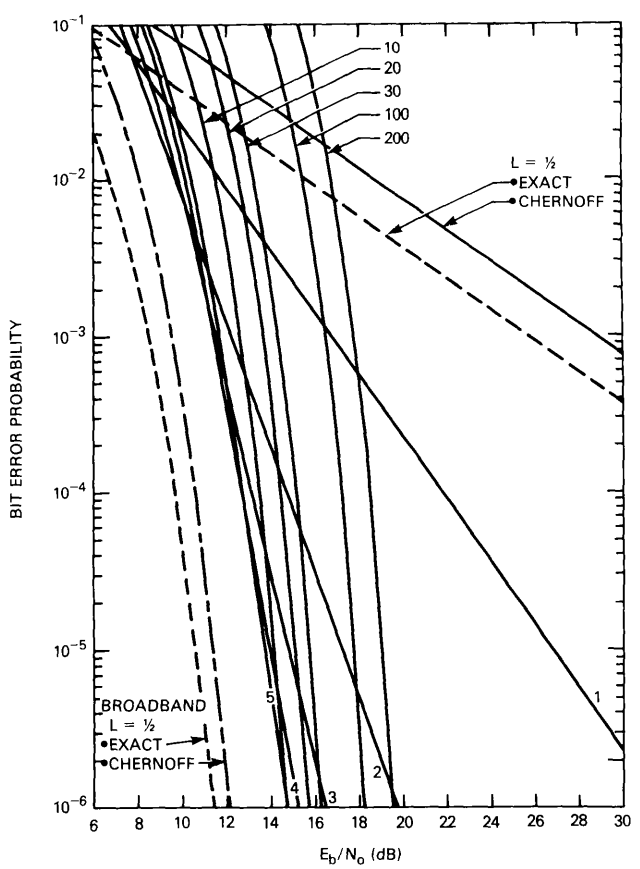

Fig. E6 — Performance of noncoherent FH 4-ary FSK in WCPB noise jamming; nonfading channel; no coding; several values of diversity; soft decision receiver; J.S.I. known (from [E1])

*This interpretation of m is appropriate in both encoded and coded systems; in the latter case the M-ary symbol is the output of the encoder.

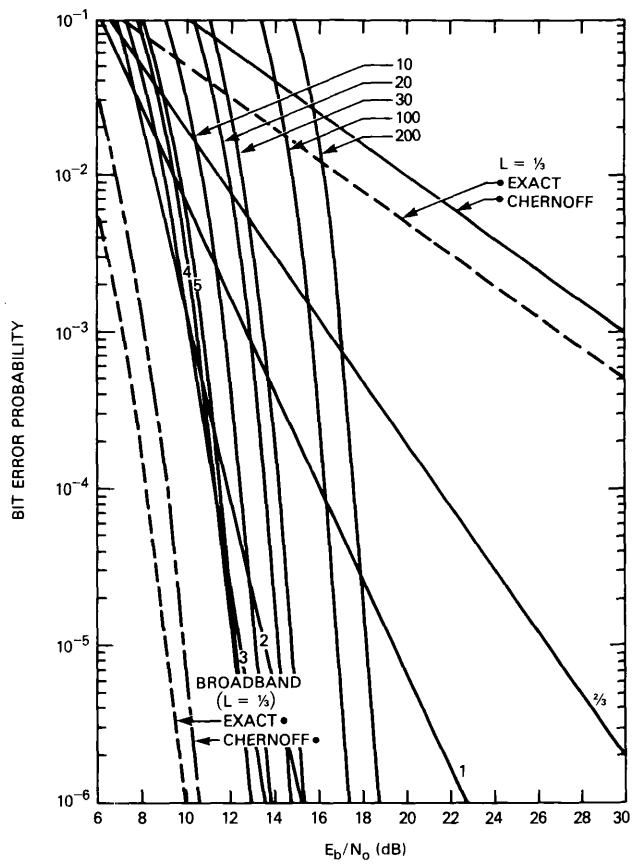

Fig. E7 — Performance of noncoherent FH 8-ary FSK in WCPB noise jamming; nonfading channel; no coding; several values of diversity; soft decision receiver; J.S.I. known (from [E1])

The results presented for various values of diversity are based on the Chernoff bound. In the no diversity ($m = 1$) no coding case, it is also possible to obtain *exact* results, which are also shown in Figures E5 through E7. These results are actually based on the union bound which is *almost* exact [E1 note #2]. For the cases with diversity, no such comparisons are possible, but it is clear that the actual bit error probabilities are somewhat lower than the values shown in the curves, and it is believed that the bounds are tight to within 1 dB for BER less than 10^{-3}.

E1.2 BER Performance on Rayleigh Fading Channels; No Coding

Although most of the links in the HF ITF Network are expected to be groundwave links that experience little dispersion, it is expected that some skywave links may be used either within the network or for communication to points external to it. An understanding of the AJ performance over fading channels is therefore essential. A fading model may also be appropriate at low hopping rates when operating at frequencies that support skywave propagation. Although a Rician (i.e., specular-plus-Rayleigh) fading model might be more appropriate than a Rayleigh model in some cases (where the groundwave signal is the specular component and the skywave signal is the Rayleigh component), a Rayleigh model is useful because it provides a worst-case bound for Rician channel performance. The results summarized here are based on [E4 and E1].

An important feature of the signal model is that each chip is assumed to experience independent fading and jamming noise, thereby resulting in random symbol errors. Considerable interleaving may be necessary, in practice, to ensure that the chips of any symbol experience independent fading. The degree of interleaving that can be achieved is limited in a system that is characterized by a short packet size. Further study is needed to assess the impact of short packet length on the validity of the random error assumption for HF channel fading characteristics.

It is interesting to note that for the uniform Rayleigh fading channel (i.e., one in which propagation conditions are uniform across the entire channel bandwidth) and for any value of diversity, the WCPB noise jammer is the broadband jammer ($\rho^* = 1$) for any of the receiver types we are considering (i.e., hard and soft decision receivers, with and without *J.S.I.* known). Since the jammer model is uniform broadband Gaussian noise, the results presented are also valid for an unjammed AWGN channel that is characterized by Rayleigh fading. For this uniform Rayleigh fading case, the BER soft decision bound is shown in Fig. E8 for several values of alphabet size M (2 to 32), and diversity, m. Also included in this figure are the results without diversity. In the absence of diversity, an E_b/N_0 value of 50 dB is required to maintain a BER of 10^{-5} in the binary case. With optimum diversity, the required value of E_b/N_0 is reduced to 18.6 dB, an improvement of approximately 31 dB.

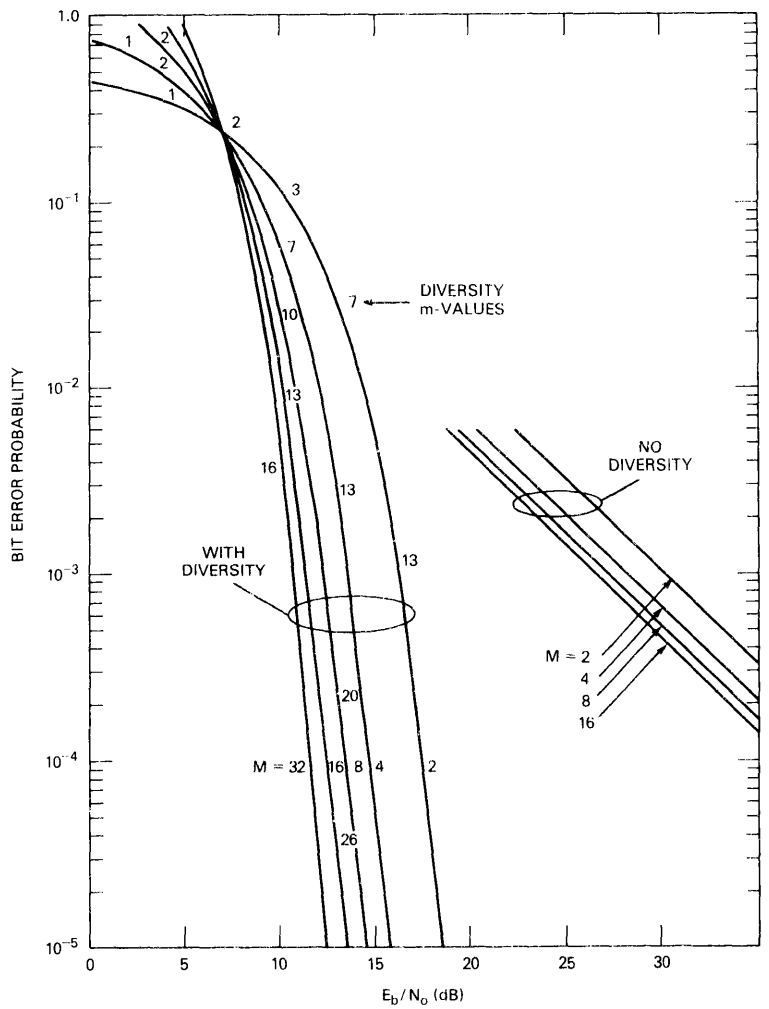

Fig. E8 — Performance of noncoherent FH-MFSK in broadband jamming on a Rayleigh facing channel; soft decision bound (From [E1])

Performance for several values of diversity (L) is shown in Figs. E9 through E11 for uncoded binary, quaternary, and 8-ary FSK, respectively. These curves are for the *ideal* soft decision receiver with J.S.I., and may be compared directly with the curves for the nonfading case given in Figs. E5 through E7. Note that the horizontal axis of the plots is now $E\{E_b\}/N_0$ rather than E_b/N_0. (i.e., the numerator is now the expected value of the received bit energy). A procedure for the evaluation of BER for hard decision receivers with or without *J.S.I.* is presented in Appendix H.

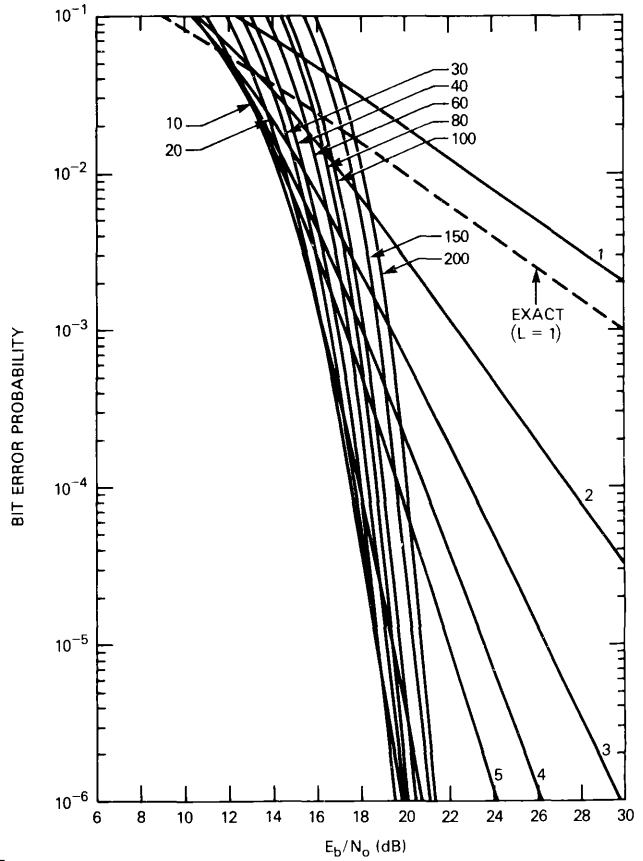

Fig. E9 — Performance of noncoherent FH Binary FSK in broadband noise jamming; Rayleigh fading channel; soft decision receiver; J.S.I. known (from [E1])

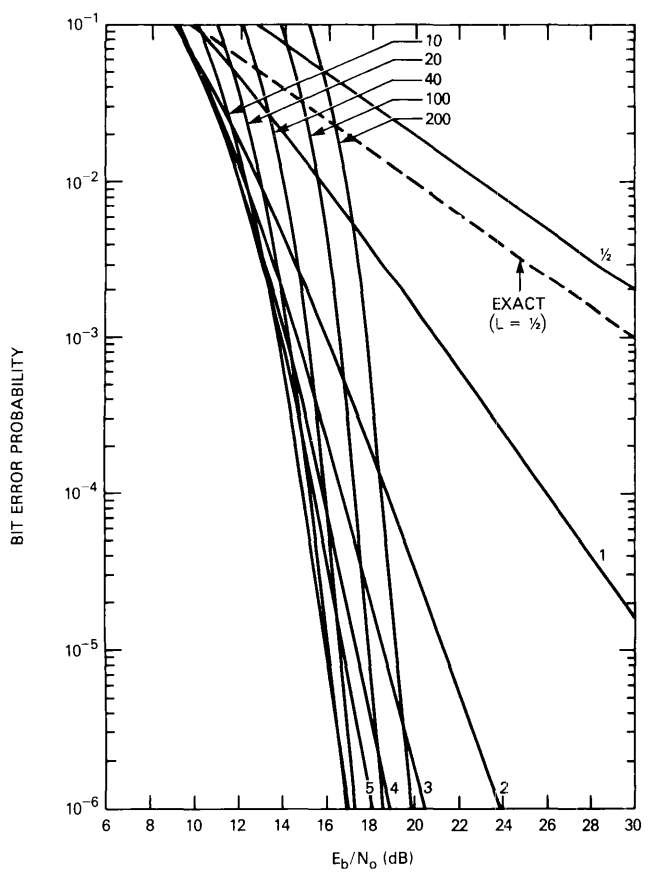

Fig. E10 — Performance of noncoherent FH 4-ary FSK in broadband noise jamming; Rayleigh fading channel; soft decision receiver; J.S.I. known (from [E1])

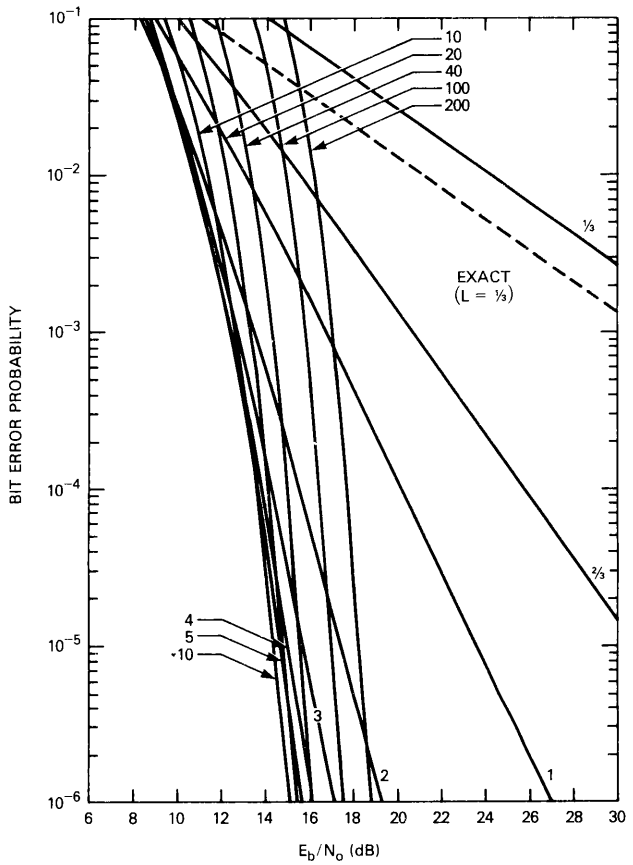

Fig. E11 — Performance of noncoherent FH 8-ary FSK in broadband noise jamming; Rayleigh fading channel; soft decision receiver; J.S.I. known (from [E1])

It is clear from these figures that higher diversity values as well as higher values of $E\{E_b\}/N_0$ are needed in the fading case as compared with the nonfading case. The reason for this is that the received signal is a sample of a Gaussian random process with a Rayleigh fading envelope. The received chip is now characterized by its expected energy $E\{E_c\}$, which is Rayleigh distributed. Thus, a fraction of the received chips will have amplitude considerably lower than the mean value as a result of fading.

E2 TOLERABLE J/S RATIOS FOR UNCODED FH-MFSK SIGNALS

The AJ performance of a signaling scheme has been expressed as the tolerable J/S ratio:

$$J/S|_{\max} = PG - E_b/N_0,$$

where each of these quantities is expressed in dB. In the previous subsection we demonstrated the BER performance as a function of equivalent E_b/N_0 for noncoded FH-MFSK signaling schemes. The processing gain (PG) is simply the ratio of the spread hopping bandwidth to the information data rate. Therefore, we are now able to compare several signaling schemes that use different values of M-ary alphabet size and several values of diversity, on the basis of the tolerable J/S protection values that can be achieved for realistic HF ITF link parameters. The performance of FH-MFSK signaling on both nonfading and Rayleigh fading channels is discussed.

This subsection considers only uncoded systems, although the use of M-ary signaling and diversity can both be thought of as special cases of coding. Tolerable J/S ratios for convolutionally coded signaling schemes are presented in Appendix F.

E2.1 Nonfading Channels

Table E1 summarizes, as a function of alphabet size, the required E_b/N_0 values for BER = 10^{-5} for both the broadband as well as WCPB Gaussian noise jammer, for both nonfading and Rayleigh fading channels. First, we consider the case of nonfading channels. The WCPB noise jammer impacts severely on the system with no diversity by causing an increase in E_b/N_0 ranging from 32.3 dB to 34.6 dB (M = 2 to 16) over that required to combat broadband jamming. The corresponding tolerable J/S protection ratios, $J/S|_{max}$, which unlike E_b/N_0 values depend on data rate and processing gain, are also shown in this table for an information data rate of 2400 b/s and a 5 MHz spread bandwidth. The processing gain of such a system is 33.2 dB. Note that the bandwidth required per frequency slot is MR_h Hz, while the channel symbol rate is KR_h binary symbols per second. Clearly, this uncoded FH-MFSK design is, indeed, *unprotected* against partial band jamming as indicated by the negative J/S values. Lowering the data rate to 75 b/s (a factor of 32) would increase the processing gain, and therefore the tolerable J/S ratio, by 15 dB. Fortunately, as discussed in Section E1.1, most of the performance degradation caused by WCPB noise jamming can be recovered via the use of coding and/or diversity, which is actually a special case of coding.

Table E1 — E_b/N_0 and $J/S|_{max}$ Values for Uncoded FH-MFSK; BER = 10^{-5}

M	Broadband Jamming	WCPB Jamming		Broadband Jamming and Rayleigh Fading	
		No Diversity	Optimum Diversity	No Diversity	Optimum Diversity
	E_b/N_0 Values				
2	13.4	45.7	16.4	50.0	18.6
4	10.7	43.7	13.6	47.9	15.8
8	9.1	42.9	12.1	47.0	14.4
16	8.1	42.6	11.0	46.4	13.5
	$J/S|_{max}$ (2400 b/s data rate; W = 5 MHz)				
2	19.8	−12.5	16.8	−16.8	14.6
4	22.5	−10.5	19.6	−14.7	17.4
8	24.1	−9.7	21.1	−13.8	18.8
16	25.1	−9.4	22.2	−13.2	19.7

The results in Table E1 apply to the case of high-quality data reception, where the BER is equal to 10^{-5}. If only medium-quality data reception is required, corresponding to BER = 10^{-3}, the results are somewhat different as illustrated in Table E2. Although the WCPB jammer still impacts severely on the system, the increase in E_b/N_0 over the broadband jamming case is only 14.8 to 16.5 dB (as compared with 32.3 to 34.6 dB in the high-quality data reception case). This fact results in positive J/S protection ratios as shown in Table E2. Therefore, an uncoded FH-MFSK design, which is *unprotected* while operating at BER = 10^{-5}, can indeed achieve some minimal degree of protection if the BER constraint is relaxed to 10^{-3}.

BER performance curves for several values of diversity have been presented in Section E1.1. From these curves the optimum values of diversity (m is approximately 10 for the 10^{-5} case and somewhat smaller for the 10^{-3} case) can be obtained and the appropriate E_b/N_0 and tolerable J/S ratios can be determined. The results are shown in Tables E1 and E2. Between 29.3 and 31.6 dB (M = 2 to 16) improvement in J/S has been realized for the 10^{-5} case by using optimum diversity. When the BER is relaxed to 10^{-3}, only 11.7 to 13.5 dB (M = 2 to 16) improvement is realized by using optimum diversity. Most important, the FH-MFSK system with diversity now becomes jam-resistant. Since in

practice it is usually not feasible to use optimum diversity, the tolerable J/S ratios would probably be somewhat lower than those shown in these two tables.

E2.2 Rayleigh Fading Channels

E_b/N_0 and $J/S|_{max}$ values for Rayleigh fading channels are also shown in Tables E1 and E2. For the Rayleigh faded jammed channel the use of optimum diversity improves performance by 31.4 to 32.9 dB ($M = 2$ to 16) as shown in Table E1 for the case of BER $= 10^{-5}$, which is roughly comparable to that achieved on the nonfaded WCPB noise jammed channel. However, the optimum value of the diversity parameter m is generally quite large ($10 \leqslant m \leqslant 30$) and is considerably greater than that required to combat the WCPB noise jammer in the nonfading channel.

Similar results are shown in Table E2 for the case of medium-quality data reception, where BER $= 10^{-3}$. For this case the diversity improvement is 13.7 to 14.9 dB as compared with 31.4 to 32.9 dB for the BER $= 10^{-5}$ case. This improvement is roughly equivalent to the diversity gain achieved on the nonfading WCPB noise jammed channel. For the fading channel the optimum value of diversity also decreases when the BER constraint is relaxed to 10^{-3} ($4 \leqslant m \leqslant 10$).

Table E2 — E_b/N_0 and $J/S|_{max}$ Values for Uncoded FH-MFSK; BER $= 10^{-3}$

M	Broadband Jamming	WCPB Jamming		Broadband Jamming and Rayleigh Fading	
		No Diversity	Optimum Diversity	No Diversity	Optimum Diversity
	E_b/N_0 Values				
2	10.9	25.7	14.0	30.0	16.3
4	8.3	23.7	11.4	27.9	13.7
8	7.0	22.9	10.1	27.0	12.4
16	6.1	22.6	9.1	26.4	11.5
	$J/S|_{max}$ (2400 b/s data rate; $W = 5$ MHz)				
2	22.3	7.5	19.2	3.2	16.9
4	24.9	9.5	21.8	5.3	19.5
8	26.2	10.3	23.1	6.2	20.8
16	27.1	10.6	24.1	6.8	21.7

In summary, performance in uniform broadband jammed Rayleigh fading channels (broadband jamming is in fact WCPB jamming for these channels) is approximately 2 dB poorer in terms of $J/S|_{max}$ values than for the WCPB jammed nonfading channel, when optimum diversity is used in both cases. As stated previously, optimum diversity values may often not be realizable. Hence, it is expected that the J/S protection ratios shown in Tables E1 and E2 will actually be somewhat less for an HF ITF design. The actual choice of diversity will be a subject of further study and will depend also on the use of coding.

E3 SUMMARY OF AJ PERFORMANCE OF UNCODED FH-MFSK SIGNALS

From the discussion of uncoded FH-MFSK signaling, we conclude that:

- An HF ITF FH-MFSK design without diversity (and/or coding, which is discussed in Appendix F) can never provide any jamming protection when the BER constraint is 10^{-5}. Relaxing the BER requirement to 10^{-3} will allow for some minimal jamming resistance.
- Diversity and/or coding are essential to a jam-resistant HF ITF FH-MFSK system design.

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E2. P.J. Crepeau and D.N. McGregor, "Anomalous Union Bound Behavior for MFSK Signaling on Inverse Linear Channels," *IEEE Trans. on Commun.* **COM-31**, 848-849, Jun. 1983.

E3. B.K. Levitt, presentation in session on "Spread Spectrum Communication in Jamming," Proc. Workshop on "Research Trends in Military Communications," organized by the Commun. Sci. Institute of the Univ. of Southern California, May 1983.

E4. D. Avidor, "Anti-Jam Analysis of Frequency Hopping M-ary Frequency Shift Keying Communication Systems in High Frequency Rayleigh Fading Channels," Ph.D. dissertation, University of California, Los Angeles, Cal., Jun. 1981.

Appendix F

AJ LINK PERFORMANCE OF CONVOLUTIONALLY CODED FH-MFSK SIGNALS

In this appendix we show that an AJ coding gain of 34 to 35 dB can be obtained by using convolutional coding, and that considerably less diversity is needed in a coded system than in an uncoded system. Again, we consider the WCPB noise jammed nonfading and Rayleigh fading channels. In the case of the Rayleigh fading channel the broadband noise jammer is actually the WCPB noise jammer. In Section F1 we consider the BER performance of such a signaling scheme using rate 1/2 constraint length 7 convolutional codes. Then in Section F2 we discuss the resulting tolerable J/S performance ratios. All of the results presented in this appendix are based on the use of an ideal soft decision receiver with Jammer State Information *(J.S.I.)*. The extension of these results to hard decision receivers is discussed in Appendix H.

F1 BER PERFORMANCE

Again, we discuss separately the cases of nonfading and Rayleigh fading channels. The curves presented in this Section are taken from Omura [F1, note 8].

F1.1 BER Performance on Nonfading Channels; Coding and Diversity

Figure F1 illustrates the BER performance of an optimal binary ($M = 2$) rate 1/2 convolutional code of constraint length 7 [F2] on the WCPB noise jammed nonfading channel. These curves are parameterized by the diversity parameter $L = m/r$, which was introduced in Section E1.1. For the case of $m = 1$ (*no diversity*) a diversity of $L = 1/r$ is inherently provided by the coding process. The addition of diversity, i.e., the transmission of each coded M-ary symbol m times, clearly increases the diversity by a factor of m. Note that r is the composite code rate, as defined in Section B4.1.

For a BER of 10^{-5}, the optimal diversity for binary signaling is approximately $m = 2(L = 4)$ for the nonfading channel. However, to minimize the diversity used (and therefore reduce the required bandwidth or increase the achievable data rate), we could choose $m = 1(L = 2)$, i.e., no diversity, for the nonfading channel, and obtain performance within 0.3 dB of optimum. If we relax the BER constraint to 10^{-3}, we do not require any diversity for the nonfading channel. Figure F2 illustrates the BER performance of quaternary (4-ary) FSK signaling with a rate $r = 1$ convolutional code that has been optimized for quaternary sequences [F3]. Similarly, Fig. F3 illustrates the BER performance of 8-ary FSK signaling with a rate $r = 1$ convolutional code that has been optimized for 8-ary sequences.

F1.2 BER Performance on Rayleigh Fading Channels; Coding and Diversity

In Fig. F4 we illustrate the BER performance of a binary rate 1/2 optimum convolutional code of constraint length 7 for the Rayleigh fading WCPB (broadband) noise jammed channel. For a BER of 10^{-5} the optimal diversity is approximately $m = 3(L = 6)$. However, performance within 0.3 dB of optimum is obtained by using $m = 2(L = 4)$. For a BER constraint of 10^{-3} we require $m = 2(L = 4)$. Figure F5 illustrates the BER performance of quaternary FSK signaling with an optimal quaternary convolutional code of rate $r = 1$. Similarly, Fig. F6 illustrates the BER performance of 8-ary FSK signaling with a rate $r = 1$ convolutional code that has been optimized for 8-ary sequences.

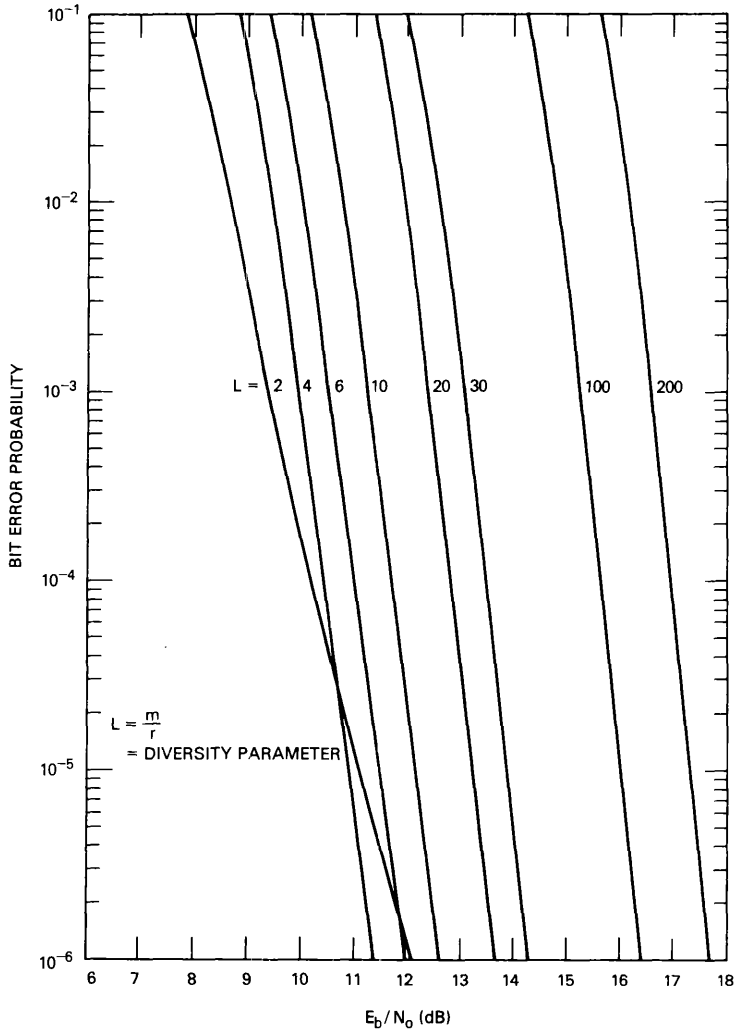

Fig. F1 — Performance of noncoherent FH binary FSK; rate 1/2 constraint length 7 optimum convolutional coding; WCPB noise jamming; nonfading channel; soft decision receiver; J.S.I. known (from [F1])

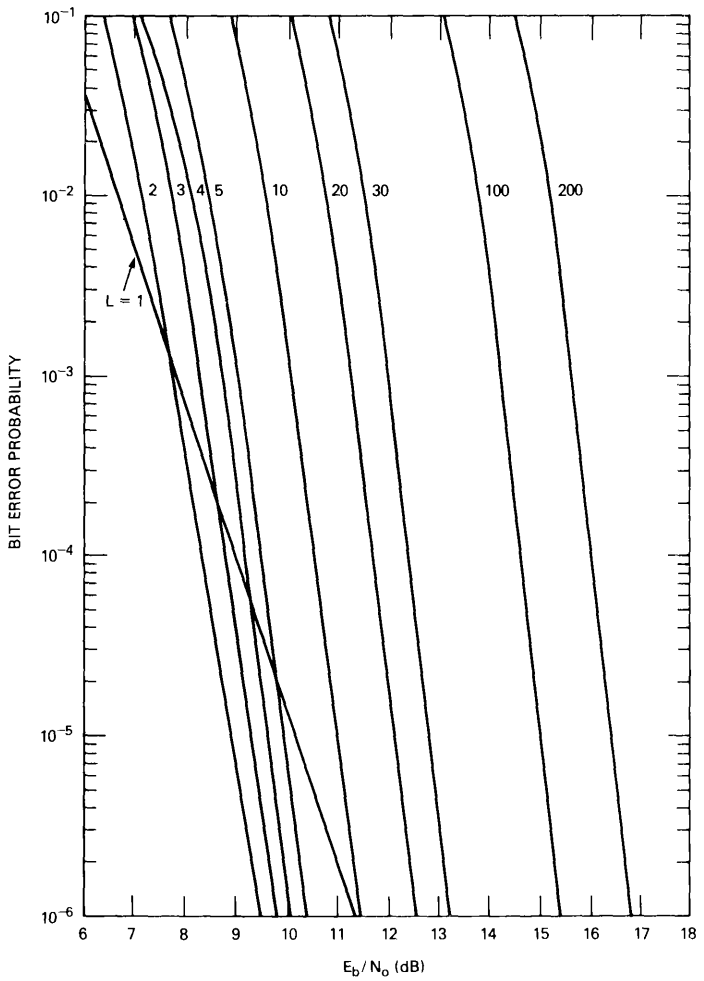

Fig. F2 — Performance of noncoherent FH 4-ary FSK; rate $r = 1$ constraint length 7 optimum 4-ary convolutional coding; WCPB noise jamming; nonfading channel; soft decision receiver; J.S.I. known (from [F1])

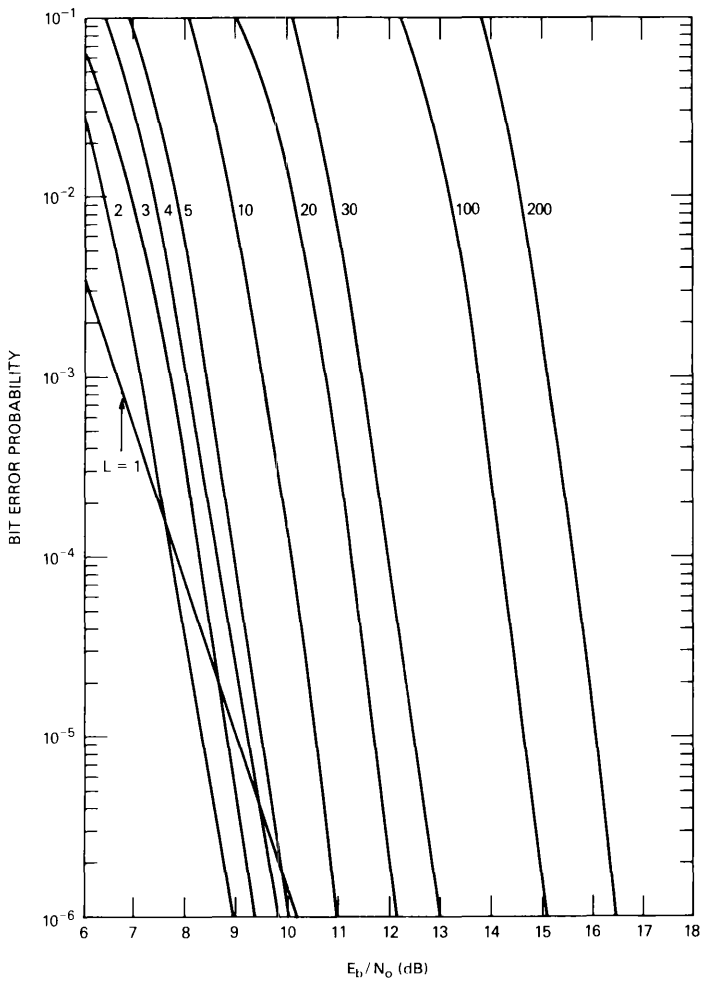

Fig. F3 — Performance of noncoherent FH 8-ary FSK; rate $r = 1$ constraint length 7 optimum 8-ary convolutional coding; WCPB noise jamming; nonfading channel; soft decision receiver; J.S.I. known (from [F1])

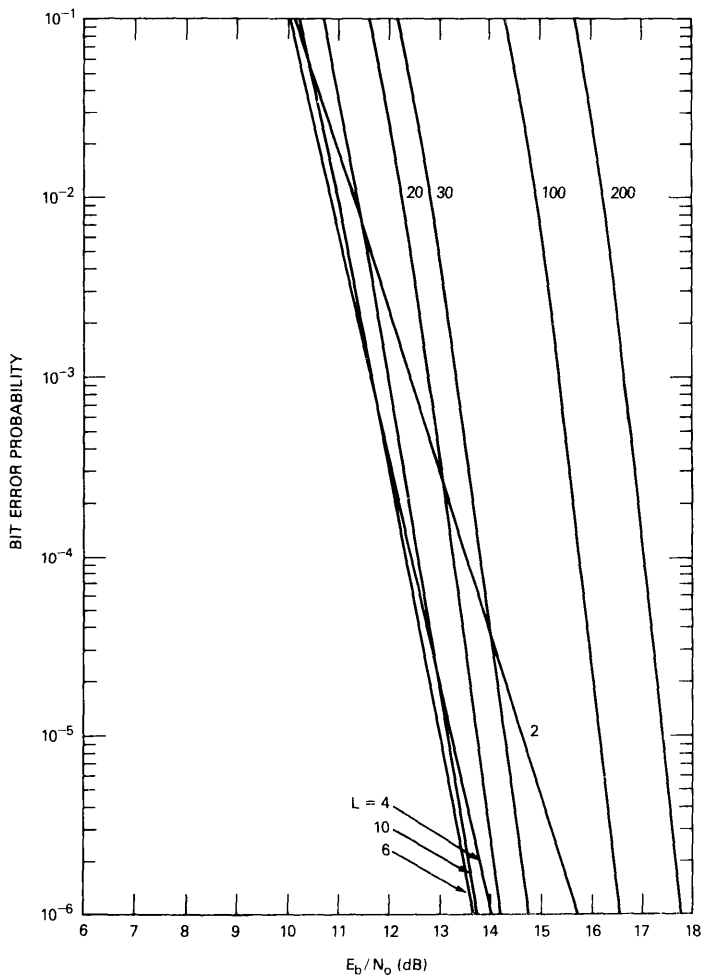

Fig. F4 — Performance of noncoherent FH binary FSK; rate 1/2 constraint length 7 optimum convolutional coding; Rayleigh fading channel; broadband noise jamming; soft decision receiver; J.S.I. known (from [F1])

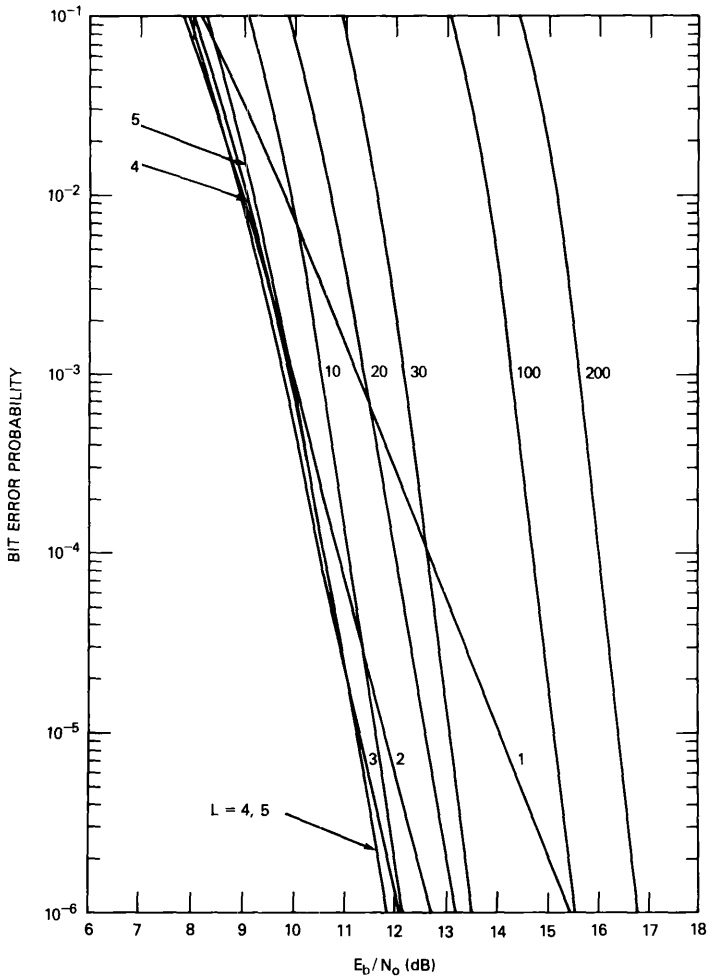

Fig. F5 — Performance of noncoherent FH 4-ary FSK; rate $r = 1$ constraint length 7 optimum 4-ary convolutional coding; Rayleigh fading channel; broadband noise jamming; soft decision receiver; J.S.I. known (from [F1])

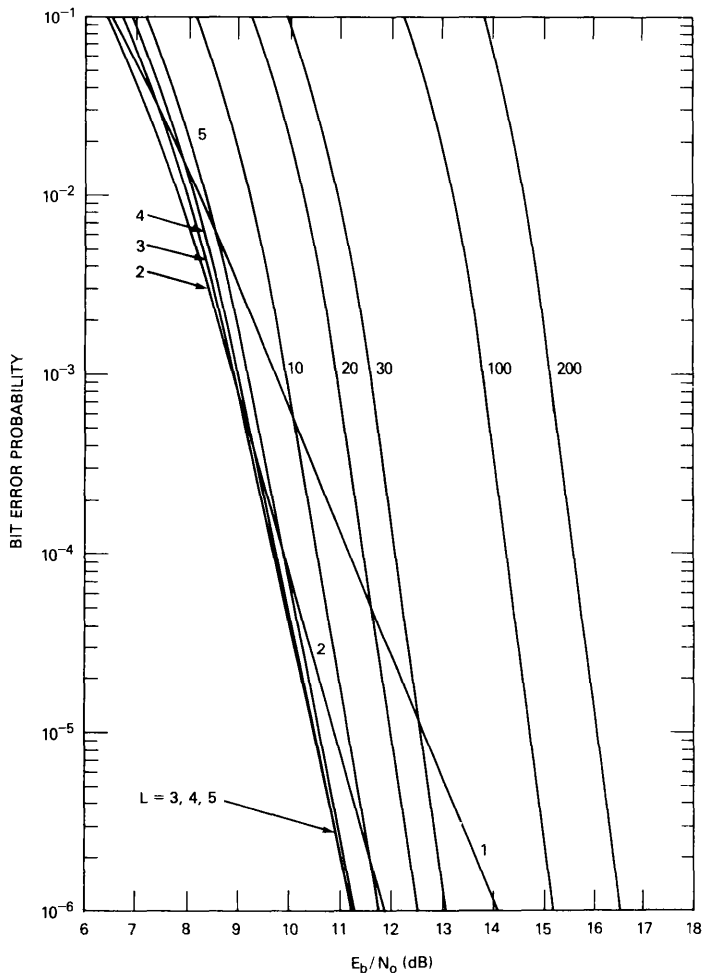

Fig. F6 — Performance of noncoherent FH 8-ary FSK; rate $r = 1$ constraint length 7 optimum 8-ary convolutional coding; Rayleigh fading channel; broadband noise jamming; soft decision receiver; J.S.I. known (from [F1])

F2 TOLERABLE J/S RATIOS FOR CONVOLUTIONALLY CODED FH-MFSK SIGNALS

When the BER is 10^{-5}, very little diversity ($2 \leqslant m \leqslant 4$) is required to obtain optimum performance when rate 1/2 coding is used. This is in direct contrast to the amount of diversity required to obtain optimum performance when coding is not used ($10 \leqslant m \leqslant 30$). Increasing the error rate to 10^{-3} also implies small diversity requirements ($1 \leqslant m \leqslant 3$) for optimum performance with coding. In the 10^{-3} case where coding is not used, more diversity is required ($2 \leqslant m \leqslant 10$). All these results serve to illustrate that the improvement in performance is principally attributed to the use of coding, and that very little additional diversity is actually required beyond that inherently provided by the coding process. A summary of the required E_b/N_0 values is given in Tables F1 and F2 where the BER is specified at 10^{-5} and 10^{-3}, respectively. Also included are the tolerable J/S protection ratios for a 2400 b/s user that hops over a 5 MHz bandwidth. By comparing Table F1 with E1 and Table F2 with Table E2, we can evaluate the performance improvement obtained by the use of coding under both high and medium data reception quality criteria.

Table F1 — E_b/N_0 and $J/S|_{max}$ Values for Convolutional Coding and M-ary Signaling; BER = 10^{-5}; Soft Decision Receiver; *J.S.I.* Known.

M	r	Broadband Jamming	WCPB Noise Jamming		Broadband Jamming + Rayleigh Fading	
			No Diversity	Optimum Diversity	No Diversity	Optimum Diversity
		E_b/N_0 Values (dB)				
2	1/2	9.7	11.2	10.9	14.6	13.0
4	1	7.5	10.2	8.9	14.0	11.2
8	1	7.1	8.9	8.3	12.6	10.5
		$J/S\|_{max}$ Values (dB) (2400 b/s data rate; W = 5 MHz)				
2	1/2	23.5	22.0	22.3	18.6	20.2
4	1	25.7	23.0	24.3	19.2	22.0
8	1	26.1	24.3	24.9	20.6	22.7

Table F2 — E_b/N_0 and $J/S|_{max}$ Values for Convolutional Coding and M-ary Signaling; BER = 10^{-3}; Soft Decision Receiver; *J.S.I.* Known.

M	r	Broadband Jamming	WCPB Noise Jamming		Broadband Jamming + Rayleigh Fading	
			No Diversity	Optimum Diversity	No Diversity	Optimum Diversity
		E_b/N_0 Values (dB)				
2	1/2	8.6	9.4	9.4	12.4	11.6
4	1	6.3	7.7	7.7	11.2	9.7
8	1	5.9	6.6	6.6	9.7	8.9
		$J/S\|_{max}$ Values (dB) (2400 b/s data rate; W = 5 MHz)				
2	1/2	24.6	23.8	23.8	20.8	21.6
4	1	26.9	25.5	25.5	22.0	23.5
8	1	27.3	26.6	26.6	23.5	24.2

Based on the results presented in Tables E1, E2, F1, and F2 the following conclusions are evident.

- When convolutional coding and diversity are not used, the M-ary (vs binary) signaling gain is approximately 2 dB for $M = 4$ and 3 dB for $M = 8$ signaling alphabets.
- Rate 1/2 convolutional coding is always better (by about 2 to 5 dB) than diversity alone for both nonfading and Rayleigh fading worst case partial band jammed channels.
- When rate 1/2 convolutional coding is used, adding diversity improves performance by about 1.5 dB ($M = 2$), 3 dB ($M = 4$), and 2 dB ($M = 8$) for the Rayleigh fading channel, under a 10^{-5} BER constraint; the improvement is even less for a relaxed BER requirement of 10^{-3}.
- The convolutional coding gain alone (as compared to a signal with no coding or diversity) is approximately 34 to 35 dB for $M = 2$ to 8 alphabets for BER $= 10^{-5}$; for BER $= 10^{-3}$ the coding gain is only 16 to 18 dB.
- The convolutional coding plus optimum diversity gain is given by:

M	r	Coding/Optimum Diversity Gain (dB)			
		WCPB Noise Jamming		Rayleigh Fading with Broadband Noise Jamming	
		BER = 10^{-5}	BER = 10^{-3}	BER = 10^{-5}	BER = 10^{-3}
2	1/2	34.8	16.3	37.0	18.4
4	1	34.8	16.0	36.7	18.2
8	1	34.6	16.3	36.5	18.0

All these results and conclusions apply to the case where we have a receiver implementation that is a soft decision device with *J.S.I.*, i.e., the knowledge of which received chips have been jammed. It is expected that the performance of practical receiver structures will be almost as good as that of such an ideal receiver. For hard decision receivers where *J.S.I.* is either known or unknown, performance curves can be derived following the procedure recently developed by Omura [F1]; this procedure is outlined in Appendix H.

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Appendix G

ON THE INTERPRETATION OF PROCESSING GAIN IN FREQUENCY HOPPING AJ SYSTEMS

In this appendix we clarify some of the features of our AJ signaling performance model. In particular, we discuss in detail our use of the term *processing gain,* a term that is often abused in the discussion of FH systems. We demonstrate that our use of the definition of processing gain as

$$PG = W/R,$$

which is based on Omura's model [G1], permits a convenient and mathematically consistent formulation of the AJ signaling problem. However, we note that this definition differs from the more traditional meaning of processing gain that refers simply to the improvement in performance that results from the frequency hopping process. In the course of our discussion we demonstrate the relationships among:

a. energies at the receiver
   — hop energy, binary symbol energy, bit energy
b. signaling rates
   — hop rate, binary symbol rate, data rate
c. code rates
   — binary, M-ary, diversity parameters m and L.

As we noted in Section 7.1, care must be used in the interpretation of processing gain in FH systems. A quantitative definition of processing gain, such as that which is used here, is valid only if the signal is hopped uniformly over the bandwidth W, and the jammer is incapable of predicting the hopping pattern or of following the signal as it hops from one frequency bin to another. It is assumed that the jammer knows everything about the signal characteristics (e.g., the hopping bandwidth, the modulation/coding/diversity scheme, and the received signal levels) except the actual FH pattern. For example, such a quantitative definition is valid for the broadband and partial band jamming examples that are considered in this report.

Other definitions of processing gain are, in fact, possible. For example, the number of frequency bins over which the signal is hopped would provide a better measure of the improvement achieved as a result of the hopping process. However, the definition we have used (which is also used e.g., in [G2]), is consistent with the problem formulation based on the equivalent E_b/N_0, and it facilitates the comparison of a number of signaling schemes on the basis of their AJ performance. Our definition of processing gain reflects in some sense the performance improvement that is achieved not only through the spectrum spreading process, but also that which is obtained from coding and diversity. The quantity PG can be made as high as desired by decreasing the code rate or by increasing the diversity. The penalty is, of course, lowered data rate.

We now review the basic signal and AJ system parameters:

$$\begin{aligned} M &= 2^K = \text{alphabet size} \\ &= \text{number of tone positions in each frequency bin} \\ K &= \text{number of binary symbols per MFSK symbol} \\ &= \log_2 M \\ R_h &= \text{hopping rate (hops per second)} \end{aligned}$$

tone spacing $= R_h$ Hz

W_b = frequency bin bandwidth $= MR_h$

W = total spread spectrum signal bandwidth (number of frequency bins times their individual bandwidth)

q = number of frequency bins in spread bandwidth
$= W/W_b = W/MR_h$

R = information data rate in bits per second

S = received signal power

J = received jammer power, summed over all frequency bins over which the signal is hopping

m = diversity parameter (i.e., number of times each encoded symbol is transmitted)

r_0 = binary code rate

r = composite code rate (see Section B4.1)
$= r_0 \log_2 M$

It is important to note that R always refers to the actual information data rate in bits per second before the application of modulation, coding, and diversity.

In Fig. G1 we illustrate the structure of a frequency bin and its relationship to the overall hopping bandwidth.

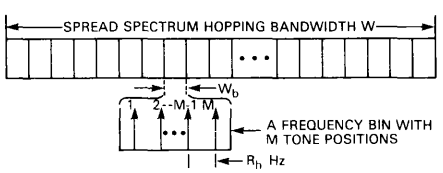


Fig. G1 — A frequency bin and its relationship to spread channel bandwidth

In Fig. G2 we illustrate the relationships among the various data rates and energies that are of interest in FH AJ systems. One M-ary symbol is transmitted per hop. Note that we reserve the units of bits/s for the information data rate exclusively. Note also that there are *symbols* and *channel symbols*. Each M-ary symbol is transmitted m times (where $m \geqslant 1$ is the diversity), thus resulting in m channel symbols. When we refer to *M-ary symbol energy* we are actually referring to the received M-ary channel symbol energy (which is equal to the received hop energy E_h).

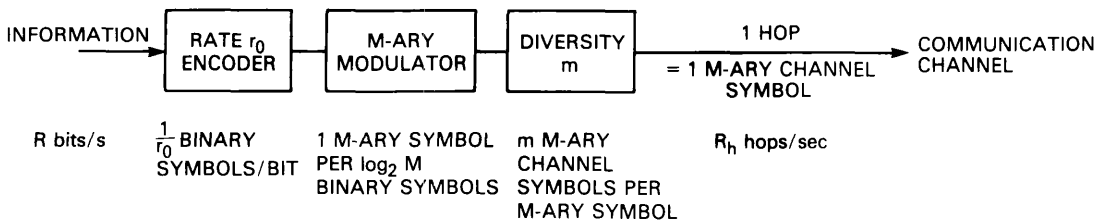
The relationships among the received signal energies are:

E_b = received energy per information bit
$= S/R$

E_h = received energy per hop*
= received energy per M-ary symbol
$= S/R_h$

E_s = received energy per binary channel symbol
$= E_h/(\log_2 M)$

*Identical to energy per chip (E_c, see Section E1.1) if one M-ary symbol is transmitted per hop.



NOTE: FOR M-ARY CODES. $r = r_0 \log_2 M$

ENERGIES AT RECEIVER

HOP ENERGY: $E_h = S/R_h$

BINARY CHANNEL SYMBOL ENERGY: $E_s = \frac{E_h}{\log_2 M} = \left(\frac{r_0}{m}\right) E_b$

BIT ENERGY: $E_b = S/R = \left(\frac{m}{r_0}\right) E_s = \left(\frac{m}{r_0}\right) \frac{E_h}{\log_2 M} = \frac{m}{r}\left(\frac{S}{R_h}\right)$

Fig. G2 — Relationships among signaling rates and energies

The average received power spectral density of the jammer noise is

$$N_0 = J/W.$$

The *equivalent bit energy to noise density ratio* is defined as,

$$E_b/N_0 = \frac{S/R}{J/W} = \frac{W/R}{J/S} = \frac{PG}{J/S}$$

The term "*processing gain*" (which has been applied to the quantity W/R by analogy with DS systems) really belongs in quotation marks because the interpretation of W/R as a true processing gain may be misleading. We do not claim, e.g., that our FH system will perform better than a narrowband system by the factor PG = W/R. If that is what one has in mind when referring to processing gain, then a better measure would indeed be q, the number of frequency bins over which the narrowband signal is hopped. However, the use of W/R in our expressions permits a consistent and more readily usable presentation of the AJ performance model as we shall demonstrate. It has been our experience that confusion easily arises over the use of our definition of processing gain. The use of a different name for this quantity might actually be the best way to reduce the possibility of misinterpretation.

Recall that we defined the *tolerable J/S ratio,* as

$$J/S|_{\max} = PG - E_b/N_0,^* \tag{G1}$$

where all quantities are in dB.* The value of E_b/N_0 used is that which is required for the desired BER for the particular modulation/coding/diversity scheme that is being used. Curves illustrating BER vs E_b/N_0 have been presented in Appendixes E and F for WCPB noise and Rayleigh fading channels. What is important to note is that these curves are presented as a function of E_b/N_0, and not of E_s/N_0 or E_h/N_0.

Processing Gain on a Per Binary Symbol Basis

We now consider the effect of adding diversity to the system or changing the code rate, and we demonstrate that our model is in fact consistent. We can write

$$E_s = \left(\frac{r_0}{m}\right) E_b.$$

*It will be obvious from the context whether we are referring to dB quantities or to actual numerical values, and so we generally omit the designation of dB in the remainder of this appendix.

Thus we have

$$E_s/N_0 = \frac{\left(\frac{r_0}{m}\right)\left(\frac{W}{R}\right)}{J/S}.$$

So we can express the tolerable J/S ratio as

$$J/S|_{\max} = \left(\frac{W}{R}\right)\left(\frac{r_0}{m}\right) - \frac{E_s}{N_0}.$$

It is clear that if the information data rate R is kept constant and m is increased from 1 (i.e., no diversity beyond that inherently provided by coding) to 2, then the hopping rate, R_h, must be doubled to provide the diversity. Doubling R_h, results in a doubling of the bandwidth of each frequency bin, W_b, and therefore a halving of the number of frequency bins because the total spread hopping bandwidth is kept fixed at W. Similar considerations apply to the use of different code rates. We could redefine processing gain as

$$PG' = (W/R)(r_0/m)$$

to reflect the impact of frequency bin bandwidth on the number of frequency bins. We would then have,

$$J/S|_{\max} = PG' - E_s/N_0. \tag{G2}$$

The two formulations for $J/S|_{\max}$ (i.e., Eqs. (G1) and (G2)) are *equivalent.* The only difference is that we have placed the factor (r_0/m) in different quantities. Omura's formulation expresses $J/S|_{\max}$ in terms of E_b/N_0, which is a more useful measure than E_s/N_0 because BER performance is normally expressed as a function of E_b/N_0 rather than E_s/N_0.

Processing Gain on a Per Hop Basis

Yet another way to look at the evaluation of $J/S|_{\max}$ is to consider the energy per hop, rather than per information bit or per binary symbol ($\log_2 M$ of which are in every hop). We have

$$E_h/N_0 = \frac{S/R_h}{J/W} = \left(\frac{S}{J}\right)\left(\frac{W}{R_h}\right).$$

We define a new *processing gain*:

$$PG'' = W/R_h$$

Therefore,

$$J/S|_{\max} = PG'' - E_h/N_0. \tag{G3}$$

This is yet a third *equivalent* expression for $J/S|_{\max}$.

Processing Gain on a Per Frequency Bin Basis

Finally, we can define processing gain in the usual manner for FH systems as follows:

$$\begin{aligned} PG''' &= q = \text{number of frequency bins} \\ &= W/W_b \\ &= W/MR_h \\ &= PG''/M. \end{aligned}$$

We obtain

$$J/S|_{\max} = PG''' + M - E_h/N_0, \tag{G4}$$

which is a fourth *equivalent* expression for $J/S|_{\max}$. Since, as we noted before, it is useful to express our results in terms of E_b/N_0, we can write,

$$\begin{aligned} J/S|_{\max} &= PG''' + M - (r_0/m)\log_2 M(E_b/N_0) \\ &= PG''' + (Mm/r_0)/(\log_2 M) - E_b/N_0. \end{aligned} \tag{G5}$$

To verify equivalence with Eq. (G1) we can combine the first two terms of the right hand side of Eq. (G5) to obtain

$$\begin{aligned} J/S|_{\max} &= (PG''')(Mm/r_0 \log_2 M) - E_b/N_0 \\ &= (W/MR_h)(Mm/r_0 \log_2 M) - E_b/N_0 \\ &= (W/R_h)(m/r_0 \log_2 M) - E_b/N_0. \end{aligned}$$

The information data rate is

$$R = (R_h\ r_0 \log_2 M)/m.$$

Therefore, we obtain

$$J/S|_{\max} = W/R - E_b/N_0,$$

which is simply Eq. (G1).

In conclusion, Omura's formulation (i.e., Eq.(G1)) is the most convenient of the five equivalent expressions for $J/S|_{\max}$, especially because BER performance results are available as a function of E_b/N_0. The quantity $PG = W/R$ is not equivalent to the conventional measure of processing gain in that it does not directly indicate the improvement that arises solely from the spectrum spreading process. PG is, in fact, related to the performance improvement that is achieved through both spectrum spreading and the coding/diversity process. Thus we perhaps should not refer to it as processing gain, but this is really just a question of terminology, rather than of substance. The use of other measures of processing gain would necessitate the introduction of *correction factors* into either the processing gain term, the E_b/N_0 term, or both, thereby unnecessarily complicating the evaluation of AJ performance.

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G2. G.C. Clark, Jr. and J.B. Cain, *Error-Correction Coding for Digital Communications*, (Plenum Press, New York, 1981).

Appendix H

THE DERIVATION OF AJ PERFORMANCE FOR HARD DECISION RECEIVERS

In Section 7 we discussed the AJ link performance that is achievable when M-ary FSK signaling is used with M-ary convolutional codes on WCPB noise jammed channels and Rayleigh fading jammed channels. Those curves were generated under the assumption of what we have termed an *ideal* receiver, i.e., one that:

a. uses soft decision decoding and
b. knows which chips are jammed.

We now demonstrate how these results may be extended to the other receiver structures of interest, namely hard decision receivers, with or without *J.S.I.* The soft decision receiver without *J.S.I.* performs poorly when jammed, and is therefore of less practical interest than the other cases, as will be discussed shortly. We refer to these other receiver structures as *nonideal* although, as noted in Section 7.1, this terminology is imprecise.

In the new technique developed by Omura [H1, note #4; H2] the coding aspects of the system are decoupled from the channel properties in the evaluation of BER performance. This decoupling is accomplished by characterizing the combined channel and modulation scheme in terms of the cutoff rate parameter R_0, which represents the maximum reliable data rate per channel use (i.e., per chip transmission) that can be achieved in a practical system. The cutoff rate is a function of the equivalent channel energy-per-chip to noise-density ratio:

$$R_0 = f(E_c/N_0) \text{ bits per channel use.}$$

R_0 depends on the type of modulation used (e.g., the value of M under MFSK), the channel characteristics (e.g., WCPB noise jamming, tone jamming, or Rayleigh fading), and the type of receiver structure; factors related to receiver structure include hard vs soft decision, the choice of decoding metric (e.g., energy vs list detection), and the availability or lack of *J.S.I.* R_0 does not depend on the particular code that is used or the degree of diversity. Therefore, E_c, rather than E_b, is fundamental to the definition of R_0. Thus, for a given modulation scheme, receiver structure, and communication channel the cutoff rate R_0 can be determined from knowledge of E_c/N_0.

For any particular code, a bound on the coded bit error probability of the form:

$$P_b < \frac{1}{2} B(R_0)$$

can be obtained. The factor of 1/2 corresponds to a tightening of bit error bounds derived by Omura [H1, note #3]. This bound is a function of only R_0. Given that R_0 is known, it is not a function of the type of receiver structure that is used (e.g., hard vs soft decision, and *J.S.I.* known or unknown), nor of the type of channel (e.g., fading vs nonfading). Since $B(R_0)$ is unique for each code, and R_0 is independent of the code, it is possible to decouple the coding from the rest of the communication system.

The four cases of interest are:

a. soft decision and known jammer state,
b. hard decision and known jammer state,
c. soft decision and unknown jammer state,
d. hard decision and unknown jammer state.

H1 NONFADING CHANNELS; WCPB NOISE JAMMING

We first consider nonfading WCPB noise jammed channels. The cutoff rate for the four cases of interest are shown in Figs. H1 through H3 for binary, quaternary, and 8-ary FSK, respectively. The concept of cutoff rate will become more apparent if particular examples are considered. First let us examine binary FSK, case (a), i.e., the ideal soft decision receiver with *J.S.I.* For $E_c/N_0 = 10$ dB we have $R_0 = 0.8$. What this means is that when a good code (e.g., an optimum convolutional code with large constraint length) is used, it should be possible to deliver data over this channel at a data bit rate equal to 80% of the transmitted bit rate (i.e., an average of 0.8 of a data bit is successfully delivered by the transmission of each binary channel symbol, or equivalently a code rate of 0.8 is sufficient for reliable communication). As long as the code rate is less than R_0, the bit error probability will decrease exponentially as code block length (for block coding) or constraint length (for convolutional coding) increases. As E_c/N_0 decreases, R_0 decreases, and the use of lower rate codes and/or additional diversity is required. For example, again for our ideal receiver, at $E_c/N_0 = 0$ dB we have $R_0 = 0.1$.

In the quaternary case, R_0 corresponds to the average number of information bits successfully delivered by each transmitted quaternary symbol for a given E_c/N_0. As E_c/N_0 becomes large, this number approaches 2. Similarly, for 8-ary signaling, R_0 approaches 3 as E_c/N_0 becomes large.

Note that the cutoff rate for case (c) (i.e., the soft decision receiver without *J.S.I.*) is zero. What this means is that the bit error probability cannot be made to decrease exponentially with block or constraint length. As a result, such a receiver would perform poorly in a jammed environment. The soft decision receiver without *J.S.I.* performs poorly because it attaches a higher weight (or importance) to signals with large amplitudes; a jammed chip may often have an energy that is greater than the sum of all other chips corresponding to the same symbol, thereby resulting in high error probability. If *J.S.I.* were available, then the jammed chips could be either excised or assigned a significantly lower weight than the unjammed chips.

We now present an example that illustrates the use of the cutoff rate curves to obtain the performance of the nonideal receivers. Consider the case of uncoded binary signaling with $L = 5$, and $E_b/N_0 = 17$ dB, as shown in Fig. H1. The resulting bit error probability for the ideal receiver is $P_b = 3 \times 10^{-5}$. We want to determine the required E_b/N_0 for the hard decision receivers, both with and without *J.S.I.*, that will result in the same error probability for the same value of diversity.

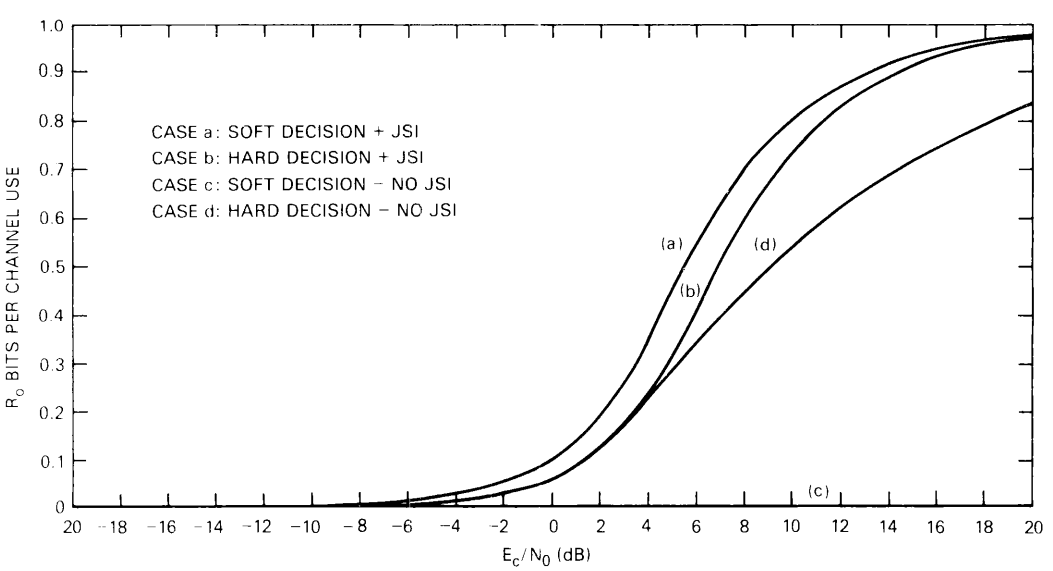

Fig. H1 — R_0 for FH-MFSK signaling ($M = 2$); nonfading channel; WCPB noise jamming (from [H1])

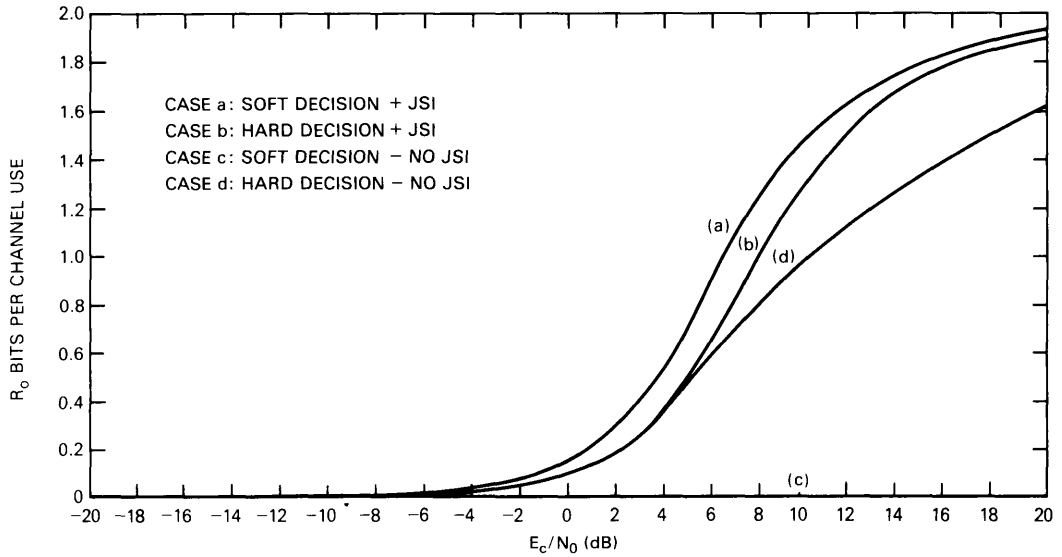

Fig. H2 — R_0 for FH-MFSK signaling ($M = 4$); nonfading channel; WCPB noise jamming (from [H1])

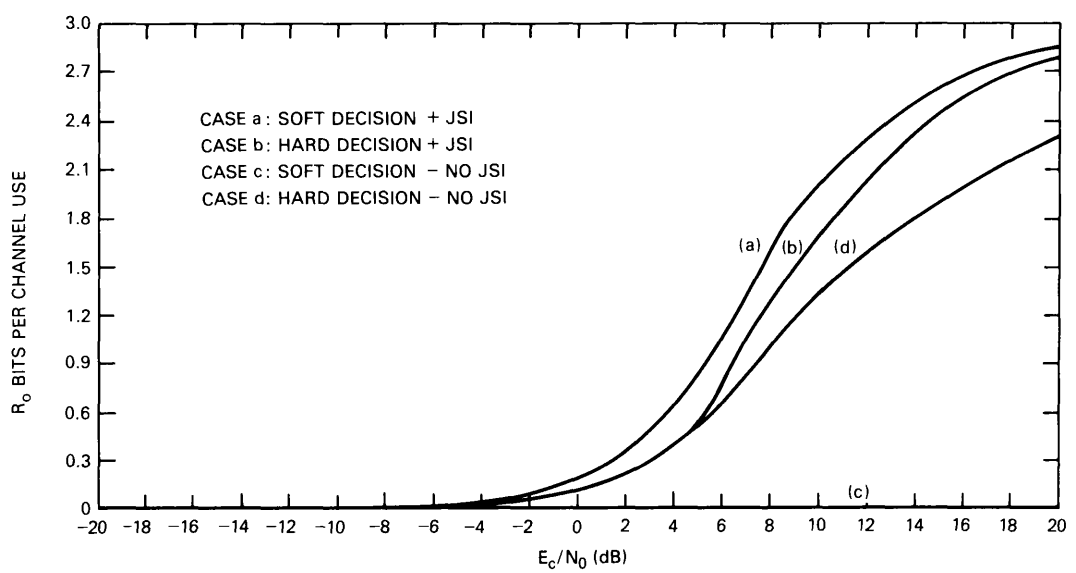

Fig. H3 — R_0 for FH-MFSK signaling ($M = 8$); nonfading channel; WCPB noise jamming (from [H1])

We begin by recognizing that the chip energy to noise density ratio is fundamental to the cutoff rate characterization, rather than the uncoded data bit (or M-ary symbol) energy to noise density ratio. Thus, knowledge of E_c/N_0 permits us to evaluate R_0. A diversity of $L = 5$ implies that $E_c = E_b/5$, and therefore $E_c/N_0 = 10$ dB (for the given value of $E_b/N_0 = 17$ dB). From Fig. H1 the corresponding cutoff rate is $R_0 = 0.8$ bits per channel use. Therefore, a cutoff rate of $R_0 = 0.8$ is required for any uncoded binary system that uses a diversity of $L = 5$ if an error probability of $P_b = 3\times10^{-5}$ is to be achieved. From Fig. H1 we see that for case (b) (i.e., hard decision with known *J.S.I.*) $E_c/N_0 = 11.3$ dB is required to achieve $R_0 = 0.8$ (as compared with 10 dB for the ideal receiver). The degradation is therefore 1.3 dB in terms of the additional signal to noise ratio that is required; this results in $E_b/N_0 =$ 18.3 dB (as compared with 17 dB for the ideal receiver). Following the same procedure, $E_c/N_0 = 18.2$ dB is required for case (d) (i.e., hard decision receiver with no *J.S.I.*), resulting in $E_b/N_0 = 25.2$ dB.

It is interesting to note that when *J.S.I.* is available, the difference in performance between hard and soft decision receivers is approximately 1.3 to 1.5 dB for all values of E_c/N_0. Another interesting observation is that for the hard decision receiver the availability of *J.S.I.* does not appreciably improve performance when E_c/N_0 is less than approximately 4 dB; however, considerable improvement is obtained for larger values of E_c/N_0. We shall demonstrate that, as a result of such behavior, the performance of hard decision systems with sufficient diversity is virtually independent of whether or not *J.S.I.* is available.

We have evaluated the BER performance for 4-ary signaling with rate 1/2 constraint length 7 convolutional coding for the three receiver structures that we are considering. In Figs. H4 through H7 we illustrate the performance for diversity values of $m = 1$ (no diversity), 2, 3 and 4, respectively. First we compare the performance of hard and soft decision receivers, both of which have *J.S.I.* As we have just noted, if *J.S.I.* is available, then the use of hard decision receivers degrades performance by about 1.3 to 1.5 dB as compared with soft decision receivers. The magnitude of this degradation is quite insensitive to the amount of diversity, because this degradation is insensitive to the value of E_c/N_0.

We next compare the performance of hard decision receivers in which *J.S.I.* either is or is not available, again for $M = 4$. For the case of no diversity, the lack of *J.S.I.* results in considerable performance degradation (6 dB at BER $= 10^{-5}$). However, as diversity increases the improvement resulting from knowledge of *J.S.I.* rapidly decreases; e.g., for $m = 4$ the improvement is only 0.1 dB at BER $= 10^{-5}$). The reason that *J.S.I.* is less significant for higher diversity values is apparent from Fig. H2. There is little difference between the R_0 curves for the two hard decision cases for $E_c < 4.5$ dB. The use of higher diversity values results in lower chip energies and therefore operation in this region of the curves.

In Fig. H8 we illustrate the BER performance for the case of quaternary signaling and a hard decision receiver with *J.S.I.* for diversity $m = 1$, 2, 3 and 4. The optimum diversity value is $m = 2$ for BER less than 2×10^{-3}; (for the soft decision case with *J.S.I.* $m = 2$ was optimal for BER $< 10^{-3}$). These curves are approximately of the same shape as those for the ideal receiver shown in Fig. F2, but translated 1.3 to 1.5 dB to the right. All of the AJ performance results (in terms of both E_b/N_0 and $J/S|_{max}$) that were obtained for the ideal receiver can thus be adapted to the hard decision receiver with *J.S.I.* simply by degrading performance values by 1.3 to 1.5 dB. Similar results hold for other M-ary alphabet sizes as well.

In Fig. H9 we illustrate the BER performance for the case of quaternary signaling and a hard decision receiver without *J.S.I.* for diversity $m = 1$, 2, 3 and 4. These curves are quite different from those of Fig. F2. For example, at BER $= 10^{-5}$ the optimum diversity is $m = 4$ (although not shown performance does in fact degrade as m is increased beyond 4); however, for $m = 3$ we come within 0.1 dB of this value, and for $m = 2$ we are within 0.8 dB.

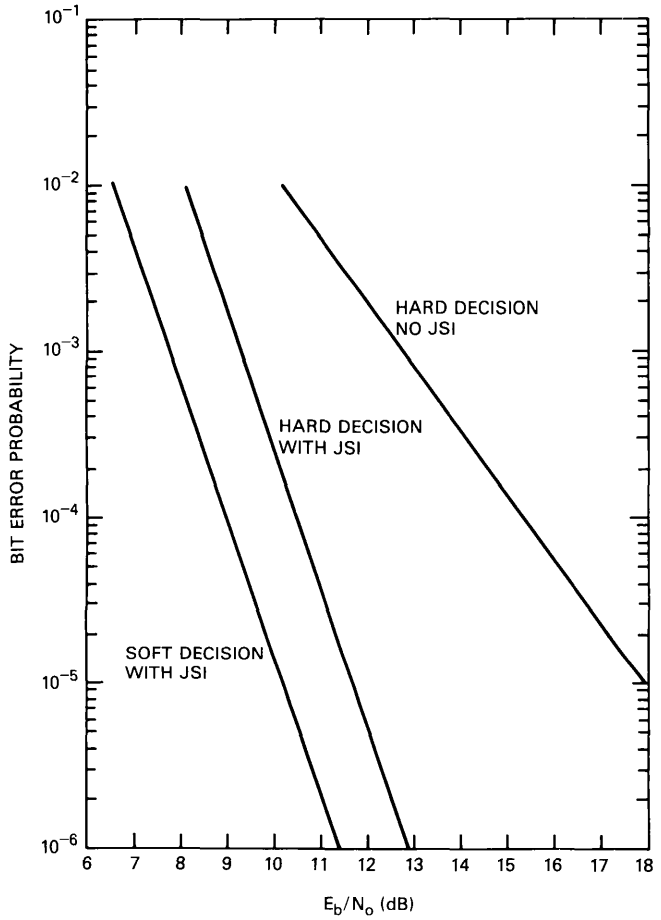

Fig. H4 — Performance of noncoherent FH 4-ary FSK; rate $r = 1$ constraint length 7 optimum 4-ary convolutional coding; WCPB noise jamming; nonfading channel; no diversity ($m = 1$)

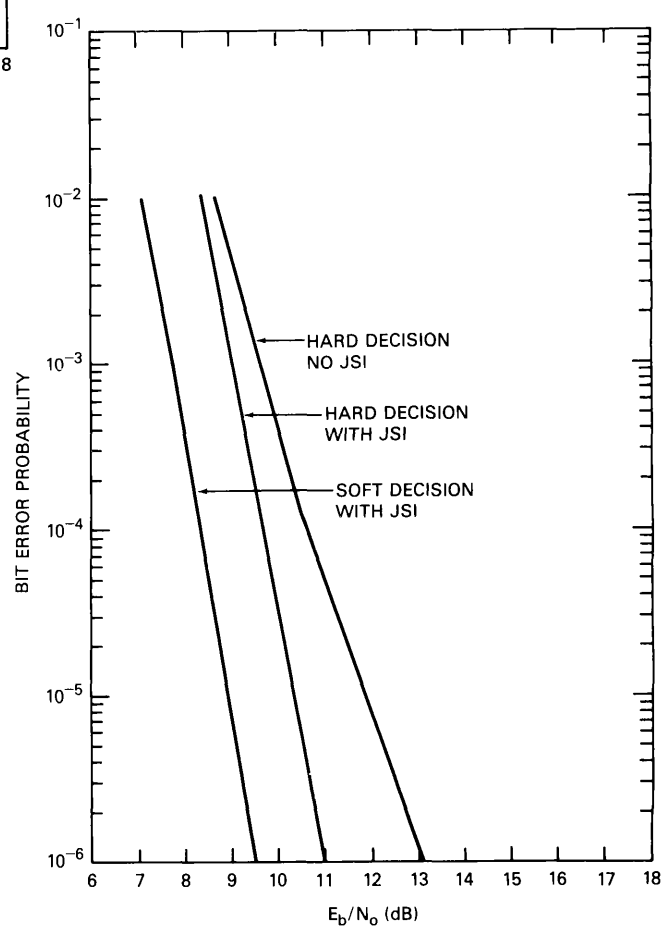

Fig. H5 — Performance of noncoherent FH 4-ary FSK; rate $r = 1$ constraint length 7 optimum 4-ary convolutional coding; WCPB noise jamming; nonfading channel; diversity $m = 2$

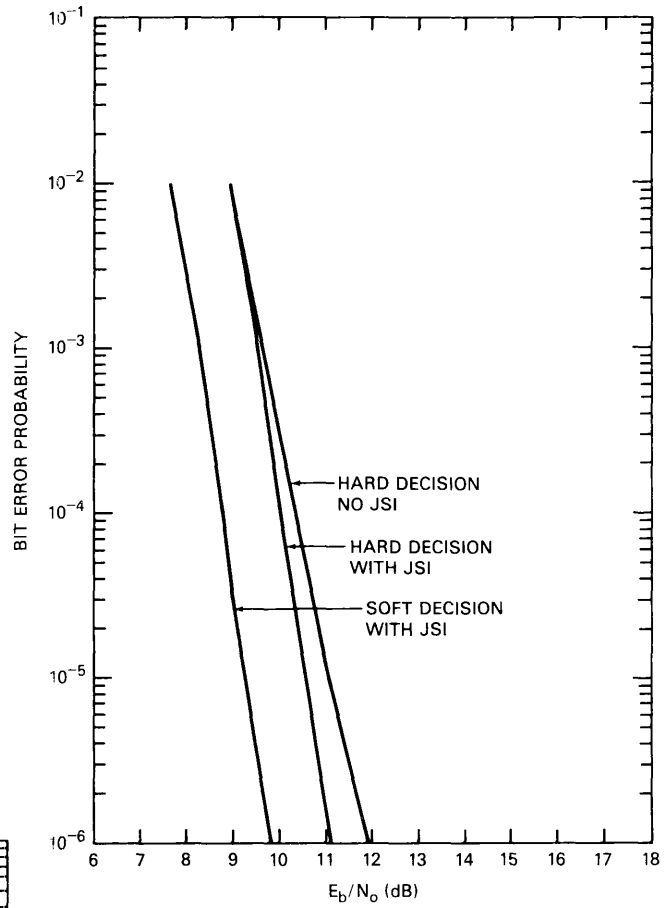

Fig. H6 — Performance of noncoherent FH 4-ary FSK; rate $r = 1$ constraint length 7 optimum 4-ary convolutional coding; WCPB noise jamming; nonfading channel; diversity $m = 3$

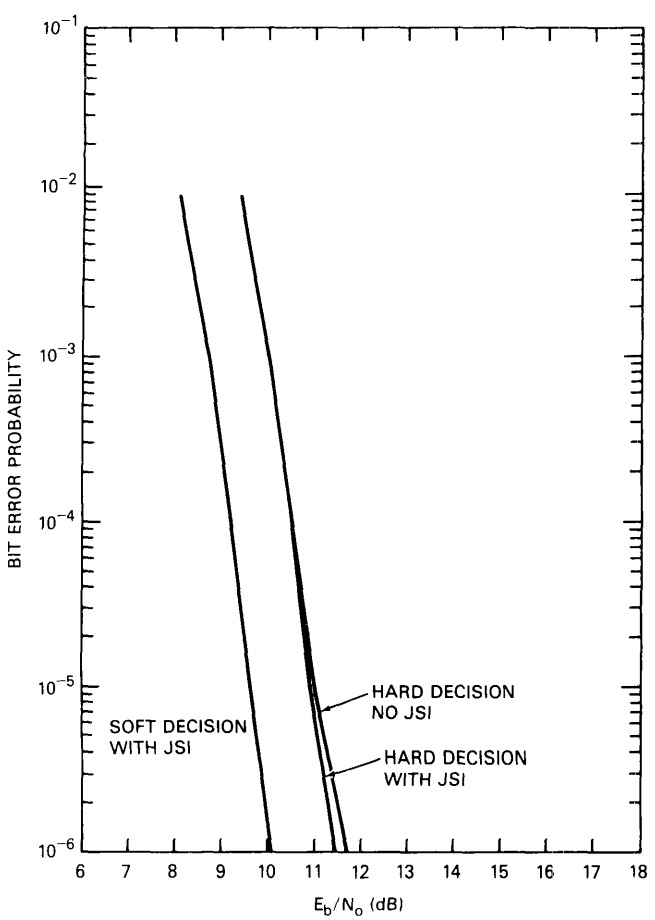

Fig. H7 — Performance of noncoherent FH 4-ary FSK; rate $r = 1$ constraint length 7 optimum 4-ary convolutional coding; WCPB noise jamming; nonfading channel; diversity $m = 4$

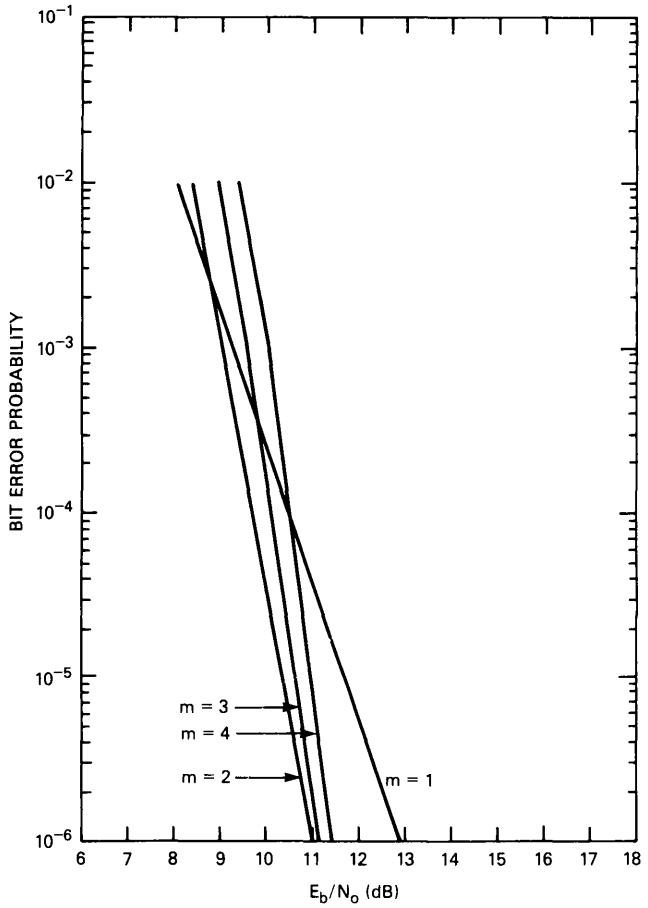

Fig. H8 — Performance of noncoherent FH 4-ary FSK; rate $r = 1$ constraint length 7 optimum 4-ary convolutional coding; WCPB noise jamming; nonfading channel; hard decision receiver; J.S.I. known; several values of diversity

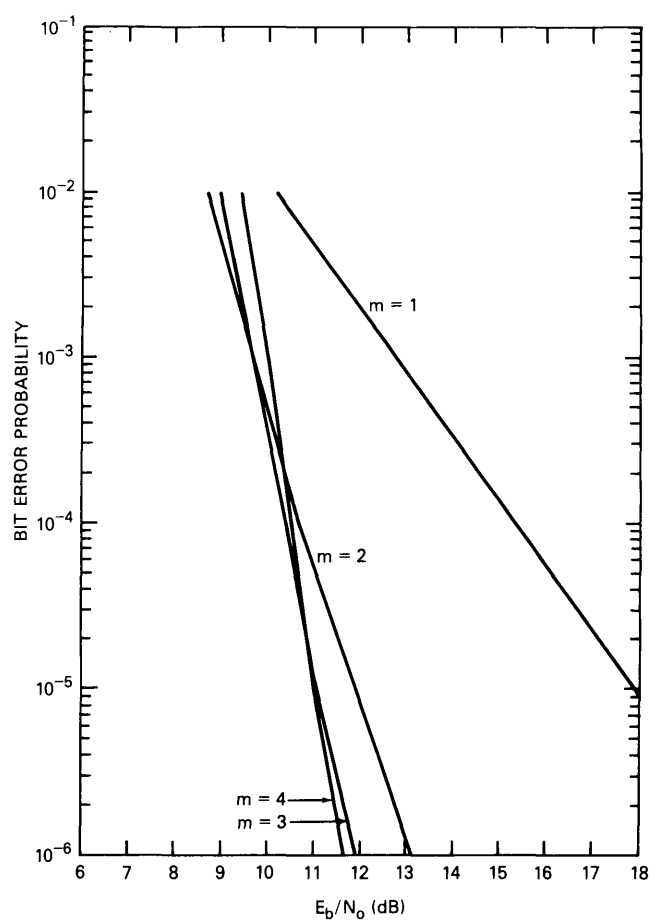

Fig. H9 — Performance of noncoherent FH 4-ary FSK; rate $r = 1$ constraint length 7 optimum 4-ary convolutional coding; WCPB noise jamming; nonfading channel; hard decision receiver; no J.S.I., several values of diversity

In Fig. H9 we see that for the case of BER = 10^{-5} failure to use diversity in a hard decision receiver without *J.S.I.* results in a 6 dB degradation as compared with the case of $m = 2$. This degradation is much more severe than for the other two receiver structures for which it is only about 1.3 dB. The use of diversity is therefore more important in hard decision systems without *J.S.I.* than it is in either hard or soft decision systems with *J.S.I.* The use of diversity results in a reduced information data rate if the signaling rate (in M-ary symbols per second) is kept constant, as we assume it is. Thus, diversity values in excess of 2 may not be worthwhile, because their use results in at best only small reduction in E_b/N_0 values, while significantly lowering the data rate.

In conclusion, for the case of a hard decision receiver without *J.S.I.* (the least sophisticated receiver structure we are considering) and 4-ary FSK a diversity of $m = 2$ appears to represent a favorable tradeoff between data rate and BER performance. In Fig. H5 we see that such a system, which is the simplest to implement, comes within 3 dB of the performance of an ideal soft decision receiver with *J.S.I.* for BER = 10^{-5}.

H2 RAYLEIGH FADING CHANNELS

Figures H10 through H12 show the cutoff rate parameter R_0 for Rayleigh fading channels for broadband noise jamming (which is in fact WCPB noise jamming for Rayleigh fading channels) as a function of $E\{E_c/N_0\}$ (the expected chip energy to noise density ratio) for binary, quaternary, and 8-ary signaling, respectively. Avidor has shown that the performance of the soft decision receiver without *J.S.I.*, is identical to that with *J.S.I.* provided that the fraction of the band that is jammed is known. Also, note that only a single curve is shown for the hard decision case; this is because under broadband jamming all chips are jammed, and so *J.S.I.* is in fact available. Also shown are curves for quantizer-limiter and limiter receivers, which are discussed in [H3]. The performance curves for the ideal soft decision receiver can be translated to the hard decision receiver by following the same procedure presented for the nonfading case.

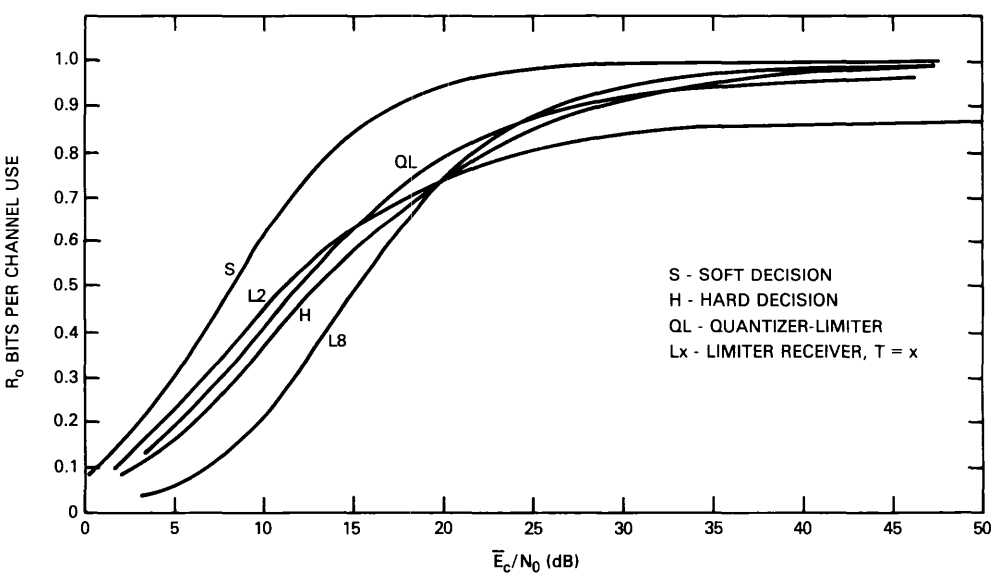

Fig. H10 — R_0 for FH-MFSK signaling ($M = 2$); Rayleigh fading channel; Broadband noise jamming (from [H3])

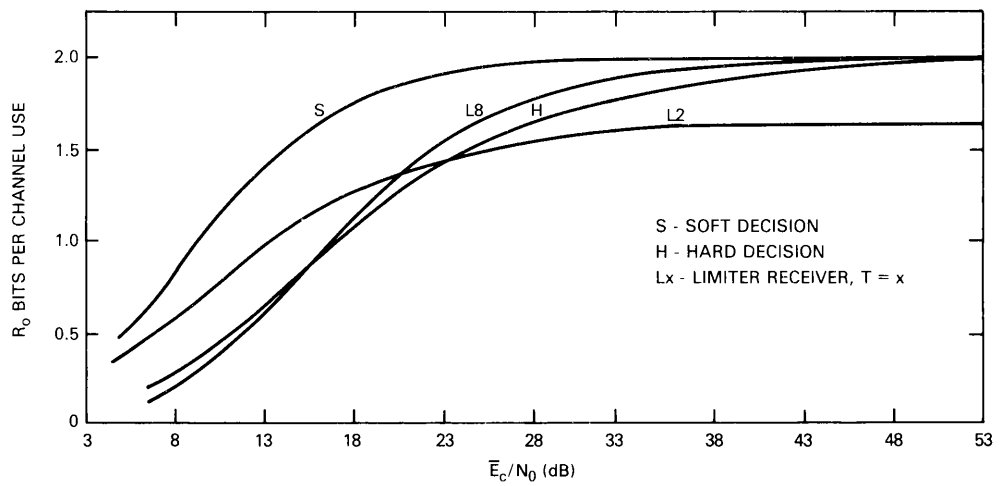

Fig. H11 — R_0 for FH-MFSK signaling ($M = 4$); Rayleigh fading channel; Broadband noise jamming (from [H3])

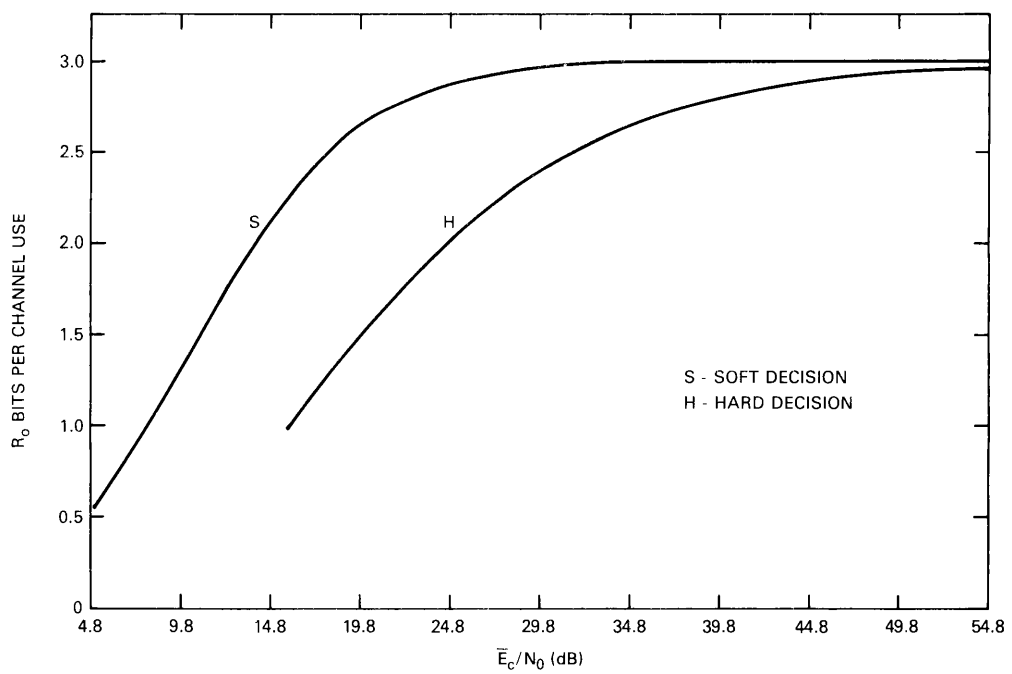

Fig. H12 — R_0 for FH-MFSK signaling ($M = 8$); Rayleigh fading channel; Broadband noise jamming (from [H3])

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Appendix I

PACKET ERROR PERFORMANCE ACHIEVABLE ON MULTIPLE USER FREQUENCY HOPPING CHANNELS

The use of spread spectrum signaling permits the sharing of a wideband channel by means of Code Division Multiple Access (CDMA) techniques. The number of users that can share a wideband channel simultaneously and the resulting performance depend on the modulation/coding scheme, channel characteristics, and receiver implementation. In a frequency hopping (FH) multiple user system *frequency hits* (i.e., the collisions of fractions of packets transmitted during individual frequency dwell times) will occur unless network-wide synchronization at the hop level is maintained and orthogonal hopping patterns are used. As discussed in Section 10.3.3, even if the FH patterns are orthogonal it will not be possible to maintain such hop-level synchronization in the HF ITF Network if hopping rates greater than about 300 hops/s are used. Therefore, frequency hits can be expected even with dedicated links, unless the hopping rate is sufficiently low. Because the hopping rates that will be employed in the HF ITF Network have not yet been determined, it is best to consider the general case in which network-wide hop level synchronization cannot be implemented.

In this appendix we consider FH spread spectrum multiple user channels in which Reed-Solomon coding is used to correct burst errors caused by other-user interference in a packet-switched environment. The model we present is based on the work of Pursley and Geraniotis [I1-I5] and Hajek [I6].

A slow FH system is considered, i.e., one in which several binary symbols are transmitted per hop. We address only the case of a fixed number of users that continuously transmit over the channel, although this model can be extended to include the case of bursty users, as was done by Hajek [I6]. Therefore, randomness in the system arises not because of changing user demands, but rather because of the pseudorandom nature of the hopping patterns.

In [I1-I6] two models are used to consider the effect of other-user interference on individual hops. In the first, all hops that are affected by other-user interference are assumed to be received with undetectable errors. In the second model it is assumed that side information is available, i.e., knowledge of which of the received hops have been corrupted by other-user interference. Such hops can be erased, thereby resulting in considerable performance improvement, because a Reed-Solomon codeword can tolerate twice as many symbol erasures as symbol errors. Wieselthier and Ephremides have proposed a third model for other-user interference in which hops can be correctly received, despite the partial overlap by other users' signals, provided that the overlap is a sufficiently small fraction of the hop duration [I7-I10]. This ability to discriminate against interfering signals results in further substantial performance improvement. We discuss the situations in which such a model would be applicable, although we acknowledge that its applicability to the HF ITF Network may be limited as a result of both the sophisticated nature of the receiver that is required, as well as the noisy signaling environment that characterizes the HF band.

I1 BASIC FH-CDMA MODEL

In FH systems the transmitter hops from one frequency bin or subchannel to another, transmitting a narrowband signal at each hop. It is assumed that there is no *spectral splatter,* i.e., that the spectrum of the transmitted signal is limited to its narrow band. The hopping pattern, which must be known by both the transmitter and receiver, is determined by a pseudorandom sequence. We assume that perfect synchronization is maintained between transmitter and receiver.

In FH-CDMA systems the code corresponds to the FH pattern. CDMA operation is usually asynchronous at the hop level, and therefore it is possible for two or more users (even when they use orthogonal hopping patterns) to transmit simultaneously in the same frequency bin, resulting in loss of data. Such collisions (of fractions of packets in this case as compared with whole packets in the usual case of *time-domain* multiple access schemes), are known as *frequency hits.*

Although network-wide synchronization at the hop level is not assumed, we note that it is reasonable to assume that the network is packet-synchronized. To do so requires guard times and a sufficiently accurate clock on each platform. As discussed in Section 10.3.1, we believe that this is a reasonable assumption for the HF ITF Network in which overall time uncertainties may be approximately 3 ms, while packet lengths may be tens of milliseconds or greater.

We consider a slow FH system, i.e., one in which several binary symbols (or the equivalent in M-ary symbols) are transmitted per hop. Errors caused by other-user interference thus tend to occur in bunches, because hits usually affect more than one binary symbol in a hop. The use of Reed-Solomon codes, which are powerful burst error correcting codes, is thus a good choice to correct the errors caused by frequency hits. We consider packet sizes chosen so that the data of a packet can be encoded as a single Reed-Solomon codeword. Each packet is divided into n *bytes,* one of which is transmitted at each hop. Each byte consists of $K = \log_2 M$ binary symbols, and is equivalent to an M-ary Reed-Solomon symbol. A Reed-Solomon (n,v) code can correct *any* pattern of as many as $t = (n-v)/2$ byte errors is a codeword (packet). It is thus a *maximum distance separable* code, and as such is considerably more powerful than binary (e.g., BCH) codes [I11]. It is important to note that, from the standpoint of codeword error probability, all byte errors are equivalent, regardless of the number of binary symbols within the byte that are in error.

We have considered two RS codes of rate approximately 1/2. The RS-(31,15) code with $M = 32$ corresponds to a packet length of 155 bits;* the packet is divided into 31 five-bit bytes, one of which is transmitted at each hop. This code is capable of correcting as many as eight byte errors in any codeword (packet). Similarly, the RS-(255,127) code corresponds to a packet length of 2040 bits, which are divided into 255 eight-bit bytes. This code is capable of correcting as many as 64 byte errors.

We note that each K-bit byte may consist of either K bits transmitted serially, a single M-ary signal (e.g., an M-ary FSK tone), or K bits transmitted in parallel (e.g., K parallel binary FSK tones). The model we present in this subsection is valid for all of these signaling schemes.

The wideband channel consists of q orthogonal frequency bins. The FH patterns are assumed to be generated by a first order Markov process such that each frequency bin is different from the previous one, but equally likely to be any of the $q-1$ other frequency bins. Alternatively, purely random schemes in which consecutive hops can be in the same frequency bin can also be considered, with little difference in performance.

Initially we assume that all frequency hits result in byte errors. This is a pessimistic assumption, because relatively strong signals will certainly have lower error probability than weaker signals, and because partially overlapping signals will not necessarily destroy the affected bytes. We consider a channel model that includes noise in addition to other-user interference. We define

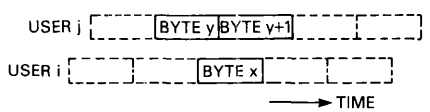
$$p_0 = \text{byte error probability resulting from noise;}$$

thus p_0 is the byte error probability in the absence of other-user interference. We assume that all byte errors within a codeword are independent, whether caused by channel noise or by other-user interference.

*In the present discussion a *bit* is equivalent to a binary symbol; we do not distinguish between information bits and transmitted binary symbols as we did in the AJ discussion of Section 7 and Appendixes E — H.

As a result of the lack of byte-level synchronization, a hit can result from either of two consecutive bytes from another channel user, as shown in Fig. I1. The relative delay among bytes involved in a frequency hit is assumed to be uniformly distributed between zero and the hop duration. (We neglect, for the moment, the possibility of multipath signal reception; however, we shall consider it shortly.) The probability that neither of these two bytes of user j will interfere with the transmission of user i, under the assumption of the first order Markov hopping patterns discussed earlier, is $(1 - 2/q)$. The resulting byte error probability given that k *other* users are simultaneously transmitting over the channel is, as discussed by Geraniotis and Pursley [I5] and Hajek [I6].

$$p_k = 1 - (1 - 2/q)^k \, (1 - p_0). \tag{I1}$$



A FREQUENCY HIT OCCURS IF EITHER BYTE y OR BYTE y + 1 OF USER j IS TRANSMITTED IN THE SAME FREQUENCY BIN AS BYTE x OF USER i

Fig. I1 — The occurrence of frequency hits

The byte error process for any user is a sequence of independent Bernoulli trials, because of the pseudorandom nature of the hopping patterns and therefore of the interference process. Since the Reed-Solomon codes under consideration can tolerate any pattern of t byte errors in an n-byte codeword, the packet error probability, given k other users, is equal to the probability that there are more than t byte errors:

$$\Pr(E|k) = \sum_{i=t+1}^{n} \binom{n}{i} p_k^i \, (1 - p_k)^{n-i}. \tag{I2}$$

Virtually all packet errors are detectable, as discussed in Appendix B, often permitting the subsequent retransmission of packets that are received with uncorrectable errors. In such cases we can interpret the packet error probability as a *packet erasure* probability.

Considerable improvement can be obtained in a system that is capable of detecting which of the received symbols have been affected by frequency hits, and of erasing the corresponding bytes. When both byte errors and byte erasures occur, a correct codeword decision can be made as long as the number of byte erasures plus twice the number of byte errors is not greater than $2t$.

The detection of hits would be straightforward in applications involving noiseless channels, i.e., channels in which the only interference is that which results from other users. For example, when binary FSK, is used, the presence of energy in both the mark and space tone positions (impossible for a valid signal) would indicate a hit.* When channels are noisy, it is more difficult to detect hits. The analysis of such a system would require the modeling of the probability of correctly detecting hits as well as the false alarm probability of incorrectly deciding that a hit has occurred in a noisy environment, in addition to a model for noise-induced byte error probability.

*If another user's signal destructively interferes with the desired signal so as to essentially cancel it, and if a third signal is present, then the latter may be perceived as the desired signal. For such an undetected symbol error to occur in a noiseless channel, a number of conditions have to be satisfied: the interfering signal would have to be in the same tone position as the desired signal, it would have to overlap for most of the hop duration, be close to 180° out of phase with the desired signal, and be of almost identical amplitude. We neglect such relatively infrequent events in our analysis.

In Fig. 10.6 we illustrated the probability of packet error as a function of the number of users that are transmitting simultaneously over a channel with 100 frequency bins. This figure demonstrates that the ability to detect frequency hits and erase the corresponding bytes results in a considerable increase in the number of simultaneous users that the FH channel can support.

The throughput of a FH channel of this type can be defined as the expected number of correct packets that can be delivered per time slot; a time slot is equal to a packet duration, and is therefore different for the different codes and packet sizes used. Under this model, it is assumed that packets incorrectly received will subsequently be retransmitted. We do not consider here the mechanism of retransmission for bursty users. We actually consider S, the throughput per subchannel, summed over all users, expressed in terms of packets per time slot and frequency bin, which, when a total of $k+1$ users are sharing the channel, is given by,

$$S = \frac{(k+1)}{q} [1 - \Pr(E|k)]. \tag{I3}$$

Channel throughput for the RS codes discussed above was shown in Fig. 10.8 for the noiseless channel, as well as for a channel in which the noise-induced byte error probability, p_0, is 0.1. Also shown in Fig. 10.8 was the noiseless case in which hits are recognized and erased, resulting in substantial performance improvement.

We now turn our attention to the modified model that permits within-hit discrimination against other-user interference, thereby improving performance considerably further.

I2 A REDUCED INTERFERENCE MODEL

We have also considered a model for other-user interference in which frequency hits can be detected, and hops (bytes) can be correctly received, despite partial overlaps by other users' signals, provided that the combined total overlap in any tone position is a sufficiently small fraction of the hop duration. A complete description of the signaling model, as well as a mathematical analysis and performance results, are presented by Wieselthier and Ephremides [I10]. Such an interference model is valid for a noiseless channel provided that the signal remains constant throughout each hop duration and that a sufficiently sophisticated receiver is used. We have investigated both M-ary FSK signaling and parallel binary tones. The ability to discriminate against interfering signals that are present for a sufficiently small fraction of the hop duration brings about dramatic increases in throughput as the tolerable hop overlap fraction is increased from 0 to 1.

The applicability of the increased throughput model to the HF ITF Network may be limited. First, the HF channel is certainly far from noiseless. In many cases, however, the other-user interference may be the predominant source of interference in the network; therefore, a noiseless channel model may in fact have some degree of validity. In the future we hope to be able to extend the model of [I10] to include the effects of channel noise. Another implementation difficulty is that a rather sophisticated receiver that is capable of extremely accurate synchronization is required to distinguish between the desired signal and the partially overlapping signals.

The comparison of performance under the three interference models clearly demonstrates that performance is strongly dependent on the receiver implementation. This study should provide insight into the benefits that are achievable through the use of receivers of different levels of sophistication.

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Appendix J

BIT ERROR PERFORMANCE ACHIEVABLE ON MULTIPLE USER FREQUENCY HOPPING CHANNELS; NONFADING AND FADING CHANNELS; NO CODING

In this appendix we discuss the raw (i.e., in the absence of coding) bit error performance that is achievable on FH spread spectrum multiple user channels. In such systems, bit errors may be caused either by frequency hits or by background noise. In particular, we consider the case of a single binary symbol transmitted per hop, noncoherent signaling (e.g., FSK), and either nonfading or fading channels. Performance curves for such a system were presented in Section 10.3.3.

We are especially interested in the class of slow, nonselective, Rician fading channels. In a Rician channel the signal consists of two components, a nonfaded or direct path component (in our case the HF groundwave signal) and a Rayleigh faded scatter component (the HF skywave signal). We also assume that Additive White Gaussian Noise (AWGN) is present at the receiver. This Rician model is quite general since it includes, as extreme cases, Rayleigh fading (where no direct component is present) and the nonfading AWGN channel (where no scatter component is present). Pursley and Geraniotis have studied nonselective- as well as selective- fading channel models for such systems [J1-J3]. Frequency selectivity (i.e., fading that varies with frequency) is especially important because it produces intersymbol interference within hopping dwell times. Its effects can by avoided, however, by transmitting symbols that remain constant throughout the dwell time of each hop. This can be accomplished either by using a sufficiently fast hopping rate so that only one (possibly M-ary) symbol is transmitted per hop, or at slower hopping rates by using multitone signals. A nonselective fading model is therefore appropriate in such cases, and in this report we consider only the nonselective case.

Although differential time delays between signal components cause frequency-selective fading, a nonselective model is appropriate for the modeling of F-layer skywave interference as long as the hopping rate is sufficiently high, even if the transmitted signal does not remain constant throughout the hop duration. Whenever the differential time delay between signal components is greater than the dwell time, the receiver will have hopped out of the particular frequency slot before the arrival of the delayed skywave component; therefore, there will be no intersymbol interference problem, except perhaps for a small fraction of the number of hops in which the hopping pattern may return to the same frequency slot following a delay approximately equal to the differential path delay. A nonselective fading model is therefore appropriate in this case of long differential path delays. The delayed skywave signals can be modeled as an additional number of independent interfering signals. If we assume that the skywave signal consists of a single (possibly faded) component, then the total number of potentially interfering signals that are contending for channel use is twice the number of actual users.

Pursley and Geraniotis [J1-J3] have considered slow frequency hopping (SFH) (i.e., one or more symbols are transmitted per hop) binary FSK signaling in which N_b bits are transmitted serially in each hop (unlike the multitone case described in Appendix I). The model of [J1-J3] assumes that partial frequency hits affect only those bits that are actually involved in the collision. Such partial hits, i.e., hits in which there is overlap for only a fraction of the dwell time, can occur because the hopping process is asynchronous at the hop level, as discussed in Appendix I. The (uncoded) bit error probability for arbitrary N_b in the case in which partial hits affect all of the bits in the hop, and in which fading is nonselective, is in fact equal to that obtained for $N_b = 1$. To avoid unnecessary complication, here we consider only the case of $N_b = 1$, i.e., a single bit per hop.

The evaluation of bit error probability resulting from hits is approached by considering a bit from an arbitrary signal (say signal i) and evaluating the probability that it suffers a frequency hit. Following the notation of [J2,J3] we define

$$\mathscr{P} = \Pr \text{ (bit from signal j interferes with signal i).}$$

For the present case of $N_b = 1$ bit per hop,

$$\mathscr{P} = 2/q - 1/q^2 \text{ for random FH patterns.}$$

If there are K signals simultaneously sharing the channel, we define

$$\begin{aligned}\hat{\mathscr{P}} &= \Pr \text{ (bit from one or more of the K-1 other users interferes with signal i)}\\ &= 1 - (1 - \mathscr{P})^{K-1}\end{aligned}$$

We emphasize that in the present discussion we are considering the case of a fixed number of users, K, that are continuously and simultaneously transmitting FH signals. The statistical aspects of this situation arise only from the random nature of the FH patterns.

The bit error probability in the absence of other-user interference (i.e., caused by fading and noise) is also needed to evaluate the overall bit error probability of the multiple access channel. We define

$$p_0 = \text{bit error probability in the absence of other-user interference.}$$

Although it is not feasible to evaluate exact expressions for multiple access error probabilities, Pursley has obtained lower and upper bounds on raw (uncoded) bit error probability, P_b. These bounds do not require the assumption of equal signal levels, and are therefore not affected by the near-far power differential problem. He has also obtained close approximations that are applicable when the signals are of equal strength. The bounds that are independent of power levels are:

Lower Bound

$$P_b \geqslant P_L = p_0(1 - \hat{\mathscr{P}})$$

Upper Bound

$$P_b \leqslant P_U = P_L + (1/2)\hat{\mathscr{P}} = p_0 + (1/2 - p_0)\hat{\mathscr{P}}.$$

The factor of 1/2 is present because the probability that a hit results in a bit error can be no greater than 0.5. We now discuss the evaluation of p_0.

We consider nonselective Rician fading, and we define:

E_T = transmitted energy per bit

$\bar{E} = (\alpha^2 + 2\sigma^2)E_T = E$ {received energy per bit}

$\alpha^2 E_T = E$ {received direct path signal energy per bit}

$2\sigma^2 E_T = E$ {received scatter component energy per bit}

$\gamma^2 = \dfrac{2\sigma^2}{\alpha^2}$ = ratio of power in scatter component to power in direct path component

N_0 = one-sided noise spectral density.

We define the following measure of expected bit energy to noise density:

$$\Lambda = \bar{E}/N_0 = (\alpha^2 + 2\sigma^2)E_T/N_0.$$

Then, for noncoherent binary FSK,

$$p_0 = \frac{\gamma^2 + 1}{\gamma^2\Lambda + 2(\gamma^2 + 1)} \exp\left\{\frac{-\Lambda}{\gamma^2\Lambda + 2(\gamma^2 + 1)}\right\}$$

(For differential phase shift keying, DPSK, the same result applies with $\Lambda = 2\bar{E}/N_0$).

In the expression for $p_0, \gamma^2 = 0$ corresponds to an AWGN channel and $\gamma^2 = \infty$ corresponds to a Rayleigh fading channel. In these limiting cases,

$p_0 = (1/2) \exp(-E/2N_0)$ AWGN

$p_0 = 1/(2 + \Lambda)$ Rayleigh Fading

Figure J1 shows the dependence of p_0 on γ^2 and Λ. p_0 is quite insensitive to changes in γ^2 for $\gamma^2 < 0.01$ and $\gamma^2 > 1.0$.

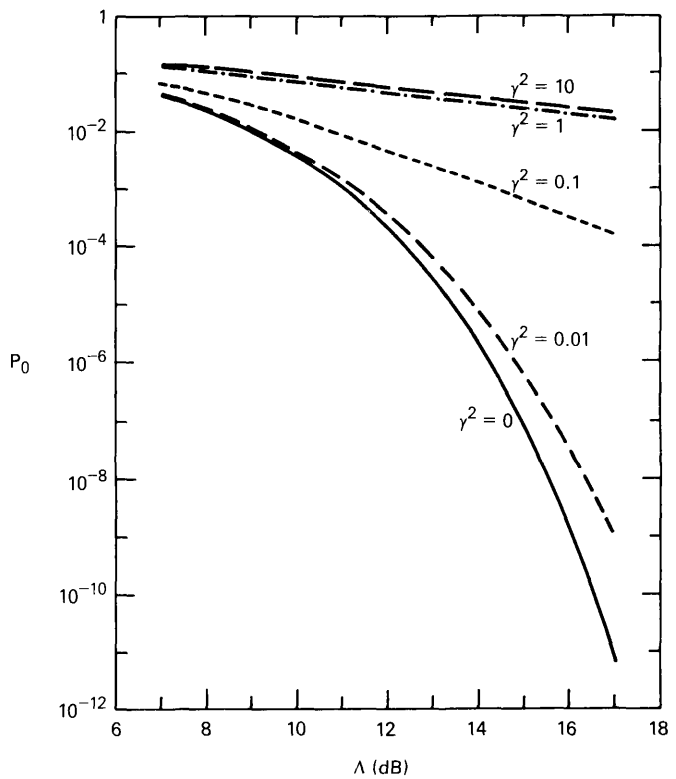

Fig. J1 — Bit error probability for Rician fading channels (no other-user interference)

The bounds on bit error probability are evaluated by substituting p_0 into the lower and upper bounds presented above. The upper bound is certainly of greater significance. It has been plotted as a function of Λ in Figs. 10.3 through 10.5 for the case of $N_b = 1$ bit per hop, $q = 100$ frequency slots, for $K = 1, 2, 5, 10$, and 20 active users, for three types of channels, i.e., AWGN, Rician fading with $\gamma^2 = 0.1$, and Rayleigh fading.

These figures show clearly that, at sufficiently high values of signal to noise ratio, the effect of other-user interference dominates the effects of noise and fading in producing bit errors.

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