Potential Use of Spread Spectrum Techniques in Non-Government Applications

Walter C. Scales

Prepared for the
Federal Communications Commission
Washington, D.C.

by
The MITRE Corporation
MTR-80W335

The following information is required by contract to be printed on the cover of this report.

1. Name and business address of contractor:

   The MITRE Corporation
   Metrek Division
   1820 Dolley Madison Blvd.
   McLean, VA 22102

2. Contract No.: FCC-0320

3. Contract dollar amount: $55,652

4. Contract award: Sole source

5. Project Officer:

   Dr. Michael Marcus
   Office of Science and Technology
   Federal Communications Commission
   2025 M Street, N.W.
   Washington, DC 20554

6. MITRE personnel responsible for the content and preparation of this report: Walter C. Scales

Technical Report Documentation Page

1. Report No.	2. Government Accession No. PB 81-165 284	3. Recipient's Catalog No.
4. Title and Subtitle Potential Use of Spread Spectrum Techniques in Non-Government Applications		5. Report Date December 1980
		6. Performing Organization Code
7. Author(s) Walter C. Scales		8. Performing Organization Report No. MTR-80W335
9. Performing Organization Name and Address The MITRE Corporation Metrek Division 1820 Dolley Madison Blvd. McLean, VA 22102		10. Work Unit No. (TRAIS)
		11. Contract or Grant No. FCC-0320
12. Sponsoring Agency Name and Address Federal Communications Commission 2025 M Street, N.W. Washington, DC 20554		13. Type of Report and Period Covered Final Report December 1979-December 1980
		14. Sponsoring Agency Code FCC/OST

15. Supplementary Notes

16. Abstract

Spread spectrum technologies have evolved primarily in military and aerospace environments, where the usual objectives are to provide resistance to jamming, multi-path and interception. Spread spectrum techniques may also have applications that fall under the jurisdiction of the FCC. In some civil applications, spread spectrum can provide communication privacy and multiple access on a transmit-at-will basis. In certain cases, spread spectrum may be a pratical, spectrum-efficient approach, although spread spectrum hardware is not currently available on a commercial basis. An example is given of a land mobile communication system using slow frequency hopping. Simulation results indicate that this technique might provide short-range service in applications when random channel access is an important feature.

17. Key Words Spread Spectrum, Spectrum Efficiency, Frequency Hopping	18. Distribution Statement Document is available to the public through the National Technical Information Service, Springfield, VA 22161

19. Security Classif. (of this report) U	20. Security Classif. (of this page) U	21. No. of Pages	22. Price $17.00 $3.50 microfiche

Form DOT F 1700.7 (8-72) Reproduction of completed page authorized

Potential Use of Spread Spectrum Techniques in Non-Government Applications

Walter C. Scales

December 1980

MTR-80W335

CONTRACT SPONSOR:
Federal Communications Commission
CONTRACT:
FCC-0320
PROJECT:
14570
DEPT.:
W-47

The MITRE Corporation
Metrek Division
1820 Dolley Madison Boulevard
McLean, Virginia 22102

TABLE OF CONTENTS

TABLE OF CONTENTS
(Continued)

TABLE OF CONTENTS
(Continued)

TABLE OF CONTENTS
(Continued)

TABLE OF CONTENTS
(Continued)

LIST OF ILLUSTRATIONS

LIST OF ILLUSTRATIONS
(Continued)

LIST OF ILLUSTRATIONS
(Continued)

LIST OF ILLUSTRATIONS
(Concluded)

1. EXECUTIVE SUMMARY

This summary presents the important features of the report in concise, nontechnical terms. In order to provide for quick, easy reference, the individual paragraphs of the Executive Summary are written in question-and-answer form. The reader who is interested in detailed technical results should turn to Section 2.2, Report Organization, for guidance.

1.1 What are Spread Spectrum Techniques?

The term "spread spectrum" has been applied to a wide variety of electronic systems and techniques. The common thread uniting these diverse areas of technology is that each uses signals requiring significantly more radio frequency bandwidth than a conventional signal would require. The expanded bandwidths provide certain features and characteristics that would otherwise be difficult or expensive to attain. Spread spectrum techniques are not new; they have been evolving since the late 1940s.

1.2 What are the Advantages of Spread Spectrum and How is it Used?

Spread spectrum techniques have evolved primarily in military environments, where they have been used to provide one or more of the following features:

- Resistance to Jamming
- Resistance to Unintentional Interference
- Resistance to Unauthorized Interception
- Sharing of a Common Radio Frequency Band by Multiple Users
- Discrete Addressing
- Accurate Distance or Location Measurements
- Pulse Compression

A more complete explanation of these features can be found in Section 2.3.1.

1.3 Why Would Spread Spectrum Techniques be Considered for Non-Government Use?

Aside from the threat of intentional jamming, many of the features and characteristics of spread spectrum have found potential applications outside of the military environment. These include nonmilitary aerospace applications now in the planning stages and several non-Government applications that have been proposed, but not implemented. (The term "non-Government," as used in this report, refers to applications outside the Federal

Government for which FCC authorization would be required.) Examples of potential non-Government applications are presented in Sections 3 and 4.

1.4 Are Spread Spectrum Systems Efficient in Their Use of Spectrum Resources?

Since bandwidth is an important measure of the spectrum resources used by a signal, the idea that a given signal should occupy more bandwidth, rather than less, runs against intuition and existing regulatory policy. Indeed, if each of many spread spectrum signals were allowed to occupy a separate, dedicated band, the resulting waste of spectrum resources would be unthinkable. However, because spread spectrum systems are resistant to interference, several spread spectrum signals can coexist simultaneously in a single, common band.

The question of whether this mode of spectrum-sharing is more effective than the existing frequency-channelized approach is not simple. Certainly, it is easy to produce examples in which spread spectrum systems make poor use of the spectrum, even when as many signals as possible share a common bandwidth. In some applications, spread spectrum signals suffer a fundamental, theoretical disadvantage when compared to frequency-channelized signals on the basis of spectrum utilization.

But the usual mode of sharing the radio spectrum on the basis of frequency channelization and geographic separation has practical limitations that prevent it from achieving the degree of efficient spectrum utilization predicted by simple theoretical models. These limitations include the need for guard bands between channels, the difficulty of maintaining a uniform demand or traffic loading on all channels, and the need for geographic gaps between regions in which a particular channel is re-used. For this reason, it is possible to produce examples in which spread spectrum techniques make more effective use of the spectrum than the conventional frequency-channelized approach. Naturally, it is also possible to improve the spectrum efficiency of the existing frequency-channelized approach without resorting to spread spectrum techniques.

When comparing practical implementations of spread spectrum and conventional narrowband technologies, it becomes evident that both approaches require a compromise between communication performance (or radiolocation performance, in the case of radiolocation systems) and efficiency of spectrum utilization. Increasing the number of users in a given band increase the utilization of the spectrum, but also increases interference, thus degrading performance. For this reason, such comparisons must be carefully formulated if the results are to be meaningful.

1.5 Can the Benefits of Spread Spectrum Systems be Achieved With Conventional Technologies?

In some cases, some of the advantages of spread spectrum systems can be achieved by other techniques. One example is resistance to unauthorized interception, which is readily achievable with conventional secure voice or data encryption techniques. On the other hand, spread spectrum techniques offer a unique method of sharing a common band between multiple users without requiring the users to coordinate their transmissions in any way. Spread spectrum systems have features and characteristics that differ from those of conventional narrrowband systems both qualitatively and quantitatively. These distinct features and characteristics are naturally better-suited to some applications than to others.

1.6 What Non-Government Applications Can Be Expected?

In the near-term, the applications that are most likely to engender practical implementations are those that require little development and offer attractive cost and performance characteristics. Most existing spread spectrum equipment was developed for military and aerospace applications. The functional and technical requirements for such applications are usually more severe than the corresponding requirements for commercial equipment. As a result, much of the existing spread spectrum equipment will be beyond the economic reach of commercial users, even if the equipment characteristics happen to be compatible with the application. On the other hand, commercial versions of existing designs could appear if the manufacturers perceive substantial markets.

In recent years, significant attention has been given to spread spectrum implementations of land mobile radio systems. However, the designs that have been proposed would require very substantial development efforts, and are not likely to be implemented in the foreseeable future. This report contains, in Section 4.1, a brief description of a simple spread spectrum technique for land mobile radio that could be implemented with significantly less development effort. This approach may entail certain performance compromises as a result of its simplicity, and thus may not be suitable for some applications. The performance and spectrum utilization aspects of this simplified approach are addressed in Section 4.1 and Appendix C.

A second application that has received considerable attention is the use of spread spectrum signals for distress alerting via satellite in the maritime mobile service. Although an experimental system has been developed and partially tested, implementation before the late 1980s is considered unlikely. A brief description of the proposed system is presented in Section 3.2.4.

There are probably a number of potential non-Government applications of spread spectrum technology that have not received serious attention simply because designers of commercial equipment are generally not well-versed in this area. A hypothetical example of such an application is described in Section 4.2.

1.7 What Economic Factors are Involved?

Most spread spectrum systems can be viewed as conventional narrowband systems to which special features have been added. This added level of complexity yields new performance characteristics at an increased cost. Clearly, the costs and benefits can be evaluated only on a case-by-case basis. However, the overall trend for the technologies used in spread spectrum systems has been one of decreasing costs during recent years. Non-Government spread spectrum systems, if developed, could be significantly less expensive than their military counterparts. Section 5.3 provides a preliminary assessment of the relative costs of several spread spectrum techniques that have been proposed for non-Government applications.

1.8 What Are the Risks of Increased Interference?

Before addressing this question directly, it is important to realize that there are several ways in which spread spectrum frequency assignments might be made. In the simplest case, a particular band might be set aside for exclusive use by spread spectrum signals in a particular geographic area. This approach might be used, for example, when moderate bandwidths are required at microwave frequencies. The alternative is to "overlay" spread spectrum signals on bands that are already in use. This approach would rely on the interference resistance and low spectral density of spread spectrum systems to minimize mutual interference.

Current users of the spectrum are likely to be concerned about the risk of interference from spread spectrum signals, especially if the "overlay" approach to spectrum management is adopted. On the other hand, the ever-increasing demand for spectrum resources will make it increasingly difficult to find suitable "dedicated" bands for exclusive use by spread spectrum systems. One potential solution is to provide frequency allocations for spread spectrum systems in the Industrial, Scientific, and Medical (ISM) bands. With one exception, current users of these bands are not protected from unintentional interference. The technical aspects of potential spread spectrum operation in the ISM bands are presented in Section 3.4.

Regardless of the approach to spectrum management, spread spectrum systems are likely to suffer interference from other spread spectrum systems. The inherent interference resistance of spread spectrum techniques will make this situation acceptable in many cases. The performance of spread spectrum systems in the presence of various types of interference has been extensively studied and can be reliably predicted given the system parameters and propagation characteristics.

1.9 What Are the Risks With Respect to FCC Monitoring and Enforcement?

This issue encompasses several facets, including the ability to monitor message content and operational procedures, the ability to measure and monitor the technical characteristics of signals, the ability to collect statistical data on spectrum utilization, and the ability to identify a particular station as the source of a particular signal.

Individual spread spectrum signals are normally difficult to detect with conventional receivers, except within a relatively short distance of the transmitting antenna. Even if they are detectable, the spread spectrum signals will probably not be intelligible on a conventional receiver. Thus, the question of monitoring and enforcement is not trivial.

Nevertheless, the FCC has reçourse to regulatory measures that could eliminate or substantially mitigate this potential problem. Some of these measures are listed in Section 5.2. Within the proper regulatory framework, monitoring and enforcement should be no more difficult with spread spectrum signals than with conventional secure voice or encrypted data signals.

1.10 What Additional Information Is Available on Spread Spectrum?

The body of technical literature on spread spectrum techniques is extensive. Of several tutorial papers that have been published, the most recent can be found in the September 1978 issue of the IEEE Communications Society Magazine. (See Reference 9 in Section 7 of this report.) A textbook entitled Spread Spectrum Systems is also available (Reference 1).

2. INTRODUCTION

2.1 Study Background and Objectives

The demand for radio spectrum from virtually all sectors of society has grown steadily and is expected to continue to grow in the future. Against a background of limited spectrum resources, the established method of dealing with this steady pressure of increasing demand involves a carefully formulated combination of regulatory and technical measures. Such measures have included, among others:

- shifts to higher frequency bands,
- use of closer channel spacings (often coupled with more stringent frequency tolerance or narrower limits on occupied bandwidth, or both), and
- the use of more efficient modulation techniques.

Such measures have been introduced carefully with the intent of minimizing the economic burden on the user groups. Although most efforts at improving spectrum utilization have been directed at increasing the efficiency of frequency division channelization, other channelization techniques have been successfully implemented using time division or other methods. The efficient sharing of a given bandwidth for communication on a time-division basis requires that transmissions from the various users be synchronized. This approach has been used, for example, in sharing a satellite repeater. A common bandwidth can also be shared on a time basis without a network synchronization protocol (i.e., by random access), but this results in interference between users and is generally effective only at low values of channel utilization, that is, when the common bandwidth is unused most of the time. Sharing on the basis of both time and frequency has been demonstrated in "trunked" land mobile systems. Again, coordination between users is required.

Another method of sharing a common bandwidth between multiple users is code division multiple access (CDMA), which has also been known as spread spectrum multiple access (SSMA). In this approach, as in time division multiple access (TDMA), users occupy a common bandwidth. In CDMA, however, multiple users may occupy the common bandwidth simultaneously. The signals from the various users are separable because each is modulated with a unique underlying code, and the various user codes are very nearly orthogonal to one another. Synchronization, in a network sense, is not required as in TDMA.* In fact, the only

* However, synchronization on individual links is necessary whenever information is actually being communicated on those links.

coordination that is required in such a system is a simple protocol between pairs of users who are (or want to be) in mutual communication.

The idea that such techniques could be used to improve the efficiency of spectrum utilization and to provide new capabilities is not original. The basis for spread spectrum communication dates back to the late 1940s, and civilian applications were discussed during the 1950s. The body of technical literature on the subject is extensive, and much of it is unclassified, despite the fact that spread spectrum technology evolved in a military environment. But high equipment cost and a conservative regulatory environment for years prevented the serious consideration of non-Government spread spectrum applications. Both of these factors are changing. The advent of low-cost large scale integrated (LSI) circuits has driven the cost of communication equipment down to the point where frequency synthesizers are commonly used in citizens band equipment, including inexpensive hand-held portable transceivers. This would have been economically unthinkable twenty years ago. Entire receivers (less tuned circuits and electromechanical components) can be built on a single "chip." Whether or not this trend will continue at anything like its previous pace is beside the point. Short of an economic catastrophe, the trend is not likely to be significantly reversed. Furthermore, the current regulatory climate is one in which innovation is encouraged, subject to common-sense cavaets.

It is clear, then, that the time has come for a fresh look at spread spectrum technology as it might be applied under the FCC's regulatory domain. A certain amount of audacity would be needed in order to imagine that all such potential applications could be examined in a single study. The purpose here is to present, in concise form, a number of potential applications that have been described prior to this study and to suggest a few new potential applications that have not been previously considered for non-Government use. The benefits, costs, and risks associated with these various potential applications are also presented.

In any effort of this sort, it is inevitable that a focus will develop on a limited area, to the detriment or exclusion of other areas. Much of the work that has been performed on military and aerospace applications of spread spectrum technology will go unmentioned, simply because there is so much of it. The interested reader can pursue these areas with the aid of the list of references presented in Section 7 of this report, or through the bibliography presented in Dixon's textbook on spread spectrum systems (1).

2.2 Report Organization

This report is organized into six major sections, plus references and appendixes. The Executive Summary (Section 1) is a primarily nontechnical presentation designed to convey the results of this study in concise form. The Introduction (Section 2) is intended to provide the reader with a working vocabulary of terms that are normally associated with spread spectrum technology, as well as a general feeling for the contexts within which that technology normally arises. Section 3 describes the potential motivations for using spread spectrum techniques under the FCC's regulatory domain and summarizes various technical approaches that have been proposed or studied for such applications. Section 4 provides two hypothetical examples of non-Government spread spectrum implementations which have not been previously studied. Section 5 describes the potential costs and risks of FCC-authorized spread spectrum implementations. The study conclusions are summarized in Section 6. References are presented as Section 7. Finally, the purely technical and analytical aspects of this study are presented in three appendixes.

Appendix A describes the theoretical spectrum efficiencies attainable with spread spectrum multiple access under various conditions. In Appendix B, the impact of frequency tolerances on the spectrum efficiency of frequency-channelized systems is examined as a basis of comparison for spread spectrum multiple access systems. Appendix C presents the development and results of a Monte-Carlo simulation of a hypothetical spread-spectrum implementation described in Section 4.

2.3 Uses of Spread Spectrum

2.3.1 Functions

The term "spread spectrum" has been used to describe a variety of techniques and applications. The common thread that unites these diverse areas of technology is the use of signals having bandwidths that are far in excess of the bandwidth of the underlying information. The signals have a "fine structure" that is used by the receivers to separate them from interference, jamming, and other undesired components that may be present at the receiver input. Spread spectrum techniques have evolved primarily in military environments, where they have been used to provide one or more of the following features.

Resistance to Intentional Jamming. Potential jammers span the range from crude radio frequency generators to sophisticated devices that imitate the desired signals, either by amplifying and retransmitting them (repeater jamming) or by discovering and using their "fine structure." Although no system is completely immune to jamming, a properly-designed spread spectrum system can make it so expensive for the enemy to jam effectively that he will consider conventional military alternatives (e.g., destroying the transmitter or the receiver, assuming that he can find them).

Resistance to Unintentional Interference. Clearly, any system that is resistant to intentional jamming will also be resistant to unintentional interference from other communication signals, radar signals, spurious emissions, ignition noise, and other sources. Again, the resistance to unintentional interference is not absolute, but is typically high compared with conventional narrowband signaling.

Resistance to Interception. In some military applications, it is important that the existence (and location) of a transmitter not be discovered. Conventional narrowband signals can be detected, identified, and triangulated upon, even by an enemy having access to only crude and inexpensive equipment. Spread spectrum signals, on the other hand, can occupy such a large bandwidth that the average signal-to-noise ratio (measured in the signal bandwidth) is significantly less than unity, except within a small area in the immediate vicinity of the transmitting antenna. Under these conditions, the difficulty of detecting the signal is greatly increased.

Even in cases where it is possible to detect the presence of a spread spectrum signal, some degree of security can be provided by designing the signal so that it cannot be "decoded" by the casual listener. The possibility that messages will be recorded and played back with the intent of creating confusion or deception is also minimized.

Low Spectral Density. Interference to narrowband receivers from co-channel signals is minimized if the interference power per unit bandwidth is made small. This end can be achieved by spreading the fixed (spread spectrum) interference power over the widest possible bandwidth. The resulting interference to a narrowband receiver may still be greater than that caused by an adjacent-channel narrowband signal, but will be significantly lower than that caused by a co-channel narrowband signal of equal power.

Multipath Resistance. "Multipath" is a term that is used to refer to the existence of more than one propagation path between a transmitter and a receiver. Signals traversing the various paths arrive at the receiving antenna with random phase angles, so that at any moment these signal components may add constructively or they may cancel one another, leaving the total received signal far below its "normal" or "average" level. Such fading effects can significantly degrade the performance of narrowband communication systems. In addition, the existence of multipath propagation can significantly degrade the performance of narrowband ranging or radiolocation systems.

If the various propagation paths are all of different lengths, they can be separated by using signals of sufficiently large bandwidth. As a trivial example, consider a system in which information is transmitted in discrete pulses, each of which is much shorter than the difference between the propagation delays associated with any two paths. In this case, the receiver will "see" a sequence of pulses (one for each propagation path) in response to every transmitted pulse (see Figure 1). If the interval between transmitted pulses is sufficiently long, no two received pulses will overlap, so the problem of multipath fading will be eliminated. The receiver may use the first received pulse in each sequence, or the strongest pulse, or it may attempt to combine the pulses in such a way that fading is mitigated or eliminated. Spread spectrum techniques can also be effective for combating diffuse multipath, in which individual paths are not separable.

Of course, narrowband techniques have also been developed to mitigate multipath fading. The most familiar of these are "diversity" techniques, which include frequency diversity (where two or more narrowband signals are transmitted on separate channels and recombined at the receiver) and space diversity (in which two or more separate receiving antennas are used). Time diversity and polarization diversity techniques have also been used.

Multiple Access. This term traditionally refers to the use of a satellite repeater by multiple signals originating from transmitters that are spatially dispersed. In a more general context, it applies when access to a channel (or group of channels) is involved, whether or not there is a repeater. The various signals may be separated in frequency (frequency division multiple access or FDMA) or time (time division multiple access or TDMA). A third multiple access technique involves the use of the entire repeater bandwidth continuously by each signal. Because the various

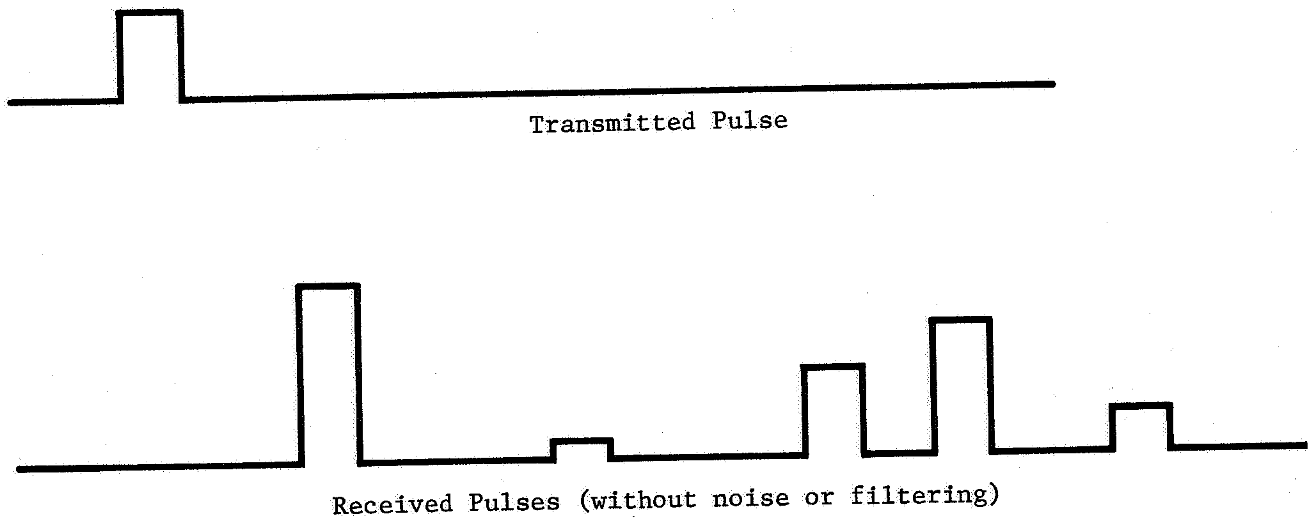

FIGURE 1
RESPONSE OF IDEALIZED MULTIPATH CHANNEL

signals overlap completely in both time and frequency, the receiver can separate the "desired" signal from the other signals only if each signal is carried by an underlying waveform that is very nearly orthogonal to all of the other signal waveforms. (Two waveforms are said to be orthogonal if their product, integrated or averaged over some fixed time period, is zero). In general, this arrangement is possible only when the signal bandwidth greatly exceeds the information bandwidth for each signal. Thus, the term "spread spectrum multiple access" (SSMA) is often used to describe this approach. Because digital codes are usually employed to form the underlying quasi-orthogonal waveforms, the term "code division multiple access" (CDMA) is also used. Such a system has the characteristic that it displays no distinct "saturation point" as the number of users is increased. Rather, the signal-to-interference ratio experienced by each user continuously declines as more and more users enter the system.

Another advantage is that the various users need not coordinate their transmissions, but simply transmit at will. SSMA techniques have the disadvantage that they often require greater transmitter power and greater bandwidth per user (for a given information rate) than FDMA or TDMA techniques.

Discrete Addressing. For some applications, it is desirable to code individual signals in such a way that they can be received only at their intended destinations. A spread spectrum system will provide this feature if each user is assigned a unique code that is employed to produce the underlying "fine structure" of his particular signal. The level of security provided by this approach is not normally comparable with cryptography, but protection from the casual listener can be reasonably assured.

Pulse Compression and Ranging. Although this application is not necessarily associated with "spread spectrum," the same underlying techniques are involved. Many radar systems transmit single, unmodulated pulses. The intrinsic range resolution in such systems improves with decreasing pulse width, but the signal-to-noise ratio also decreases with decreasing pulse width, since decreasing the pulse width, for a given peak pulse power, decreases the total energy of the pulse. Good range accuracies can be attained by using very high peak pulse power, but this imposes a corresponding economic burden.

An alternative approach is to design the transmitted pulse so that its width can be significantly reduced at the receiver. (Hence the term "pulse compression.") The resulting short pulse contains all of the energy of the longer transmitted pulse, but has the intrinsic range resolution implicit in the reduced pulse width. Again, the basic approach entails giving the transmitted pulse a "fine structure" that can be efficiently removed at the received.

2.3.2 Ongoing Programs

The spread spectrum concept, in its basic form, dates back to the late 1940s (1). Since that time, a substantial number of spread spectrum systems have been developed and implemented. Others are currently under development. A few of the more well-publicized applications are listed below. The list is by no means exhaustive.

JTIDS (Joint Tactical Information Distribution System). This system is currently being developed by the Department of Defense (DoD) to provide secure, jamming-resistant communications, navigation and identification for combat elements such as aircraft. The communication functions are digital, but a provision is made for digitized voice (2). JTIDS is of particular interest because its spectrum is "overlaid" on existing 960-1215 MHz aeronautical radionavigation assignments (3). The details of the band-sharing arrangement will be described in Section 3.3.5.

Packet Radio is a project of the Defense Advanced Research Projects Agency (DARPA) that is aimed at extending a technology known as "packet switching" into the area of radio communications. Spread spectrum signals, operating in the 1710-1850 MHz band, are used to provide jamming resistance, security, multiple access, and multipath protection (4). An experimental system has been under evaluation for several years.

Global Positioning System (NAVSTAR/GPS) is a DoD effort that will provide accurate satellite-derived location fixes anywhere in the world. Spread spectrum signals centered at 1227 MHz and 1575 MHz provide the basis for range measurements and data transfer. The features provided by the spread spectrum signal format include resistance to jamming and interference, multiple access, good range resolution, multipath rejection, and security (5). The system is currently under development.

Tracking and Data Relay Satellite System (TDRSS) is a NASA program aimed at providing improved control, communication, and ranging for low-orbiting satellites. Spread spectrum signals from the user satellites (including the space shuttle) are relayed to ground stations via TDRS. The spread spectrum signal format provides resistance to multipath and unintentional interference, multiple access, and good range resolution (6). TDRSS is scheduled to become operational in 1981.

Position Location Reporting System (PLRS) is an Army/Marine Corps UHF system designed to provide digital data and position location for land vehicles, aircraft, and manpack users. PLRS provides data security and resistance to jamming, as well as location and data communications.

Single-Channel Ground and Air Radio (SINCGARS-V) is the Army's developmental VHF voice radio system. SINCGARS-V will also have an ancillary capability for handling digital data. The system is designed to provide resistance to jamming and interception in communication between combat leaders on the battlefield.

2.4 Spread Spectrum Technology: A Brief Overview

A variety of technologies have been used in the design and implementation of spread spectrum systems. Tutorials on spread spectrum techniques have been published on several occasions (7,8,9) and a textbook on the subject is available (1). Rather than trying to duplicate any significant amount of this literature, this section presents a brief introduction to spread spectrum technology, with emphasis on developing the technical vocabulary that will be used in the remainder of the report.

2.4.1 Basic Techniques

Three basic techniques have been used in spread spectrum systems. They are known as frequency hopping (FH), time hopping (TH), and direct sequence (DS). Hybrids employing two or more of these techniques have also been implemented. A fourth basic technique, known as "chirp," or linear FM, will also be described. These techniques distinguish spread spectrum from other bandwidth-expanding techniques like pulse code modulation (PCM) and ordinary wideband FM.

2.4.1.1 Frequency Hopping

In this technique, the carrier frequencies of the transmitter and receiver are abruptly changed at regular intervals. The transmitter and receiver ideally switch carrier frequency at the same time. The transmitted sequence of carrier frequencies follows a pseudo-random pattern, as shown in Figure 2. The receiver attempts to remain tuned to this time-varying carrier frequency, so that the signal at the receiver's intermediate frequency output (Figure 3) is "de-hopped." Thus, the receiver must contain a stored replica of the transmitter's frequency hopping pattern.

A narrowband interfering signal will appear only occasionally at the receiver's i.f. output, as the frequency-hopping signal and the interferer briefly "collide" (Figure 4). (This assumes that the interfering signal is not so strong that it is still measurable after being attenuated in the stop-band of the receiver's i.f. filter.) If there are N separate hopping frequencies, each being used for an equal fraction of the transmission time, then the interference power, averaged over all hops, is reduced by a factor of N.

In principle, information can be modulated onto the transmitted carrier in any convenient form, since the carrier is de-hopped at the receiver. In practice, however, it is not always possible to maintain a constant or predictable carrier phase between frequency hops. For this reason, coherent demodulation techniques are seldom used. In analog FM transmission systems using frequency or phase modulation, the phase discontinuities between frequency hops produce background noise ("clicks") that cannot be mitigated by increasing the signal power (10).

For some purposes, it may be convenient to classify frequency hopping systems as fast hopping or slow hopping. This report will use the term "fast-hopping" when the hopping rate significantly exceeds the information rate. In systems of this type, multiple hops are typically added (i.e., integrated) to re-form the information signal. Examples will be given in Section 3.2.2. In slow-hopping systems, the hopping rate is comparable to or less than the information rate. An example of a slow-hopping system is presented in Section 4.1.

2.4.1.2 Time-Hopping

In this approach, the transmitter emits short pulses or bursts at pseudo-random times. The sequence of transmission times is stored in the receiver, which tracks and demodulates the transmissions but otherwise ignores the channel. Thus, a jammer must either discover the hopping pattern, or spread its power over a high duty cycle (thus wasting much of its energy) or be content with randomly jamming only a small fraction of the transmitted pulses. Repeater jamming can be thwarted by using sufficiently narrow pulses.

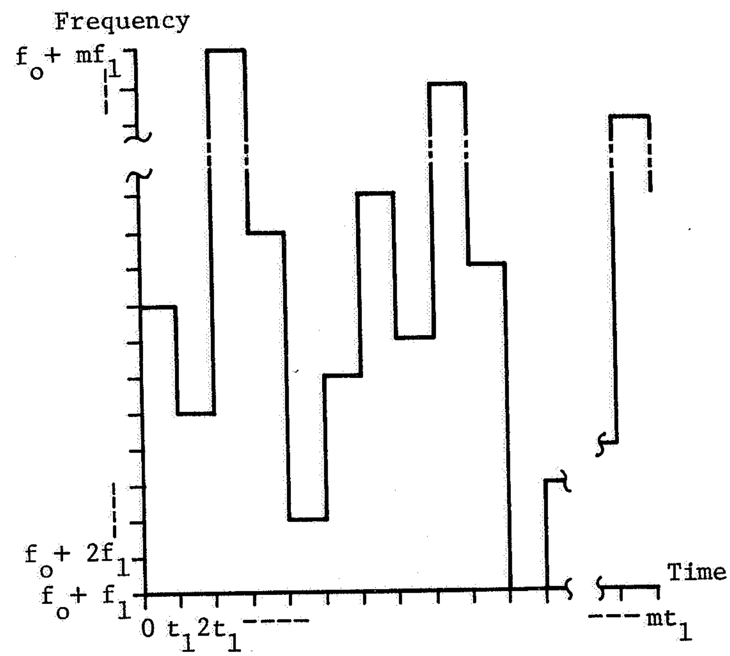

FIGURE 2
SIMPLIFIED FREQUENCY VS. TIME TRAJECTORY OF A FREQUENCY HOPPING CARRIER

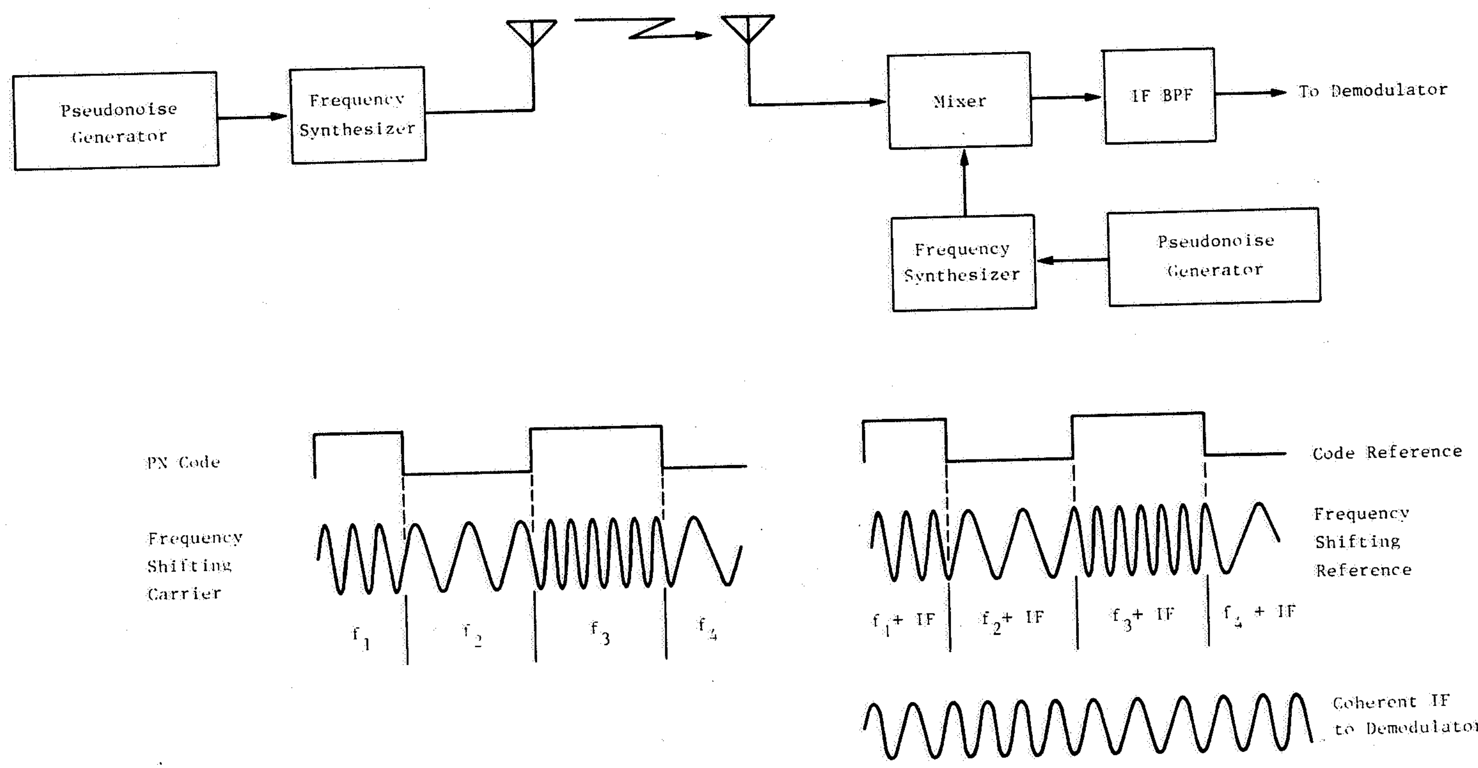

FIGURE 3
BASIC FREQUENCY HOPPING SYSTEM WITH WAVEFORMS

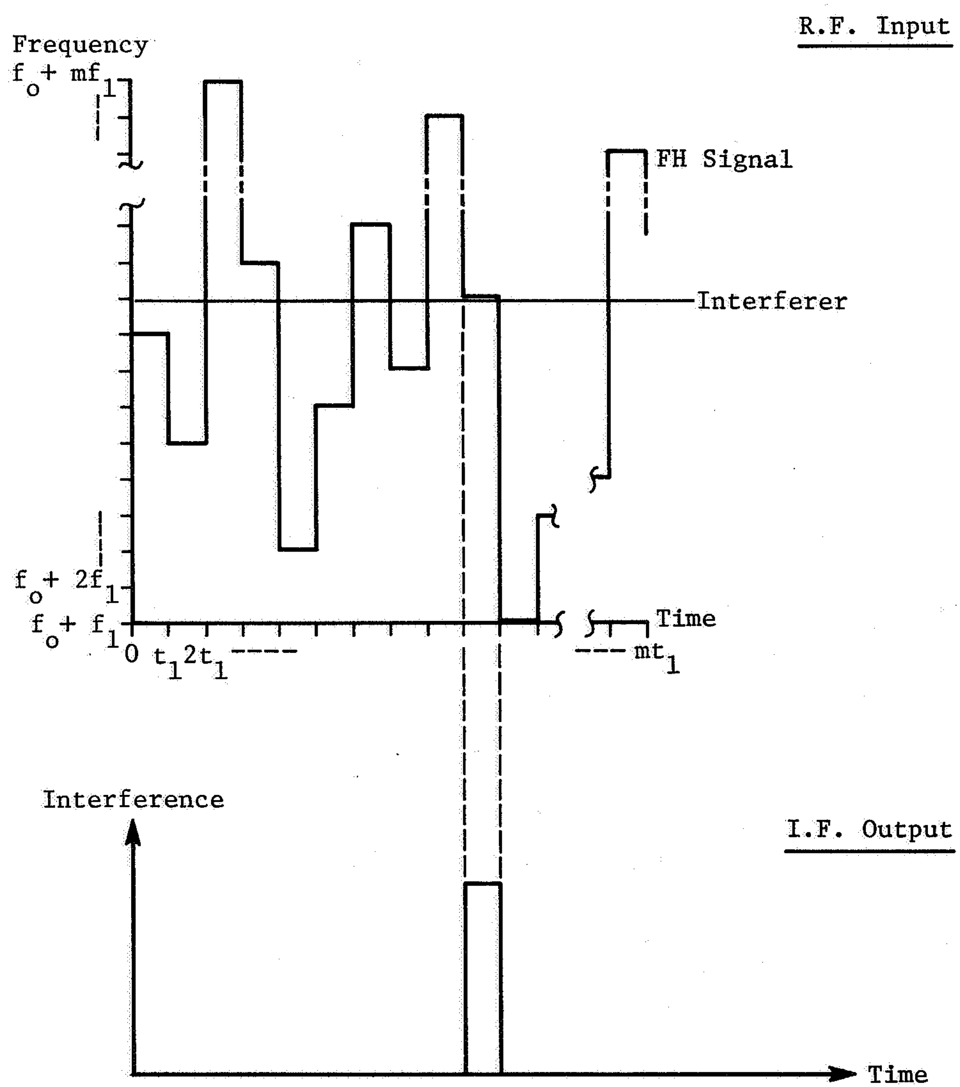

FIGURE 4
EFFECT OF NARROWBAND INTERFERENCE ON A FREQUENCY HOPPING SYSTEM (SIMPLIFIED)

Pure time-hopping has not found widespread application outside of the military sector.

2.4.1.3 Direct Sequence

This method, which is also known as pseudo-noise, or PN, consists of switching the phase of the transmitted carrier at regular intervals. In the simplest version, the carrier is switched between two phases that are 180 degrees apart, according to a pseudo-random binary pattern. The receiver tracks these pseudo-random phase inversions using a stored replica of the binary pattern, thus reproducing the original carrier. A period of constant carrier phase is called a "chip" in order to avoid use of the term "bit" in reference to both information bits and the smaller spread spectrum elements. The term "chip" has also been applied to a frequency hopping interval.

Figure 5 illustrates the operation of a simplified DS system. At the transmitter, the information signal is multiplied by a rapidly-switched sequence of chips. Its bandwidth is thus significantly expanded beyond the information bandwidth. At the receiver, the signal is restored to the information bandwidth by re-multiplying with the pseudo-random chip sequence. Interference, on the other hand, is subjected to only one multiplication process (this occurs in the receiver) so that its bandwidth is increased to at least the bandwidth of the chip waveform. Filtering the receiver output effectively removes interference energy beyond the information bandwidth.

Thus, the signal-to-interference ratio at the receiver output is higher than the channel signal-to-interference ratio by a factor that is approximately equal to the ratio of the spread spectrum bandwidth to the information bandwidth. This factor is known as "process gain" or "processing gain."

Other versions of DS spread spectrum involve the use of four or more carrier phases, or the use of two carrier phases that are separated by an angle other than 180 degrees.

In principle, virtually any conventional technique can be used to modulate the carrier with the information, either before or after spreading. In practice, most DS systems have been used for digital data transmission, although analog modulation systems have also been implemented.

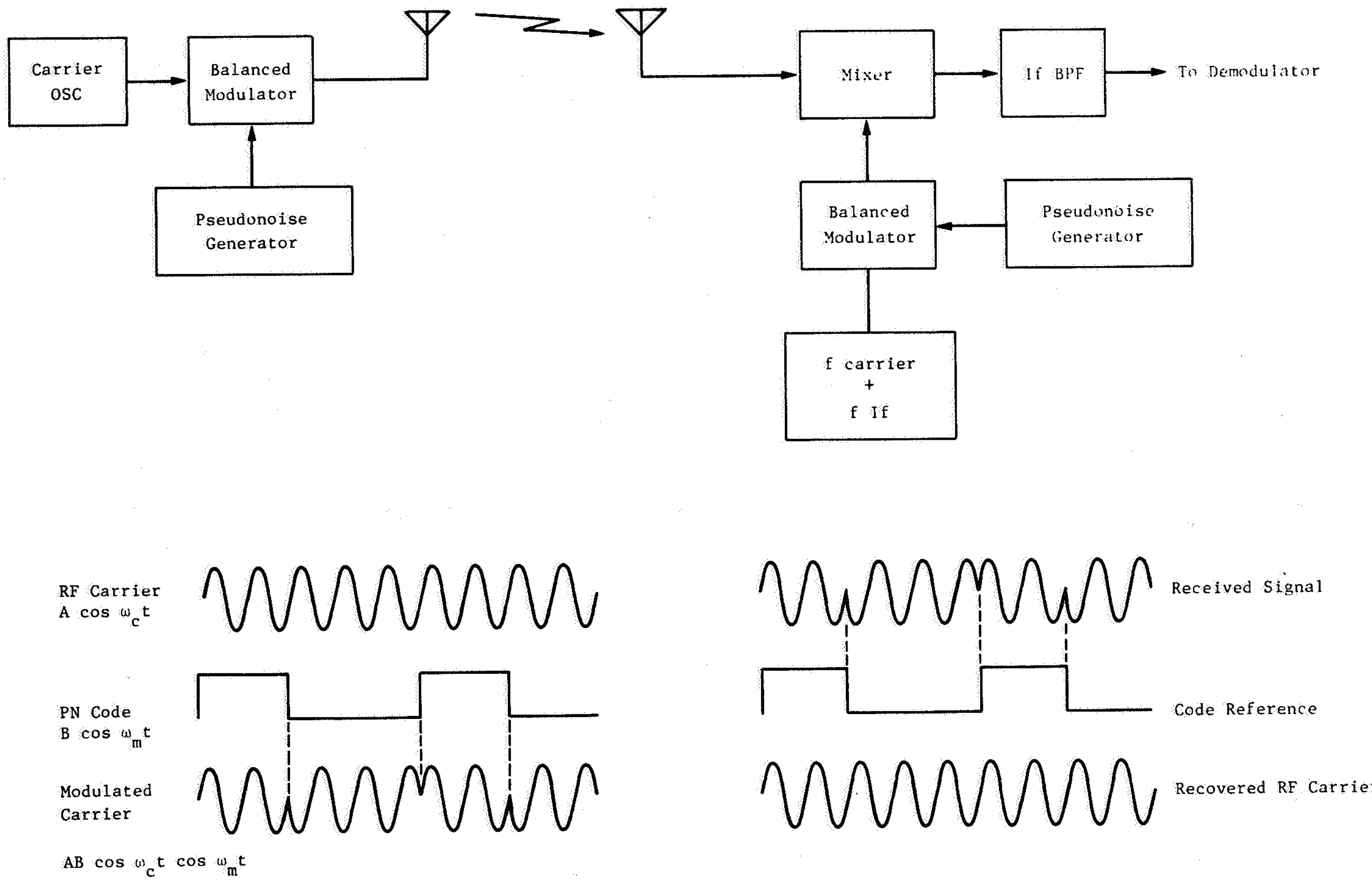

FIGURE 5
OVERALL DIRECT SEQUENCE SYSTEM SHOWING WAVEFORMS

2.4.1.4 Linear FM ("Chirp")

Linear FM techniques are based on constant envelope, swept-frequency waveforms. These techniques have traditionally been associated with pulse compression radar systems, and are not necessarily associated with the usual connotation of spread spectrum.

In the basic chirp technique, the carrier frequency of the transmitted pulse is linearly swept over a given bandwidth. At the receiver, the signal is processed in a dispersive delay line that compresses the pulse to a width roughly equal to the reciprocal of swept bandwidth. The gain in signal-to-noise ratio during the compression is approximately equal to the time-bandwidth product of the transmitted pulse. Time-bandwidth products on the order of 1000 have been demonstrated in chirp systems.

2.4.2 Discussion

At this point, it may be useful to make several observations about the techniques that have just been briefly described.

I. None of the spread spectrum techniques described above provides any improvement over narrowband techniques when the noise or unintentional interference has a constant, uniform spectrum over the entire spread spectrum band. (However, forcing a willful jammer to spread his limited power over this wider bandwidth is advantageous.) Thus, "pure" spread spectrum techniques do not provide performance improvements against background noise alone. However, digital spread spectrum systems are frequently used in conjunction with error-correcting codes that do provide such improvements.

Spread spectrum techniques can provide performance improvements against multipath-related effects, but only if they are specifically designed to do so.

II. Multiple access can be implemented with virtually any spread spectrum technique. In the usual implementation, each transmitter (or group of transmitters) is assigned a unique digital code that underlies either its hopping sequence or its phase-keying sequence. The set of codes is usually designed to minimize interaction between transmitters and receivers that are not in communication with one another.

III. It has been a common practice to combine two or more spread spectrum techniques to produce hybrid systems. Thus, an early concept known as RADA--Random Access Discrete Address--combined frequency hopping and time hopping (11). JTIDS and

PLRS uses a combination of frequency hopping and DS techniques. TDRSS and GPS, on the other hand, use pure direct sequence techniques (5,6).

IV. An important feature of slow frequency hopping is that advantage can be taken of the selectivity of inexpensive crystal filters. Unless a narrowband interferer is extraordinarily strong, it will produce interference only 1/N of the time, where N is the number of FH channels or frequency slots. Subject to the constraint just stated, the amount of interference produced is limited to a fraction that is independent of the signal-to-interference ratio at the receiver input.

In order to see the significance of this, consider a multiple access application in which a receiver must contend with a single interferer that is 40 dB stronger than the desired signal. (This is the so-called "near-far" problem. For this example, non-fading conditions will be assumed.) Suppose, furthermore, that a 10 dB output signal-to-interference ratio is required at least 95 percent of the time under these conditions. A direct sequence system would then require a processing gain of 40 dB + 10 dB = 50 dB. As a good approximation, this means that for an information bandwidth of 3 kHz a spread spectrum bandwidth of 300 MHz (= 3 kHz x 10^5) would be required. Aside from the fact that this is an extraordinarily large bandwidth, direct sequence chip rates on the order of 300 MHz are near the limit of the present state of the art. It is easy to produce examples in which multi-gigahertz chip rates would be required.

With a frequency hopping approach, the problem is substantially alleviated. As long as the design is such that adjacent channels do not "splatter" through the i.f. passband, or overcome the i.f. filter's stopband, or desensitize the receiver, or produce intermodulation products, the interferer (now assumed to be another slow FH signal) will produce interference only 1/N of the time. Thus, a properly-designed twenty-channel system would provide the required output signal-to-interference ratio 95 percent of the time. For an information bandwidth of 3 kHz, this number of channels would typically represent a total bandwidth of no more than 1.0 MHz, assuming a maximum channel spacing of 50 kHz. Furthermore, the technology required to implement such a system is well within the state-of-the-art.

2.4.3 Implementation Constraints

2.4.3.1 Synchronization

The problem of synchronizing the receiver with the incoming signal at the chip or hop level represents a major design issue in most spread spectrum applications. Although most conventional digital receivers require synchronization at the bit level, synchronization at the chip or hop level is often more difficult to achieve. For direct sequence systems, the SNR for a single chip is typically less than unity. In addition, the frequency of the received carrier may not be known accurately enough to allow the coherent integration of a large number of chips. Thus, it may be necessary to perform an exhaustive search over two dimensions: carrier frequency and code chip epoch.

As a hypothetical example, suppose that the carrier frequency uncertainty is such that one hundred different trial carrier frequencies need to be tested in order to assure that the signal will be found sufficiently close to one of the trial frequencies. If the DS code has a period of 127 chips, and the possible code epochs are tested at 1/2 chip intervals, a total of 100 x 127 x 2 = 25,400 trial integrations must be performed in an exhaustive search. After each integration, a test is performed in order to determine whether or not a signal is present. Even if this process requires only 1 millisecond per trial integration, the resulting worst-case acquisition time of 25.4 seconds could be unacceptable for many applications.

Slow FH systems can be easier to synchronize because the SNR for a single hop is typically sufficient to make reliable decisions on the presence or absence of the signal. Since the receiver "knows" the hopping sequence, it can simply observe a single frequency until a signal is detected. If energy is detected on subsequent hops, acquisition is assumed.

2.4.3.2 Matched Filter vs. Correlation Receivers

The process of integrating a number of chips to reconstruct a data bit at a DS receiver can be performed in two distinct ways. The first technique, which was illustrated in Figure 5, consists of multiplying the received signal by a synchronized replica of the spread spectrum phase code. The resulting waveform is integrated in a low-pass filter or in an integrate-and-dump circuit. This implementation is known as a correlation receiver.

In a second approach the spread spectrum phase code coefficients are stored as part of a linear filter, as shown in Figure 6. The two leading technologies for implementing such matched filters or "passive correlators" are charge transfer devices (CTDs) and surface acoustic wave (SAW) devices (12,13,14). Digital implementations have also been demonstrated.

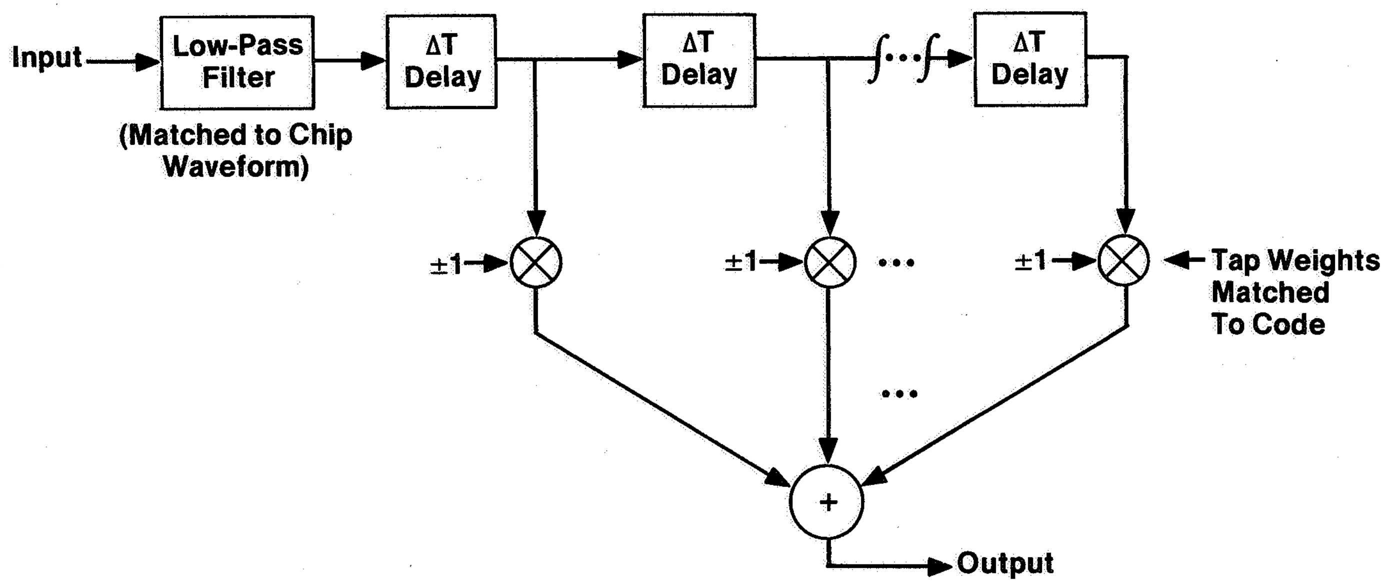

FIGURE 6
TYPICAL MATCHED FILTER (PASSIVE CORRELATOR) FOR DIRECT SEQUENCE SPREAD-SPECTRUM RECEPTION

Matched filter receivers and correlation receivers ideally provide the same SNR performance after synchronization. Matched filter receivers facilitate rapid acquisition and synchronization, but are limited to integration over at most a few thousand chips (15). Correlation receivers allow integration over a virtually unlimited number of chips, but suffer acquisition-time disadvantages, as noted above. Economic factors are also involved in the trade-off between matched filter receivers and correlation receivers.

2.4.4 Spread Spectrum and Error-Correcting Codes

The word "code" as used earlier has applied to the underlying digital structure of a spread spectrum signal. In a separate context, a "code" is used to mean a set of digital elements employed for forward error correction. Since forward error correction is often used in conjunction with spread spectrum systems, the distinction is important.

The relation between spread spectrum and error correction coding was recently explained in a tutorial by Viterbi (16). Error correction uses increased bandwidth to provide improved performance. Unlike "pure" spread spectrum techniques, error correction coding can provide performance improvements in additive white Gaussian noise (i.e., background noise). Error correction coding is beneficial in most, if not all, digital spread spectrum applications, although economic tradeoffs are obviously involved. Multiple access concepts (i.e., CDMA) based on error correction coding have been analyzed (17). In addition, low rate coding (i.e., error correction coding with a large amount of redundancy) has been suggested as a substitute for conventional spread spectrum techniques (87).

3. THE USE OF SPREAD SPECTRUM IN NON-GOVERNMENT SERVICES

3.1 Potential Benefits

Given that spread spectrum techniques have evolved largely in response to military requirements, and in view of the fact that they require large bandwidths (relative to the information bandwidth), it is reasonable to ask why anyone would consider spread spectrum techniques for non-Government* applications. A review of the functional uses of spread spectrum, as presented in Section 2.3.1, provides a background for this issue. Specifically, only the anti-jamming property of spread spectrum seems to be unique to military environments. The other uses, including resistance to unintentional interference, resistance to interception, discrete addressing, multipath resistance, multiple access and pulse compression all have potential civilian applications. Indeed, certain civilian applications have already been exploited; TDRSS and GPS provide well-known examples**, although both will operate under Government authorizations.

In general terms, it is possible to identify four potential motivations for introducing a new communication or radiolocation technology:

- Reduced Cost
- Improved Communication or Radiolocation Performance
- Expanded Capabilities
- Improved Spectrum Utilization

Each of these will now be addressed in the context of spread spectrum techniques.

I. Reduced Cost: Spread spectrum equipment designs require added complexity in both the transmitter and the receiver. Thus, it seems most unlikely that spread spectrum equipment will, in the foreseeable future, be less expensive than comparable narrowband equipment. MITRE has not been able to identify any potential spread spectrum application in which an existing conventional system would be replaced by a spread spectrum system for the purpose of achieving reduced costs. However, a caveat should be added

* The term "non-Government," as used in this report, coincides with the regulatory use of the term. That is, non-Government means non-Federal-Government, but includes state and local governments.

** Although GPS is a DoD program, civilian applications are anticipated.

here. Because of the performance improvements that are possible with spread spectrum, it is conceivable that under certain conditions, a particular spread spectrum system could be less costly -- due to reduced transmitter power or the elimination of ancillary circuits -- than a narrowband system offering the same level of communication or ranging performance. An example of such a tradeoff might be spread spectrum versus polarization diversity for communication in multipath environments. Comparative cost analyses are not generally available in this area.

II. Improved Communication (or Ranging) Performance: As noted earlier, spread spectrum systems can provide significant resistance to unintentional interference and multipath fading. To the extent that error correction coding is used, spread spectrum systems provide improved performance against additive white Gaussian noise. (However, this is also true of conventional systems.) Although conventional diversity techniques also provide fading resistance in communication systems, high-resolution ranging systems frequently depend on wide bandwidths for good performance.

III. Expanded Capabilities: Spread spectrum systems can provide user privacy, discrete addressing, and multiple access on a transmit-at-will basis. All of these features can also be provided without the use of spread spectrum, but performance and cost tradeoffs are usually involved.

IV. Improved Spectrum Utilization: The notion that spread spectrum techniques could provide improved spectrum utilization may be at first surprising. J. P. Costas, in a well-known 1959 article, seems to have been the first to raise this possibility (18). More recently, Cooper and Nettleton have predicted improved spectrum efficiency for high-capacity spread spectrum mobile radio systems (19-24).

It is easy to produce examples in which spread spectrum systems display poor spectrum efficiencies. However, it is also becoming increasingly clear that, depending on the basis of comparison, there are conditions under which the spectrum efficiencies of spread spectrum systems are comparable to or better than the efficiency of conventional narrowband frequency-channelized approaches. In addition, there may be particular bands and/or geographic areas that are not generally suitable for narrowband signals. If spread spectrum signals can be used in such cases, spectrum utilization might be improved, even if the intrinsic spectrum efficiency of the spread spectrum signals is poor.

Despite the potential benefits of applying spread spectrum techniques in non-Government applications, substantial interest has yet to be shown by potential users and manufacturers. Although a few manufacturers have expressed an interest in pursuing specialized markets (25-27, 82), some of the most visible proponents of non-Government spread spectrum applications have been from the academic sector (19-24).

3.2 Spread Spectrum in Dedicated Bands

From the viewpoint of communication performance, the most favorable implementation of spread spectrum would involve the use of dedicated bands for spread spectrum applications. The alternative, as discussed in Section 3.3, is to share or "overlay" the spread spectrum allocation with conventional services. The issue of communication performance for spread spectrum in dedicated bands has been treated in the technical literature (see, for example, 1, 5, 10-12, 15-24, 28, 29). A significant topic that remains to be addressed in this case is efficiency of spectrum utilization. Various aspects of this issue will now be presented. The discussion on efficiency of spectrum utilization will be followed by several descriptions of application concepts.

3.2.1 Efficiency of Spectrum Utilization

Over the years, a number of measures have been applied to the concept of spectrum efficiency (30-38). The usual formulations are cast in terms of a benefit/cost ratio, where the benefits are defined in terms of the number of simultaneous links or users accommodated or the network information throughput; the costs are normally defined in terms of the amount of spectrum resources occupied. Typical examples are number of users per unit bandwidth and number of users per unit bandwidth per unit area (for a given information rate or information bandwidth). Comparisons of various modulation techniques and system concepts may be sensitive to the measure of spectrum efficiency upon which the comparison is based.

One of the most straightforward measures of spectrum efficiency for a network or an associated group of users is total information throughput per unit bandwidth. Under this formulation, the rates of information transfer for all user links are summed and the total is divided by the total bandwidth required to accommodate all the links. All links are assumed to be in continuous use, and no reference to physical area or volume is made. This definition can be used to measure the spectrum efficiencies of satellite links, for example.

On this basis, the spectrum efficiency of spread spectrum signaling can be quite low. Detailed analyses supporting this conclusion are presented in Appendix A and in References 17, 29, and 39-41. Basically, the reason for this result is that under idealized conditions, complete orthogonality between the signals of various user links is more closely approachable with TDMA or FDMA than with SSMA, unless the various SSMA user links can be mutually synchronized. In designing a set of signals for an asynchronous* multiple access application, it is found that the best achievable worst-case cross-correlation between signals degrades (i.e., increases) if each signal is required to be spread uniformly across the available bandwidth (76). If the cross-correlation is minimized, then signals with highly nonuniform spectra result, as in FDMA. More will be said about this in Section 3.3.6.

Example 1

As an example, consider an asynchronous direct sequence SSMA system in which the various user links employ bi-phase keying to embed the information in the spread spectrum signal. All links operate at the same data rate, but each link uses a separate spread spectrum code. The codes are assumed to be quasi-orthogonal. Equal power levels and a non-fading channel are assumed. In addition, the number of active users is assumed to be sufficiently large that, for the purpose of computing the bit error rate, the interference from other links that is experienced by any given receiver can be modeled as a Gaussian random process. Coherent detection (without error-correction coding) is assumed.

Under these conditions, the spectrum efficiency, as a function of the required bit error rate, is shown in Figure 7 (from Appendix A, Figure A3). The lower curve applies when the ratio of received energy per bit to background noise power density (E_b/N_o) is 10. It can be seen that the spectrum efficiency is quite low (less than 0.2) for bit error rates below 0.01. As shown by the upper curve, this conclusion is not substantially changed if background noise is eliminated completely, that is, if $E_b/N_o = \infty$. (The term "background noise", as used above, refers to the noise level in the absence of interference from other spread spectrum links.) If asynchronous frequency hopping with binary frequency shift keying and noncoherent detection are used instead of direct sequence signaling, spectrum efficiency is degraded even further (see Figure A5, Appendix A). The existence of fading can also degrade spectrum efficiency relative to Figure 7. (See Appendix A, Figures A6 and A7.)

* In the context of multiple access, "asynchronous" means that the various links need not be mutually synchronized.

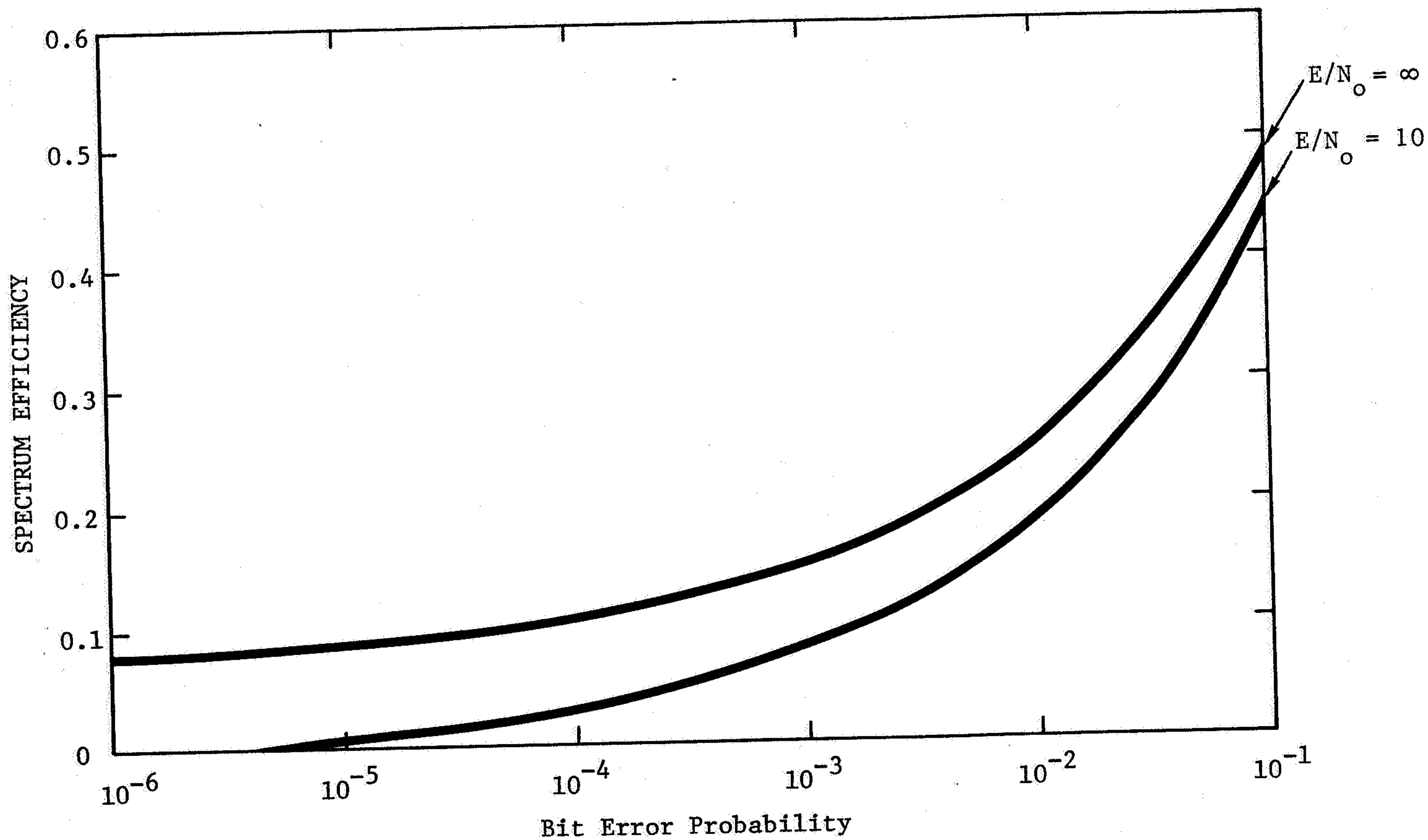

FIGURE 7
SPECTRUM EFFICIENCY OF ASYNCHRONOUS DIRECT SEQUENCE SSMA

As a basis of comparison, an FDMA system requiring 2.5 Hz of r.f. bandwidth for each bit-per-second of transmitted information has an efficiency of 1/2.5 = 0.4 at low error rates, using the definition set forth above. In the absence of frequency errors, this level of performance appears to be attainable with existing narrowband modulation techniques (42, 43) as long as the received power levels in adjacent channels are not greatly different. Thus, at low error rates FDMA can provide at least twice the spectrum efficiency of SSMA for the conditions of this example.

Of course, there is more to the tradeoff between FDMA and SSMA than spectrum efficiency. In addition, there are several factors that can mitigate the relatively negative view of SSMA spectrum efficiency presented above. One of these is the impact of guard bands in FDMA, which are necesary to provide for frequency tolerances, Doppler shifts, modulation sidebands (i.e., "splatter"), finite filter roll-offs, or a combination of these factors. As shown in Appendix B, the existence of frequency tolerances can significantly reduce the spectrum efficiency of FDMA, particularly when the data rate per channel is low and the assigned frequency is high. The following is presented as an example of how this factor can influence the comparison of SSMA and FDMA in terms of spectrum efficiency.

Example 2

The basic assumptions for direct sequence SSMA are as in Example 1, except that a frequency tolerance of 2 parts per million is assumed (see Appendix B, Example 1). In addition, the following parameters apply:

R = information rate = 100Hz

M = Maximum number of users = 100

A_1 = predetection SNR in the absence of interference = 30 dB

A_m = required predetection SNR at full capacity = 10 dB

From Appendix B or reference (44), the required number of code chips per information bit is approximately

$$n = (1/3)(M-1)(1/A_m-1/A_1)^{-1} = 333$$

The required r.f. bandwidth is approximately twice the chip rate, or 2 X 333 x 100 Hz = 66.6 kHz.* In addition, we add an equal bandwidth to provide for guard bands. The total SSMA channel bandwidth is then 133.2 kHz, plus twice the frequency tolerance. (The frequency tolerance is 2×10^{-6} times the center frequency.) At a center frequency of 1.0 GHz, the spectrum efficiency is $(100\text{Hz})(100 \text{ users})/(133.2 \times 10^3 + (10^9)(2)(2 \times 10^{-6})) = 0.073$.

As a comparison, consider an FDMA system having the same number of users, operating at the same data rate, with the same frequency tolerance (i.e., 100 users, 100 Hz per user, and 2 parts per million). Let the bandwidth expansion factor for the FDMA users be 2.0. The total required r.f. signal bandwidth is then $(100\text{Hz})(100 \text{ users})(2.0) = 20$ kHz. If 100 Hz guard bands are added between channels, with 50 Hz guard bands at each of the band edges, the total required bandwidth is approximately 20 kHz + $(100 \text{ Hz})(100 \text{ channels}) + 2(2 \times 10^{-6})(100)f_c = 30 \text{ kHz} + 4 \times 10^{-4}f_c$, where f_c is the center frequency of the band. The spectrum efficiency at $f_c = 1.0$ GHz is $(100 \text{ Hz})(100 \text{ users})/(30 \times 10^3 + (4 \times 10^{-4})(10^9)) = 0.023$, or less than half the efficiency of SSMA. As shown in Appendix B, the break-even frequency at which the spectrum efficiencies of FDMA and direct sequence SSMA are equal (for this example) is 261 MHz. At higher frequencies, SSMA can provide better spectrum efficiency than FDMA.

Similar examples can be produced using frequency-hopping SSMA.

A second factor that alters the trade-off between SSMA and FDMA is the inclusion of geographic coverage in the definition of spectrum efficiency. For example, the spectrum efficiency of a land mobile (voice) communication system might be defined in terms of number of users per unit bandwidth per unit of coverage area. Clearly, if SSMA provides better geographic reuse of the spectrum than FDMA, the trade-off will be shifted in favor of SSMA. The advantage of SSMA in this context is that each user has access to the entire system bandwidth regardless of his geographic location. In conventional frequency-channelized systems, each channel may be available over as little as one-seventh

* In a practical design, the number of code chips per information bit might be increased to the next-highest value of 2^L-1 in order to permit the use of L-bit linear shift register generators. In this case, a 511 chip code would allow the use of 9-bit linear shift-register generators. However, the number of chips per bit is generally not constrained to values of 2^L-1.

of the coverage area*. In small cell systems, the number of user links per channel per cell is typically on the order of 0.045 to 0.055 (see Reference 55). For example, in a 667-channel system, 30 to 37 channels per cell would be available. This apparent disadvantage is overcome by providing a large number of cells in each system.

Unfortunately, a general in-depth analysis of SSMA spectrum efficiency in land mobile radio (LMR) systems has not yet been performed, although some steps in this direction have been taken (19, 29, 45). One of the basic problems is that, particularly in urban areas, there is a great deal of variability in transmission loss over various paths of equal distance for given antenna heights, within a single geographic area (46). There is also variation in path loss characteristics from one city to another (47-49), so the incorporation of simple path loss models (e.g., $1/R^N$ law) in spectrum efficiency calculations is at best an expedient approximation. In addition, the wide variety of possible spread spectrum implementations (including hybrid techniques and code variations) and possible geographic configurations make it unlikely that a general analysis will be performed.

Another dimension affecting spectrum utilization is time. In many services, the average utilization of channels (with respect to time) is low. Nevertheless, the assignment of channels to users on a fixed basis makes it more likely that a particular user will find his channel blocked when it is needed. A spread spectrum multiple access system averts this problem by making a large bandwidth continuously available to each user. Performance degrades gradually as the number of active users increases.

Trunked narrowband systems also provide a solution to this problem, by dynamically assigning channels to users only when the channels are actually needed. Trunked systems assign channels to users on a demand basis until all available channels are in use. At this point, new calls cannot be placed until one or more ongoing calls is terminated. Thus, trunked systems are capable, in principle, of attaining high efficiencies of spectrum utilization. However, trunked systems require a central control of channel assignments, and exhibit a "hard" saturation characteristic when the number of users in the system becomes equal to the number of available channels. It seems likely that there will always be applications in which, for one reason or another, trunking is not practical.

---

* This is based on a cellular approach with one ring of hexagonal cells between base stations using the same frequency.

In summary, the question of efficiency of spectrum utilization, as it applies to spread spectrum systems, should be viewed in the context of the major qualitative differences that exist between conventional frequency channelized or TDMA approaches and random access spread spectrum. Spread spectrum may be an attractive alternative where uncoordinated use of the spectrum over one or more geographic areas is an important requirement. In making comparisons of spectrum efficiency, it becomes clear that the actual capacity of practical FDMA approaches may differ significantly from their theoretical capacities computed on the basis of uniform instantaneous channel loading. The need for geographic frequency reuse intervals and guard bands also limit the spectrum efficiency that can be attained by conventional frequency channelization in practical systems.

3.2.2 Fast Frequency Hopping in High Capacity Land Mobile Radio Systems

3.2.2.1 Differential PSK

In 1977, Cooper and Nettleton (50-52) proposed a unique approach to high-capacity land mobile communications. The following salient features were incorporated in their technique:

- Digitized voice with differential PSK modulation
- Fast frequency-hopping
- A small-cell network implementation
- A unique receiver design based on the use of tapped delay lines and multiple bandpass filters
- Dynamic control of transmitter power levels

The following benefits have been claimed for this approach (23):

1. Resistance to fading is provided by the frequency diversity that is inherent in the fast frequency-hopping technique.

2. Each user has access to the entire system bandwidth. Thus, there are no "blocked calls."

3. There is no hard limit on the maximum number of simultaneously-active users. Communication performance degrades gradually as more and more users enter the system.

4. The system offers privacy from casual listeners.

5. Since it is not necessary for users to switch channels when they cross a boundary between cells, "forced termination" of calls cannot occur.

6. The hardware employed by each (mobile) user is identical, except for the filters associated with that user's unique signal set.

7. Priority messages can be accommodated even when the system is heavily loaded.

8. Under conditions of reduced capacity, the system may be able to co-exist with conventional narrowband services.

The disadvantages of this approach, as recognized by its advocates, include (23):

1. The need for dynamic control of mobile transmitter power levels in order to mitigate the "near-far" problem (i.e., the problem of having a mobile located near the base station interfere with a mobile far from the base station).

2. The requirement for digitized voice, together with the spread spectrum signal format, will increase the complexity of transmitters and receivers.

3. A vehicle location technique of modest accuracy is necessary in order to monitor the vehicles as they move from cell to cell. (This would also be required for conventional small-cell techniques -- see Reference 53).

4. Fully coherent detection is not possible in a rapid-fading LMR environment.

5. The proposed approach is not economically attractive for large-cell systems.

In addition, Mikulsky (54) has pointed out that, because of the receiver implementation technique advocated by Cooper and Nettleton, realization of the "no calls blocked" and "no forced terminations" claims would require that each base station accommodate all possible user codes. Since each code requires a unique delay-line/filter configuration in the receiver, the economic impracticability of this is evident. Thus, blocked and termiated calls would occur as in any cellular system.

In References 19 and 23, Cooper and Nettleton present comparisons of their proposed technique with conventional cellular systems in terms of spectrum efficiency (defined in terms of calls per unit bandwidth per cell). These comparisons show estimated spectrum efficiencies for fast frequency-hopping of 2.0 to 4.7 times the spectrum efficiency of conventional narrowband systems, depending on the specific configurations being compared. These estimates were based on a baseband data rate of 30 kbits/s for digitized voice.

In a subsequent independent analysis, Henry (55) found the DPSK/fast-FH technique to be inferior to conventional narrowband FM techniques on the basis of the average number of usable channels per cell. In his example, narrowband FM was superior by a factor that ranges from 1.2 to 2.8, based on a 32 kbit/s digitized voice rate for the fast frequency hopping system. (If a 30 kbit/s rate is used as in the earlier analysis by Cooper and Nettleton, Henry's spectrum efficiency penalty (for fast FH) is 1.12 to 2.62.)

Further improvements in the spectrum efficiency of this spread spectrum approach can be attained by lowering the digitized voice rate. Digital voice has been used successfully at rates as low as 2.4 kbits/s. However, low data rates usually result in increased equipment cost or degraded voice quality, or both. If degradations in voice quality are allowed, then comparable narrowband systems can also increase spectrum efficiency by using closer channel spacings (with or without corresponding reductions in the r.f. bandwidth). On the other hand, digital voice technology is advancing rapidly, and it is possible that low cost systems for high quality voice transmission at data rates on the order of 10 kbits/s (with an acceptable tolerance for transmission errors) will be available in the future. (See Flanagan, Reference 100, for an overview of digital voice technology.) If so, then the fast FH/DPSK technique will exhibit superior spectrum efficiency even in Henry's example.

It has also been suggested that the spectrum efficiency of the Cooper-Nettleton approach might be further improved by controling the mobile transmitter power to optimize the signal-to-interference ratios at all base stations, rather than trying to maintain a constant mean received signal power at the base station with which the mobile is communicating (89, 90).

3.2.2.2 Multilevel FSK

In a recent anlaysis of a concept originally described by Viterbi (101), Goodman and others have evaluated the performance of a fast frequency hopping multiple access technique that uses multi-level FSK as a basis for signaling (102-105). The approach is quite similar to the one described by Cooper and Nettleton, except for the method in which the transmitted information is embedded in and recovered from FH waveforms.

Figure 8 illustrates the operation of the FH/FSK transmitter. The digital data stream representing the voice input is divided into blocks of k bits (typically, k = 8). Each k-bit block is used to select one of 2^k time-frequency sequences. Each such sequence has a length of L time chips, or hops (typically, L = 19). To provide for discrete addressing, the basic time-frequency sequence used by each transmitter is different. Information is imposed on this basic time-frequency sequence by selecting one of 2^k possible cyclic shifts of the basic sequence in the frequency domain for each k-bit input data block. Thus, 2^k channels are required.

The operation of the receiver is illustrated in Figure 9. The transmitter's address code (i.e., its basic time-frequency sequence) is used to de-hop the intended signal, yielding a matrix of 2^k channels by L time chips. A hard decision is made on the presence or absence of a signal in each of the $Lx2^k$ matrix elements. In the absence of noise, fading, and interference, the de-hopped signal will occupy a single row of the matrix; all of the other 2^k-1 rows will be vacant, indicating the absence of signal energy. With noise, fading, and interference, the row corresponding to the correct signal can contain deletions; other rows can contain insertions. Interference in the row containing the correct signal nominally has no effect. The simplest decoding algorithm involves selecting the row containing the largest number of signal "hits."

For 32 kbit/s digitized speech, k = 8 and L = 19, about 170 users can be supported at an error rate of 10^{-3}. This estimate includes the effects of multipath and noise, as well as mutual interference in an isolated system, but it does not include the effects of urban shadowing, synchronization errors, "splatter" between adjacent channels, or interference from nearby cells in a cellular system. Nevertheless, it appears that the number of users per unit bandwidth under this approach is several times the corresponding number for FH/DPSK, as described in the previous section, and is at least comparable to the number of users per unit bandwidth that can be supported in a conventional narrowband FM system in which frequencies must be re-used over a particular service area.

A major disadvantage of this approach is that it requires that a real-time spectrum analysis be performed in the receiver. For k = 8 and L = 19, and a 32 kbit/s data rate, this analysis must be performed every 13.2 microseconds and it must resolve 2^8 = 256 separate channels over a total bandwidth of about 20 MHz. While this requirement is entirely within the state-of-the-art, the implementation costs could be excessive.

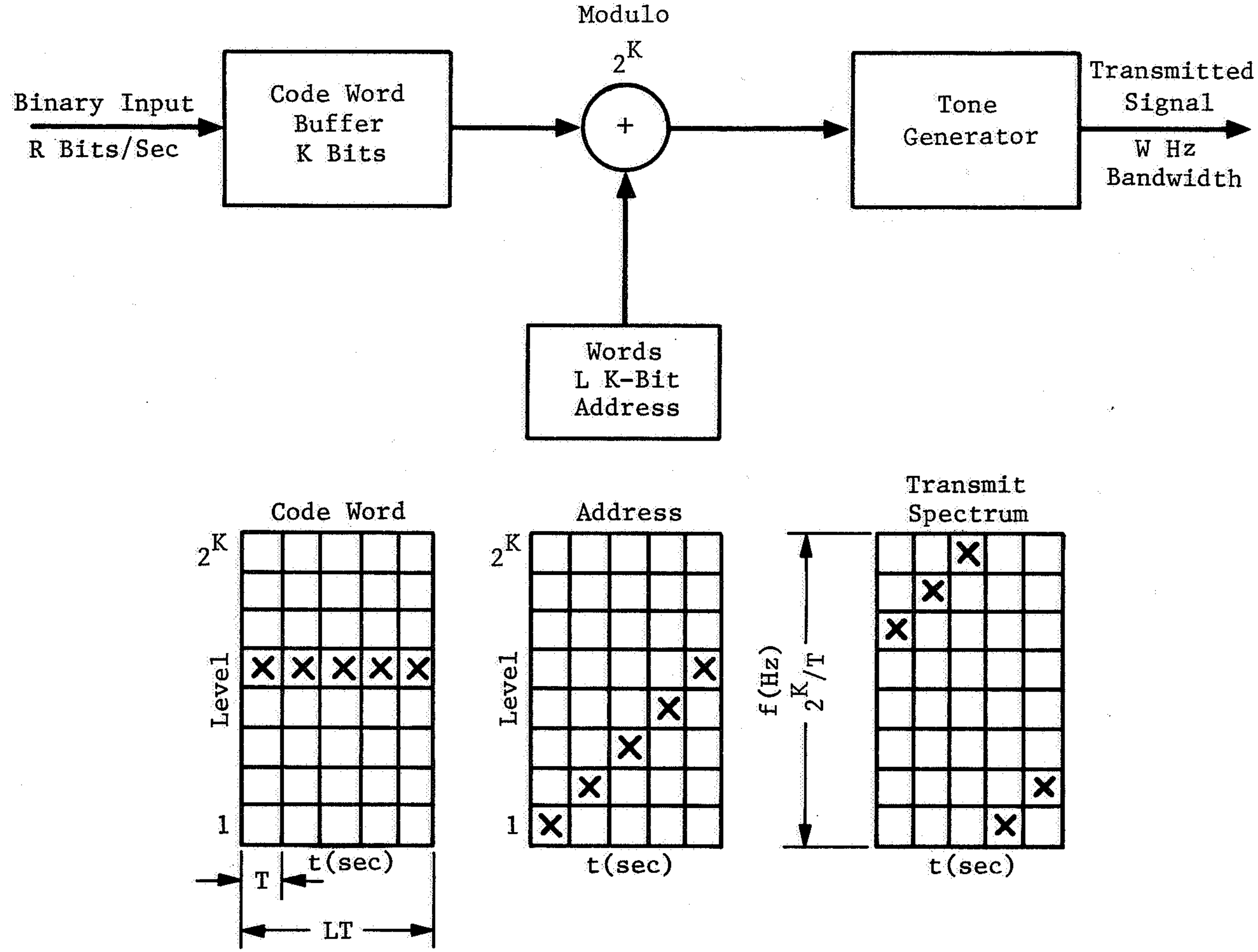

FIGURE 8
FH/FSK TRANSMITTER

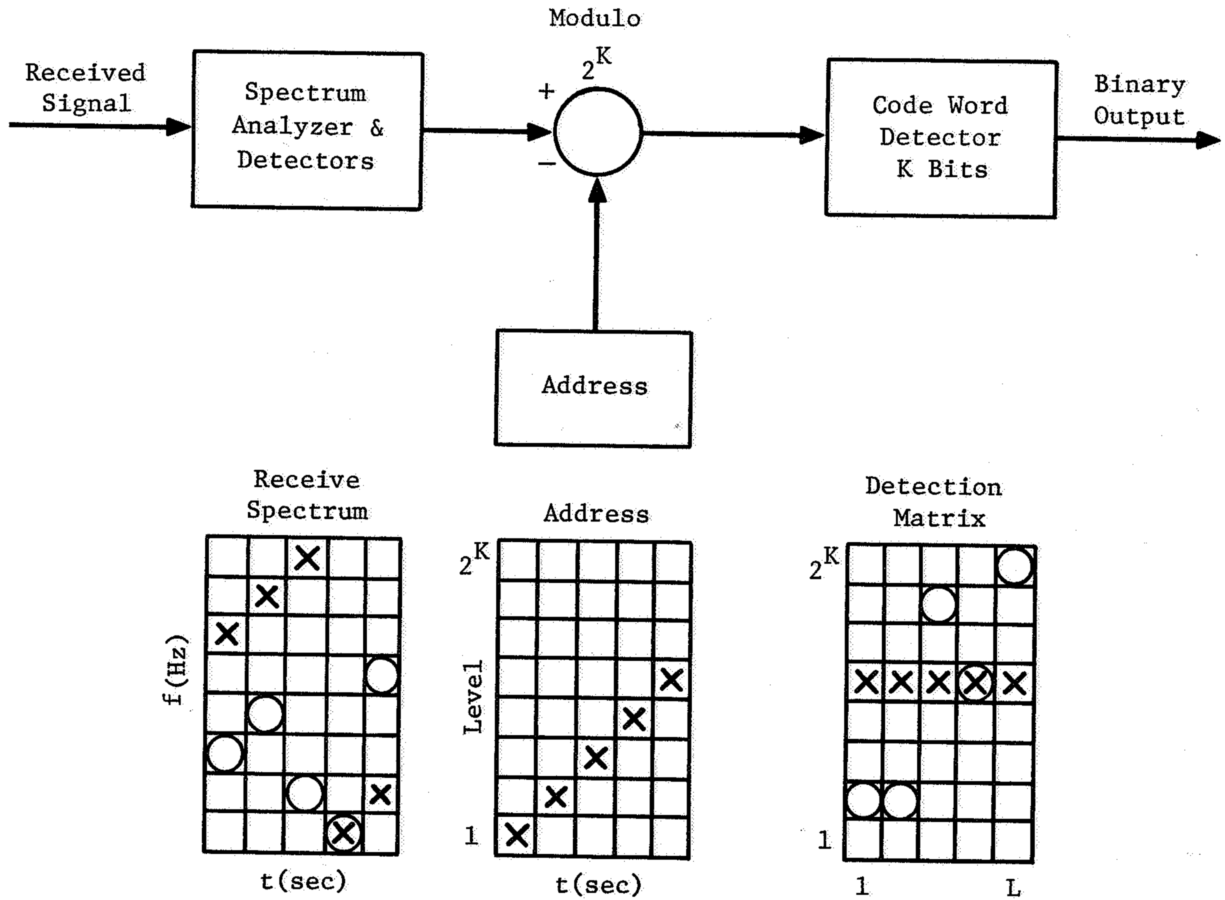

FIGURE 9
FSK/FH RECEIVER

3.2.3 Independent Land Mobile Radio Systems

In a recent study performed by NTIA for the FCC's UHF Task Force, Berry and Haakinson (45) analyzed multiple independent land mobile radio (LMR) systems using spread spectrum in a common band. It was found that the spectrum efficiency of spread sprectrum in this case was low compared with conventional techniques. The analyses apply to direct sequence techniques but do not explicitly treat the case of frequency-hopping.

The study also contains a section on spread spectrum "overlay" with conventional services (see Section 3.3.2 below) as well as tutorials on spread spectrum and land mobile radio.

3.2.4 Spread Spectrum in the Maritime Mobile Service

Experimental use of spread spectrum for maritime mobile applications dates back to 1974, when NASA and the Maritime Administration performed tests via NASA's ATS-5 and ATS-6 satellites to evaluate direct sequence ranging for maritime navigation purposes (88). Later efforts to use spread spectrum signaling for maritime satellite navigation have focused on GPS.

More recently, spread spectrum has been suggested for two other applications in the maritime mobile service. They are: emergency communications via satellite and overlaid communications in the terrestrial VHF marine band. The latter will be discussed in Section 3.3.3. This section will provide brief descriptions of proposed spread spectrum signaling for distress alerting via satellite.

SAMSARS (Satellite Aided Maritime Search and Rescue System) is a recent concept for the application of spread spectrum techniques to maritime distress alerting (56-60).* Under the current version of this concept, a commercial ship in distress would activate a low-power (about 10 watts), battery operated transmitter operating at L-band in a channel reserved for distress alerting. These signals would be received and translated to C-band by a commercial maritime communication satellite. At the satellite ground station, the distress message (which would typically include vessel identification, location coordinates, and a distress code) would be decoded and relayed to a rescue coordination center by conventional communication links. In order to provide multiple access and resistance to unintentional interference, a spread spectrum signal format is used.

---

* This concept was developed and analyzed independently by the MITRE Corporation. Subsequent development and test phases were funded by the Maritime Administration and the Coast Guard.

In this application, conventional frequency channelization is not practical for the following reasons:

1. Since it is, by nature, a one-way signaling application, the user cannot listen for a clear channel before transmitting nor can he be instructed to switch to a clear channel by the ground station.

2. If only 1000 units were built and deployed, the bandwidth required to give a separate frequency channel to each would be about 3.2 MHz (= 2 x 1600 MHz x 10^{-6} x 1000) for a frequency tolerance of 1 part per million. (This bandwidth could be reduced by assigning several transmitters to one channel, at the expense of occasional interference.) Less severe frequency tolerances would result in even larger bandwidths.

SAMSARS would use a 200 kHz-wide dedicated channel to accommodate its 128 kHz chip rate. For reasons of economy, each transmitter would employ the same direct sequence code. Multiple signals are separated by code epoch, random carrier frequency offset (resulting from a 10 ppm tolerance) and transmitter duty cycle. At least 58 simultaneously active transmitters can be accommodated within the field of view of any satellite, with a 95 percent message detection probability and a bit error rate of 10^{-5} (Reference 60).

A similar direct-sequence spread spectrum technique for maritime distress signaling has been proposed by Japan (67). The Japanese system would operate with a 406 MHz up-link.

Although narrowband alternatives to SAMSARS have been proposed (61, 62), they would not use frequency channelization in the usual sense. Instead, they would rely on transmitter duty cycle and random frequency offsets to provide multiple access, with occasional interference.

3.2.5 Spread Spectrum in Commercial Satellite Communications

The use of spread spectrum for multiple access in satellite communication systems provides for privacy and resistance to interference. However, as noted earlier, spread spectrum satellite networks tend to require more power and more bandwidth per user than equivalent FDMA or TDMA networks. These characteristics run directly against the current trend of optimizing the use of spacecraft power and bandwidth in commercial satellite communication systems. For this reason, it seems unlikely that spread spectrum or CDMA techniques will come into widespread use in commercial satellite systems within the foreseeable future.

Within this overall trend, spread spectrum techniques may find specialized applications, such as the emergency alerting system described in Section 3.2.4.

In a 1977 study funded by NASA, various multiple access techniques were evaluated for satellite systems providing service to non-Government mobile users (40). FDMA was selected on the basis of its low implementation cost. The low spectrum efficiency of CDMA in this application was also noted.

3.3 Spread Spectrum "Overlay" with Conventional Services

Despite the ever-growing demand for spectrum resources in most non-Government services, the average utilization of the spectrum, even at peak hours in urban areas, is probably low. A recent FCC report (63) indicates that as few as 48 percent of the 25-470 MHz land-mobile channels monitored in Los Angeles had "very high occupancy" (defined as at least 60 percent peak hour message occupancy). The corresponding rate of "very high occupancy" channels in San Diego was 25 percent. Substantial numbers of channels were found to have zero occupancy. Analogous results were obtained in a 1977 study of VHF Maritime Mobile channel occupancy in the New Orleans area (64).

Such observations, however well-founded, are of little use to someone who is trying to get a new frequency assignment in a congested area (or to someone who is trying to use an occupied channel for which he is already licensed). The user of a busy channel (or set of channels) has only one legitimate way to "average out" the variations in channel occupancy--he must wait for a clear channel. He cannot switch to a temporarily-vacant channel that is allocated to a service in which he is not authorized to operate. Indeed, he will only be aware that the channel or channels on which he is authorized to operate are busy.

If spread spectum signals were "overlaid" on existing allocations, it might be possible to improve the overall efficiency of spectrum utilization, even if the spectrum efficiency of the individual spread spectrum signals were quite poor. This prospect has given rise to a number of studies, which will be summarized below. First, however, a brief review of the basic technical issues will be presented.

3.3.1 Basic Technical Issues

In considering spread spectrum band-sharing with conventional services, three distinct interference modes must be considered:

1. Interference to the conventional services from spread spectrum signals,

2. Interference to spread spectrum signals from conventional services, and,

3. Interference to spread spectrum signals from other spread spectrum signals.

The third interference mode--multiple access interference to spread spectrum signals has been treated in the technical literature (6, 11, 17-24, 28, 29, 40, 44, 45, 50-52) and in Appendix A. In general, meaningful analyses of spread spectrum self-interference require at least partial definitions of the application and the specific spread spectrum techniques to be used.

Spread spectrum interference to conventional receivers can take two forms. In the most serious form, the spread spectrum signal produces interference when the intended conventional signal is present at the receiver input. The second form occurs when the intended conventional signal is absent, but the spread spectrum signal or signals become noticeable above the level of the background noise. In either form, the effective interference is normally reduced relative to the interference that would have resulted from a co-channel narrowband signal. The amount of interference reduction depends on the specific spread spectrum technique and the type of conventional receiver to which the spread spectrum signal is applied.

As an example, consider a direct sequence signal that has an approximately uniform power spectrum over a bandwidth of 1.0 MHz. If the interference at the input of the conventional receiver is not so great as to cause nonlinear operation of a receiver stage "upstream" from the i.f. filter, and if the i.f. filter has a bandwidth of 14 kHz, for example, then the effective interference power is reduced by approximately 10 log (14 kHz/1000 kHz) = -18.5 dB, due to the "mismatch" between the i.f. bandwidth and the spread spectrum bandwidth.

A frequency-hopping signal can produce short bursts of relatively unattenuated interference in a narrowband receiver. A compendium of methods for analyzing spread spectrum interference to conventional receivers is available in Reference 65.

The recoverability of spread spectrum signals in a band occupied by conventional signals depends on the number, power, and frequency spacing of the conventional signals and on the particular spread spectrum implementation. In direct sequence systems, the interfering (conventional) signals are attenuated by an amount equal to the receiver's processing gain (approximately the ratio of the r.f. bandwidth to the information bandwidth). Slow frequency-hopping systems can be more severely degraded by a number of interferers at widely separated frequencies than by a single high-power narrowband interferer (66). As in conventional systems, interference to a spread spectrum receiver can degrade the intended signal or it can create "nuisance" effects when the intended signal is not present.

In bands that have been channelized at intervals of 25 to 50 kHz, a substantial fraction of the spectrum is committed to guard bands. In these cases, it may be possible to insert frequency-hopping signals between the existing channel assignments. Channels that are especially vulnerable to interference (public safety, for example) could be "notched out" of the FH spectrum by simple logic changes. (See Section 3.3.5.)

3.3.2 Overlay for Land Mobile Service

The prospect for overlaying spread spectrum signals in a common band with conventional narrowband land mobile communications was discussed by Cooper (68) and by Berry and Haakinson (45) in 1978. Both studies presented encouraging preliminary results, although neither explicitly addressed the "near-far" problem. Later, more detailed analyses by Dvorak (69) and Juroshek (70) yielded substantially more negative results. Specifically, Dvorak concluded that the overlay strategy would be practical only for small cell spread spectrum networks. In examples, he demonstrated that, for the 150 MHz land mobile band:

1. Mobile-to-mobile (SS to FM) interference could be unacceptable if the distance between mobiles were less than 24 meters.

2. Mobile-to-base interference (from a single spread spectrum transmitter to an FM receiver) could cause a 3 dB degradation in receiver noise at distances of 15-20 km.

(Part of this analysis was implicitly restricted to direct sequence or fast frequency-hopping signals.)

Juroshek does not treat the small-cell case, but assumes that the spread spectrum transmitters and the FM transmitters have comparable power levels. His analysis is applicable to fast frequency-hopping and direct sequence techniques. The major conclusion of his study is that, over the 150-900 MHz frequency range, spread spectrum and conventional FM land mobile systems cannot share a common band without significant interference to both types of systems. Again, the required separation distances (between the interfering transmitter and the receiver) were found to be typically on the order of 10 km or more.

3.3.3 Overlay for Maritime Mobile Service

In a study funded by the Maritime Administration, NTIA is assessing the potential for overlaying spread spectrum signals on the VHF maritime mobile band (156-162 MHz). An analysis was performed, with emphasis on direct sequence techniques. The results were applied in a hypothetical case study for the New Orleans area, which is known to suffer from radio frequency congestion in the VHF maritime mobile band. Although the results of this analysis have not been published at the date of this writing, a number of tentative conclusions have evolved. These are summarized as follows:

1. The "near-far" problem for the direct sequence spread spectrum, even without narrowband interference, requires an equalization of received signal powers if two or more mobile stations transmit at once to a single base station. This could be done, in principle, by restricting transmitter/receiver distances or by controlling the transmitter power levels. It may be possible to use a direct sequence spread spectrum system to simultaneously communicate with a number of ships in the New Orleans area with acceptably low mutual interference between the spread spectrum signals if a single base station can be located 30 km or more north or south of New Orleans to provide nearly equal mobile-to-base distances, thereby insuring nearly equal received spread spectrum signal powers at the base station.

2. A practical spread spectrum system requires one frequency band for base-to-mobile communications and a second non-overlaping frequency band for mobile-to-base communications.

3. When a direct sequence spread spectrum system with a single base station and multiple mobile stations operate in the same frequency band and geographic area with one or more coventional narrowband FM land mobile systems, mutual interference severely reduces the ranges of all systems.

4. The use of a small cell configuration for the spread spectrum stations can reduce the total spread spectrum interference to FM systems sharing the same frequency band and geographic area. However, this approach will not reduce the FM interference to the spread spectrum system.

In order to verify the preliminary conclusions, tests were performed in the New Orleans area. To avert the necessity of actually radiating a spread spectrum signal, conventional signals in the 156-162 MHz band were picked up by a test antenna, downconverted to an intermediate frequency, amplified, and combined with a direct sequence spread spectrum signal. The composite waveform was used as the input to a direct sequence demodulator.

Although the results of these tests have not yet been published, it is understood that they are consistent with the conclusions of the earlier NTIA analysis. Again, the negative results apply only to direct-sequence systems.

3.3.4 Overlay with Television

More than 400 MHz of bandwidth has been allocated to broadcast television in the United States. Of this, only a fraction of the potential channel assignments are actually used in any given city. This condition has already resulted in the allocation of 470-512 MHz band (channels 14-20) on a shared basis between television broadcast service and land mobile service. It is logical to ask whether the television bands might be shared with one or more spread spectrum allocations.

Preliminary studies on band-sharing between spread spectrum and television signals were published in 1978 by Juroshek (71) and Ormondroyd (72). Both studies included laboratory experiments involving the use of standard TV receivers and spread spectrum signals. Using a black-and-white TV receiver, Ormondroyd found that spread spectrum signals generated by modulating an AM signal with a direct sequence pseudonoise code caused much less degradation than conventional narrowband AM signals. On the other hand, Juroshek found that direct sequence spread spectrum signals produce about the same amount of interference to color TV as narrowband FM signals of the same level, as long as the spread spectrum bandwidth is about 2.0 MHz or less. For spread spectrum bandwidths greater than 6 MHz, spread spectrum should cause reduced interference relative to narrowband FM signals, simply because of the TV receiver's ability to attenuate interference outside of its nominal 6 MHz passband.

In 1979, Coll and Zervos proposed the use of spread spectrum techniques for multiplexing digital data with video in television-based information distribution systems (73). A laboratory experiment was performed in which the direct sequence chip rate was 63 times the horizontal scan rate (63 x 15,734 Hz = 991.25 kHz). A spread spectrum level 40 dB below the peak-to-peak video amplitude was used. Reliable operation of both the television and the data signal was reported.

3.3.5 JTIDS - A Case Study in Band-Sharing with Spread Spectrum

The U. S. table of frequency allocations was recently amended to permit the operation of JTIDS* in the 960-1215 MHz band. Although JTIDS is a Government system, the process by which its allocation was obtained, as well as the band-sharing characteristics of the system, are of interest because they demonstrate the use of a common frequency band by several independent wideband systems.

The design of JTIDS led to the early definition of a requirement for 150 MHz of bandwidth and in the range of 200-2000 MHz(3), based on technical and operational considerations. It was clear that obtaining this much bandwidth below 2 GHz on a dedicated basis was out of the question.

JTIDS shares the 960-1215 MHz aeronautical radionavigation band with TACAN/DME and the Air Traffic Control Radar Beacon System (ATCRBS), both of which use pulse signals. JTIDS uses a frequency-hopped signal format that is designed to minimize emissions in the 1030 MHz and 1090 MHz ATCRBS bands. An illustration of the composite JTIDS spectrum is shown in Figure 10. The 1030/1090 MHz "notches" in the spectrum are achieved by the elimination of the corresponding frequency slots from the hopping sequences, in conjunction with controlled pulse rise and fall times, continuous phase shift modulation, and sidelobe filtering of the modulating waveform (3).

JTIDS was found to be compatible with all present and planned uses of the 960-1215 MHz band, including ATCRBS monopulse, Discrete Address Beacon System (DABS), Beacon Collision Avoidance System (BCAS) and the DME portion of the Microwave Landing System (MLS/DME), as well as TACAN/DME and ATCRBS. However, the frequency coordination process, if it can be called that, was unquestionably one of the most expensive ever undertaken. Assuring compatibility between JTIDS and all present and planned uses of the 960-1215 MHz band entailed approximately 1500 hours of bench tests, 1000 hours of flight tests and eight man-years of analysis (3).

The resulting data were compiled in a 2000 page report (74). The magnitude of this effort was related partly to the fact that services protecting the safety of human life were involved, and partly because JTIDS is one of the first spread spectrum systems to share a band with such services.

---

* Joint Tactical Information Distribution System -- See Section 2.3.2.

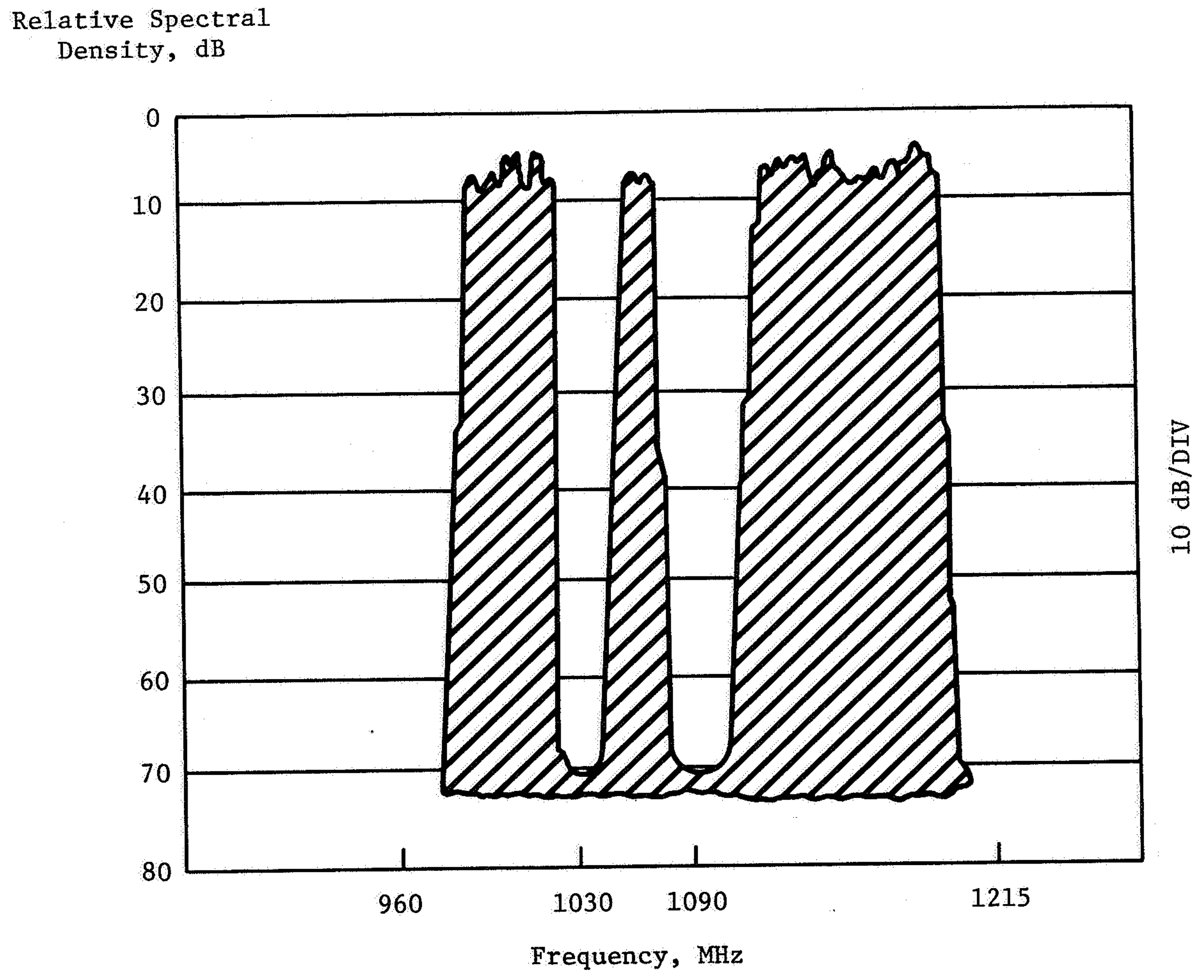

FIGURE 10
COMPOSITE JTIDS SPECTRUM

3.3.6 Theoretical Considerations

In designing a SSMA system to operate in a band with narrowband signals, it is desirable to provide a set of spread spectrum user codes with the following properties:

1. The interference between spread spectrum users should be minimal.

2. The interference produced by a narrowband signal at the worst-case frequency should be minimal.

3. The interference suffered by a narrowband receiver at the worst-case frequency should be minimal.

These goals place conflicting requirements on the spread spectrum code set. A practical design thus entails a tradeoff. Consider first a direct sequence system.

Property 1 implies a uniformly low cross-correlation between different user codes. Properties 2 and 3 imply a "maximally flat" power spectrum for each code, which, for rectangular chips, means, in effect, a $(\sin kf)^2/(kf)^2$ power spectrum of the type produced when the periodic auto-correlation function has no sidelobes.

That a tradeoff is necessary in direct sequence systems is implied by the Welch bound (75). In any phase-coded signal set, the largest of the cross-correlations or auto-correlation sidelobes has a minimum attainable value. Specifically, if C_1 is the largest cross-correlation in a signal set and C_2 is the largest auto-correlation sidelobe in the signal set, then

$$C^2_{max} \geq f(M,L)$$

where $C^2_{max} = \text{Max}\ (C_1^2,\ C_2^2)$

M = number of codes in the code set

L = number of elements (chips) in each code

The particular form of f(M,L) depends on whether periodic or aperiodic correlations are being considered. Both forms result in a positive value of f(M,L) for values of M greater than 1. If uniform spectra are required, then $C_2=0$ and $C_1^2 \geq f(M,L)$. If low cross-correlations are required, then $C_1^2 < C_2^2$ and $C_2^2 > f(M,L)$. Thus, it is impossible to achieve uniform spectra ($C_2 = 0$) and arbitrarily low cross-correlations.

A second derivation of the tradeoff between low auto-correlation sidelobes and low cross-correlation in a signal set was developed by Sarwate (76). This bound has the form

$$C_1^2 + g(M,L) > C_2^2 L$$

Sarwate also shows that in any set of L signals, each containing M = L elements, a requirement for zero cross-correlations implies that the magnitude of the periodic auto-correlation takes on its maximum value for every possible shift of every signal. That is, $C_1 = 0$ implies $R_k(j) = R_k(0) = L$, for all j and all k. (Here, $R_k(j)$ is the periodic auto-correlation function of the k^{th} sequence.) What can be said about the spectra of such signals is that they are highly non-uniform, since uniform spectra imply $R(j) = 0$ for $j \neq 0$.

Although such bounds have not been explicitly formulated for frequency-hopping systems, it seems likely that a similar tradeoff exists. For example, the so-called "one-coincidence" codes* (77) provide low mutual interference between FH users while producing a relatively uniform power spectrum. However, the number of such codes is only equal to the number of FH channels (i.e., frequency slots). In order to increase the number of user codes, the orignal one-coincidence code set can be modified in such a way that low mutual interference between FH users is preserved, but the individual modified codes no longer use all available frequency slots (23). Interference to and from conventional narrowband signals at the worst-case frequency is therefore increased. Alternately, the number of user codes could be increased while using every frequency slot once (or N times) in every code, but mutual interference between FH users would be increased with respect to the one-coincidence codes.

3.4 Spread Spectrum in the ISM Bands

Frequencies allocated for Industrial, Scientific, and Medical (ISM) purposes provide for the unlicensed operation of devices which use radio waves for the purposes other than communication. Such devices include, for example, medical diathermy equipment, industrial heating equipment, and microwave ovens. Because of the inherent interference resistance of spread spectrum receivers, it is logical to examine the possibility of operating spread spectrum systems in these bands.

* So-called because any two codes or their time shifted replicas occupy the same frequency slot during no more than one hop per code cycle.

3.4.1 Current Rules and Practices

There are currently seven ISM bands below 25 GHz, as shown in Table 1. These allocations encompass a total of more than 526 MHz. They range in width from 13.56 kHz (at 13.56 MHz) to 250 MHz (at 24.125 GHz). Besides accommodating ISM equipment operated under Part 18 of the FCC Rules and Regulations, these bands provide, on a shared basis, communication and radiolocation functions as shown in the last column of Table 1. In addition, field disturbance sensors (e.g., intrusion alarms and theft detectors) and low power communication devices are operated in certain ISM bands under Part 15 of the Rules and Regulations. Note also that most bands are shared with Federal Government stations or are allocated to such stations except through coordination with the FCC.

In general, radiocommunication services and radio frequency devices operated under Part 15 are required to accept interference from ISM devices. In addition, radiocommunication services in the 890-902 MHz and 928-940 MHz bands, which are adjacent to the 902-928 MHz band, are required to accept ISM interference from devices which met existing FCC standards on their date of manufacture. The only operation within the ISM bands that is protected from interference is Airport Surface Detection Equipment (ASDE) operating in the 24.-24.25 GHz band.

3.4.2 Characteristics with Respect to Potential Spread Spectrum Use

The only ISM band below 900 MHz having a significant bandwidth (if 320 kHz can be called a significant bandwidth) is the 26.96-27.28 MHz band. However, this band is already extensively in use as citizens band channels 1-27. As of February 1980, there were nearly fifteen million licensed citizens band stations, authorized to operate the range of 26.965-27.405 MHz. In addition, the upper 50 kHz of the 27.12 MHz ISM band is authorized for use by other services, as indicated in Table 1. A small number of Governemnt stations also operate in this band. The 13.56 MHz and 40.68 MHz bands may find limited spread spectrum applications in special cases where very low information rates are used.

The 915 MHz ISM band encompasses a total bandwidth of 26 MHz that appears to have few non-ISM uses, aside from Government and experimental stations. Although the 902-912 MHz and 918-928 MHz portions of this band are available to automatic vehicle monitoring systems (on the basis of non-interference to Government stations), only one such system has been operationally deployed in the band. A search of the FCC frequency listing for this

TABLE 1

ISM BAND ALLOCATIONS AND USES

CENTER FREQUENCY, MHz	TOLERANCE	DESIG-NATION*	TYPICAL ISM USES (78)	TYPICAL NON-ISM USES
13.56	±6.78 kHz			• Aeronautical Fixed Service (where landline comm. is not available at en-route stations)
27.120	±160 kHz		• Commercial Food Processing • Medical Diathermy • Land Transportation Radio Serv. • Low Power Comm. (under Part 15)	• Class C and D Citizens Band • Public Safety Radio Serv. • Industrial Radio Serv.
40.680	±20 kHz	G	• Commercial Food Processing	• Telemetry from Ocean Buoys and Wildlife (Secondary)
915	±13.0 MHz	G	• Commercial Food Processing • Medical Diathermy • Home Microwave Ovens • Tracking, Telemetry, Control and Comm. (Government)	• Automatic Vehicle Monitoring Systems (Secondary) • Field Disturbance Sensors (Intrusion & Theft Alarms)
2450	±50.0 MHz	G, NG	• Commercial Food Processing • Medical Diathermy • Medical Research • Plasma Research • Microwave Power Transmission Experiments • Home Microwave Ovens	• Amateur • Fixed • Mobile • Radiolocation • Field Disturbance Sensors
5800	±75 MHz	G, NG	• Radiolocation • Amateur • Field Disturbance Sensors	
24,125	±125 MHz	G, NG	• Radiolocation • Amateur • Airport Surface Detection Equipment (Primary) • Field Disturbance Sensors	

* G = (Federal) Government
NG = Non (Federal) Government

band produced only 35 assignments, other than experimental, ISM*, and Government listings, as of September 2, 1979. A total of 112 unclassified Government listings (many of which include multiple stations) were on file with IRAC (Interdepartment Radio Advisory Committee) as of February 27, 1980. A small number of classified Government listings were also on record.

The increased path loss and slightly higher equipment costs for the 915 MHz ISM band (relative to lower bands) may be of concern to potential users. However, the increased path loss (for transmit and receive antennas of fixed gain) is primarily relevant to attaining given levels of performance in the presence of natural background noise and receiver noise. For interference-limited systems using omnidirectional antennas, performance is relatively independent of frequency, since the interferer-to-receiver links will exhibit the same frequency dependence as the intended transmitter-to-receiver link. Indeed, frequencies in the vicinity of 1 GHz may be favorable for such systems, because the external background noise is low.

Although equipment costs for the 915 MHz band may be somewhat higher than for lower bands, commercial communication equipment for the 800 MHz land mobile bands has been available for some time.

One of the current uses of the 915 MHz ISM band is for home microwave ovens, although most models operate in the 2450 MHz ISM band. Microwave oven emissions typically occupy several megahertz of bandwidth. Emission levels up to 81 dB above one microvolt per meter have been observed in a 30 kHz bandwidth at a distance of 1000 feet from a single oven, with no intervening obstacles between the oven and the receiving antenna (79). Since most emissions occupy less than 10 MHz, an upper bound on the total field strength can be placed at about 81 + 10 log (10 MHz/30 kHz) = 106 dB above one microvolt per meter at 1000 feet.

The observed interference levels were several dB lower when outdoor measurements of emissions from home-installed ovens were made (79). Building penetration loss in the vicinity of 900 MHz is on the order of 10 to 25 dB (80).

The 2450 MHz ISM band (2400-2500 MHz) is extensively used for home microwave ovens and for commercial, medical, and research purposes. In addition, this band is occasionally used for non-ISM purposes by such diverse groups as industry, local govern-

* ISM equipment must be licensed if and only if it does not comply with the standards set forth in Part 18 of the Rules and Regulations.

ments, fire, police, railroads, and highway maintenance services. The FCC Master Frequency List contained 319 non-Government assignments (other than experimental and ISM listings) in the 2450 MHz ISM band as of September 2, 1979. Only nine unclassified Government assignments were recorded with IRAC as of February 27, 1980.

The 2450 MHz ISM band has also been proposed for the transfer of microwave power from a solar power satellite to earth (114). If implemented, the total radiated power from such a system would be on the order of several Gigawatts (cw) in a beam having a diameter of several kilometers at the surface of the earth. The beam sidelobes, combined with scattering in the ionosphere, could make this band unusable over wide areas.

The principal use of the 5800 MHz ISM band (5725-5825 MHz) is for Government-operated radar systems. As of February 27, 1980, there were 121 Government assignments in this band, virtually all of them for radar or radar-related systems. Only four non-Government listings were on file with the FCC as of September 2, 1979, exclusive of experimental and ISM uses.

The 24,125 MHz ISM band (24,000-24,250 MHz) is near the 22,235 MHz water-vapor absorption line and is, for that reason, better suited to short and medium range applications than to long range applications. Total atmospheric attenuation, under good meteorological conditions, is typically 0.05 to 0.2 dB per kilometer at the surface of the earth (81). The principal use of this band is for police radar systems. Over 1600 licenses had been granted for non-Government use of the band (exclusive of experimental and ISM licenses) as of Septemebr 2, 1979. There were only 31 IRAC listings for government stations as of February 27, 1980. Most of these are for police radar.

3.5 Alternatives to Spread Spectrum

In order to achieve a complete view of the spread spectrum issue, it is necessary to recognize that in some cases, there are narrowband alternatives to spread spectrum that can provide most of the benefits of spread spectrum techniques. It would be impossible to list all of the potential alternatives, since this would require a study of all potential applications, present and future. However, a few examples will now be given. The examples are ogranized according to the type of benefit that the alternative technique is intended to achieve.

Privacy. Spread spectrum techniques provide a measure of user privacy that falls between encryptation and completely uncoded transmissions. There are a number of alternatives.

The simplest is to substitute use of the public telephone system for radio communications whenever possible. Although this approach is not entirely "secure" in an absolute sense, it provides more privacy than uncoded private land mobile radio. Radio paging is available in most metropolitan areas, and this service can be used to establish telephone communications when a call originating from a fixed location is placed to a mobile user. Since the paging signals are coded, they are not readily intercepted by unsophisticated listeners.

The use of improved mobile telephone service (IMTS) or advanced mobile phone service (AMPS) also provides some privacy. Although the transmissions are uncoded, a listener trying to intercept a call from a particular talker must scan all the IMTS or AMPS channels. The degree of protection is somewhat higher in small-cell systems, because a mobile unit changes channels frequently as it passes from cell to cell.

Finally, it has been possible for some time to purchase voice communication security units ("scramblers") for use with conventional mobile transceivers (83).

Discrete Addressing. The discrete addressing feature of spread spectrum has two benefits. The first of these is privacy, which was just reviewed. The second is selective calling. It is important in many applications to be able to automatically route calls to particular stations. This feature is included in all IMTS and AMPS designs, as well as earlier mobile telephone systems and paging systems. Selective calling equipment is available from the major manufacturers of private land mobile radio equipment. Selective calling is also available for certain conventional maritime mobile equipment (84).

Resistance to Multipath Propagation. A primary manifestation of multipath propagation in narrowband communication systems is fading. As noted in Section 2.3.1, certain spread spectrum implementations (as well as simple, wideband pulse techniques) can substantially mitigate fading. However, diversity techniques can be used with narrowband signaling to reduce fading (85). Diversity techniques have been only partially successful in mitigating the impact of multipath errors in narrowband ranging and radiolocation systems (86).

Spectrum Efficiency. Examples were given in Section 3.2.1 and References 19 and 23 of cases in which spread spectrum implementations use the spectrum more efficiently than conventional narrowband implementations. However, the outcomes of such examples obviously depend on the particular narrowband implementation with which a spread spectrum approach is compared. The spectrum efficiency of some narrowband systems could potentially be improved by techniques such as trunking, channel-splitting, or the use of single sideband.

In low data rate microwave FDMA systems where frequency tolerances limit spectrum efficiency, a common frequency reference, coupled with a frequency distribution system, may be appropriate. In other applications, the use of ovenized oscillators (instead of temperature-compensated crystal oscillators) may be justified. However, these alternatives are obviously not viable when the frequency tolerances are due primarily to Doppler shifts.

Resistance to Interference. In some cases, spread spectrum techniques may offer the most economical approach to mitigating interference. In other cases, narrowband techniques, combined with the use of directional antennas, special filters or interference blanking may be preferable. In advanced applications, the use of adaptive array antennas may be justified, with or without spread spectrum. It is also possible to envision "smart" narrowband feedback communication systems that would automatically detect the presence of narrowband interference, locate a clear channel, and switch to the new channel, all without human intervention.

Clearly, the choice between the use of spread spectrum techniques and narrowband techniques (where such a choice exists) requires evaluation on a case-by-case basis. The tradeoff may affect cost, performance, spectrum efficiency, and the availability of operational features.

4. EXAMPLES OF HYPOTHETICAL IMPLEMENTATIONS

Up to this point, a few Government applications of spread spectrum have been briefly described (Section 2.3.1), and reference has been made to several non-Government applications that have been proposed or analyzed: fast FH/DPSK and FH/FSK for high capacity land mobile service (Section 3.2.2) and SAMSARS, which uses direct sequence signaling for maritime distress alerting via satellite (Section 3.2.4). Naturally, it is impossible to predict what new non-Government applications, if any, will be proposed in the future. However, in order to provide a better grasp of the variety of potential spread spectrum implementations, two hypothetical examples of such implementations are presented in this chapter. These examples are not intended as recommendations, but only as illustrations of potential spread spectrum applications for non-Government services. In selecting these examples, particular emphasis has been placed on approaches that have simple implementations. It is recognized at the outset that the price of simplicity is often reduced performance. Nevertheless, there is a limited payoff in considering potential spread spectrum systems having such complex implementations that the prospects for their practical realization in the foreseeable future are small.

4.1 Slow Frequency Hopping with FM Voice

4.1.1 Objectives

This approach would provide privacy, multiple access without coordination, discrete addressing, and resistance to narrowband interference. It would not provide any resistance to wideband interference.

4.1.2 Potential Applications

In a general context, the applications considered here are mobile applications where the number of potential users in a given area significantly exceeds the number of available narrowband channels. For the purpose of illustration, a UHF citizens band application has evolved as a principal example. Other mobile applications may be equally appropriate.

4.1.3 Basic Concept

Most late-model citizens band transceivers are based on the use of low-cost phase-lock loop frequency synthesizers for channel selection. Low-cost frequency synthesizers are also used extensively for VHF maritime mobile and aeronautical mobile transceivers, for IMTS transceivers, and for other communication equipment. The concept presented here is based on the premise that such transceivers could be economically modified for slow frequency hopping.

As in conventional signaling, mobiles and base stations might transmit on the same set of frequencies (as in the present 27 MHz citizens band) or in separate sub-bands separated by typically three to five megahertz (as in many land mobile systems). The former option generally results in the least expensive equipment configuration, particularly if direct mobile-to-mobile operation (without repeaters) is required. The latter provides for the possibilities of full duplex and repeater operations.

In a practical implementation, transceivers could provide the option, by switch selection, of operating in a conventional mode or in a frequency-hopped mode. This possibility opens up a second set of system design alternatives, as follows:

1. Provide separate sub-bands for conventional channels and frequency-hopped channels.

2. Allow the same set of channels to be used for conventional communications and frequency-hopping.

3. Allow most channels to be used for either frequency-hopping or conventional communications, but reserve some channels for FH acquisition sequences.

4. Reserve a separate set of channels for conventional communications only, as an adjunct to alternative 2 or 3.

Since this is only a hypothetical example, no attempt is made to analyze the advantages and disadvantages of the various system alternatives. Instead, it has been assumed that base and mobile transmissions occupy separate sub-bands and that conventional and FH transmissions are similarly separated. A hypothetical channelization plan reflecting this strategy is shown in Figure 11a. Although an allocation in the 912-918 MHz band is shown, the basic approach is applicable, in principle, to any available VHF or UHF band. Figure 11b shows an alternate frequency plan in which only the FH channels use separate bands for base and mobile frequencies.

Note that mobile-to-mobile transmissions can occur in two modes: by repeater, or by providing mobile receivers that can be switched to either the uplink or the downlink band.

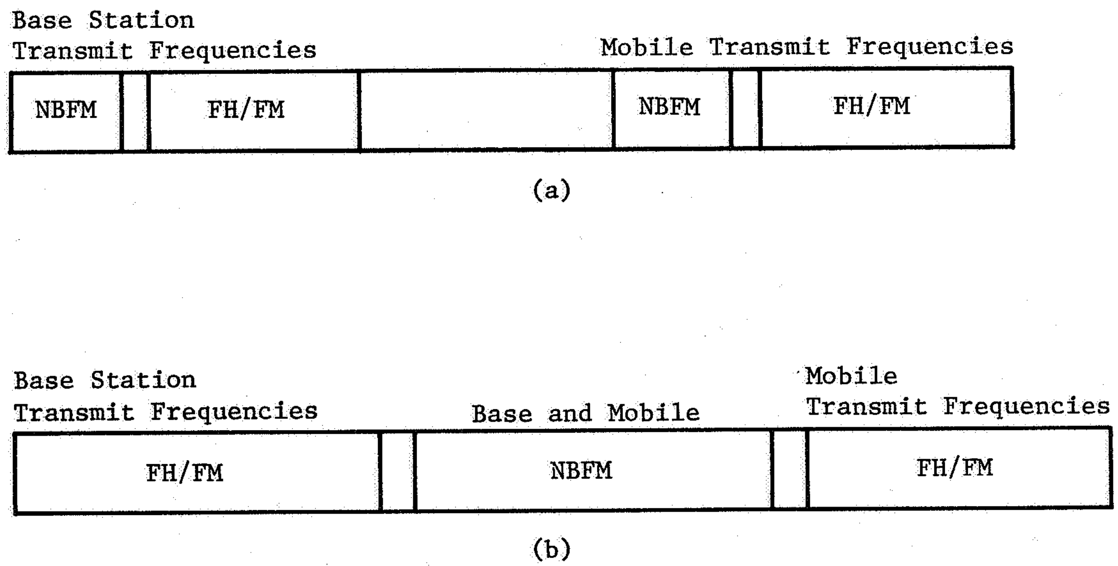

FIGURE 11
HYPOTHETICAL FREQUENCY PLANS FOR FREQUENCY HOPPING MOBILE COMMUNICATION

4.1.4 Equipment Configuration

Figure 12 shows a simplified block diagram of a conventional synthesizer-driven transceiver. This configuration differs from older designs only in that the carrier and local oscillator (L.O.) frequencies, instead of being generated by a separate crystal for each channel, are produced in a low-cost synthesizer using one or two crystals. In some cases, the synthesizer consists of as few as two large scale integrated (LSI) circuits, in addition to a few dozen passive elements. A programmable read only memory (PROM) is sometimes used to restrict the available channels to those for which the unit is licensed*. In transceivers designed for CB or marine use, this function can be accomplished in a permanent fashion by the use of a read only memory (ROM) or by decoding logic between the channel selector and the synthesizer.

Figure 13 shows how a conventional synthesizer-driven transceiver design might be modified for frequency-hopping. It can be seen that this design differs from the conventional design only in the addition of a "Frequency Controller" function and a PROM containing the hopping sequence. The frequency controller provides for switching the synthesizer output frequencies according to a PROM-selected FH pattern. A simplified diagram of the frequency controller is shown in Figure 14. The PROM containing the hopping sequence is cycled through its successive addresses by a binary counter, which is driven at the hopping rate, f_{HOP}. In general, the number of FH channels may be less than the number of available memory locations in the PROM. In this case, the counter must be augmented with a simple circuit to reset the counter at the end of each hopping cycle.

When the transceiver is switched from receive to transmit (typically, by grounding a push-to-talk line), a reset pulse, having a duration of one or more hops, is applied to the counter. This returns the synthesizer to the initial frequency of the hopping sequence. The hopping sequence then proceeds at a rate equal to the quiescent frequency of the voltage-controlled oscillator (VCO). Optionally, a separate crystal-controlled clock (not shown in Figure 14) could be used to control the hopping rate during transmissions.

When the transceiver is switched from transmit to receive, the counter is again reset, returning the synthesizer to the initial frequency in the hopping sequence. However, this time the

* The PROMs are programmed by the equipment manufacturer, not by the user.

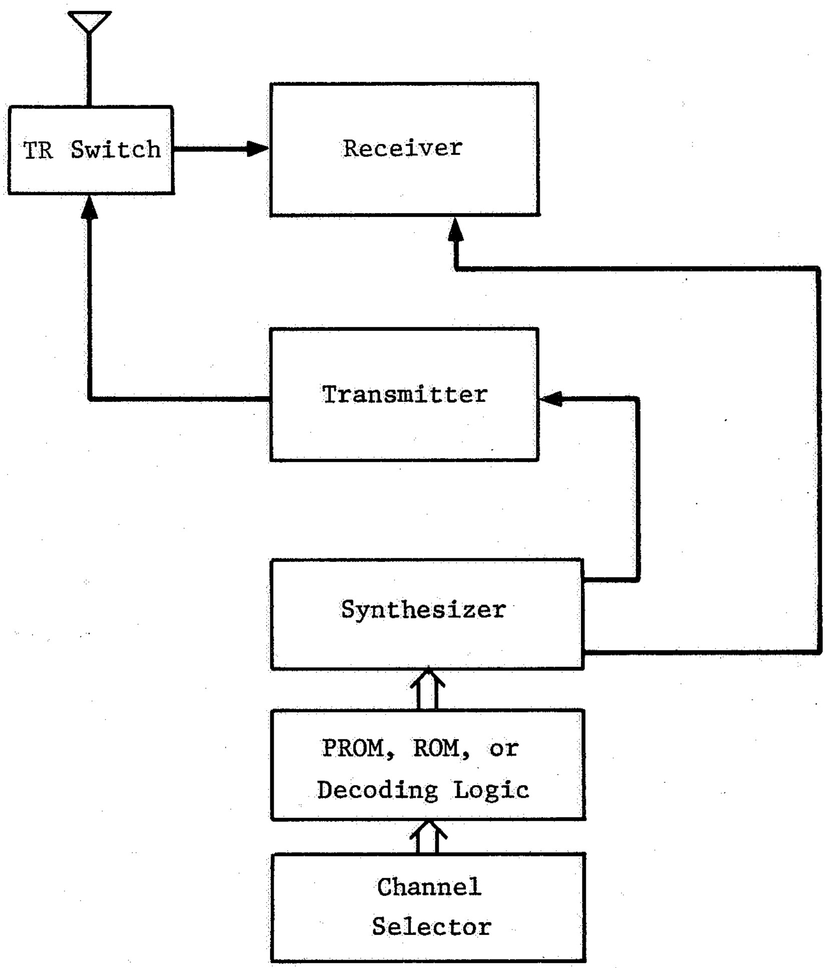

FIGURE 12
SIMPLIFIED BLOCK DIAGRAM OF A CONVENTIONAL HALF-DUPLEX SYNTHESIZER-DRIVEN TRANSCEIVER

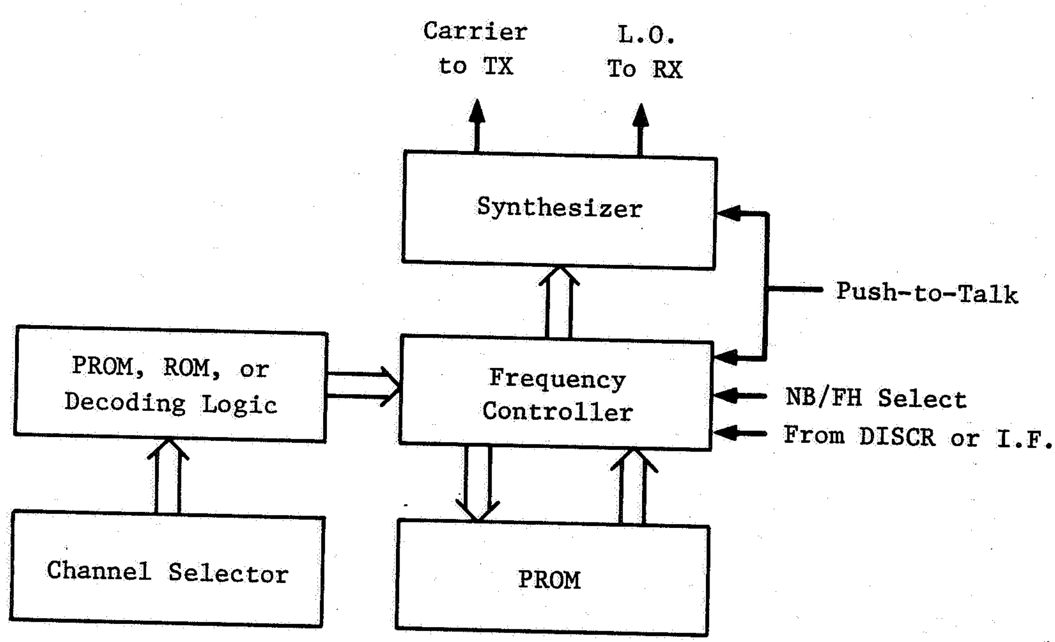

FIGURE 13
BASIC DESIGN MODIFICATION FOR FREQUENCY HOPPING

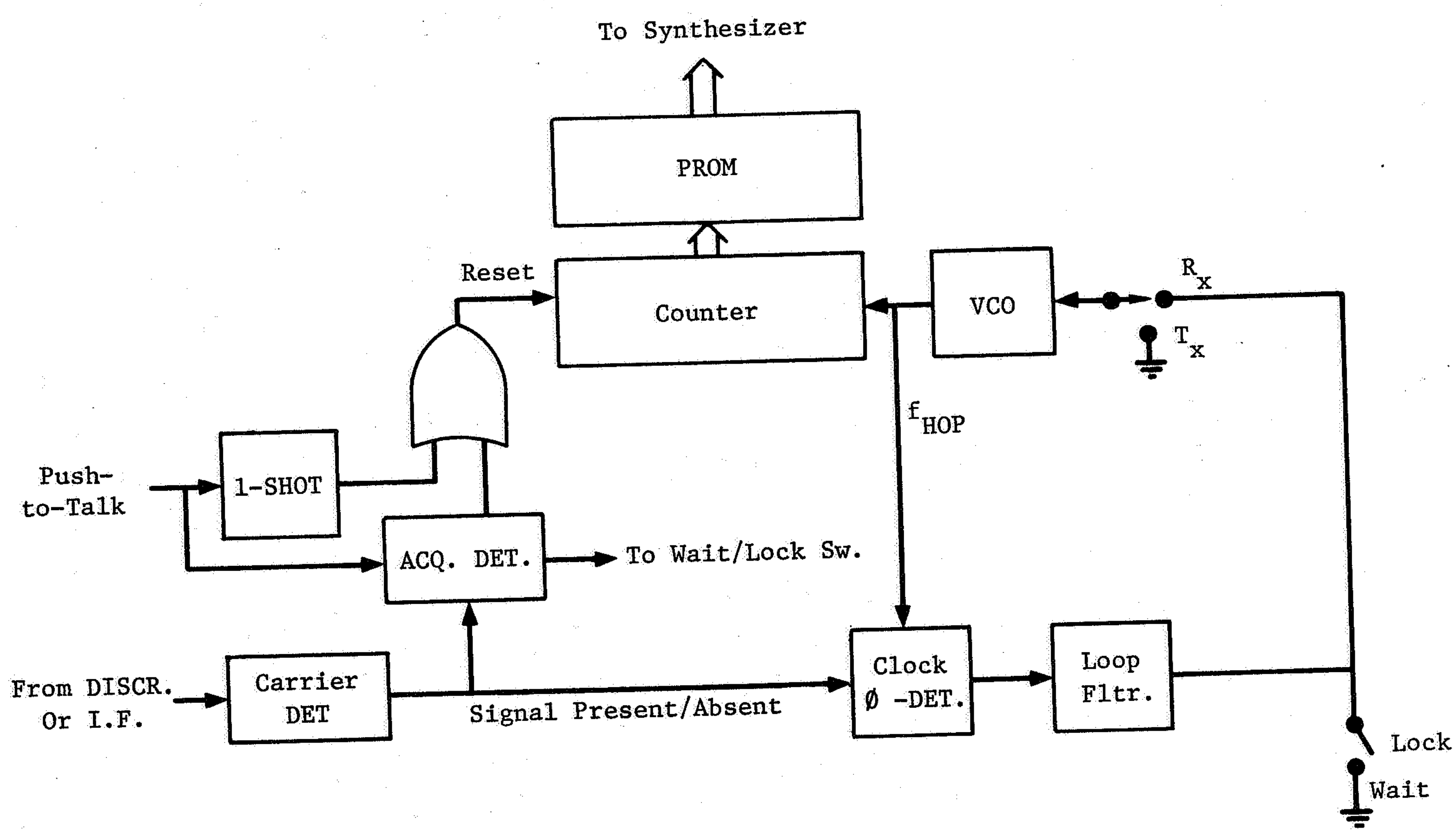

FIGURE 14
FH CONTROLLER

counter is held in the reset condition until the presence of a signal is detected on that frequency. At this point, the counter is enabled and the hopping sequence proceeds. A phase detector measures the phase error of the VCO output with respect to the hops of the received signal. This error signal is used to correct the VCO frequency in a conventional tracking loop. If a signal is detected on each of several subsequent hops, acquisition is declared. If the signal is absent on subsequent hops, or if it disappears for a significant interval after acquisition has been declared, the counter is reset and the receiver returns to the initial frequency in the hopping sequence to wait for a signal.

A simplified state diagram of the process described above is shown in Figure 15. In operation, the receiver may experience several false starts if it encounters interference on the initial hopping frequency, f_o. After each false start, it returns to f_o in search of a valid FH signal.

Note that operating the transceiver in a conventional (non-FH) mode merely entails controling the synthesizer from the channel selector rather than from the FH PROM. In the FH mode, the channel selector is disabled.

This section has provided a brief description of how a slow-FH, half-duplex transceiver might be easily implemented. Alternate implementations are obviously possible. For example, it might be desirable to use a pseudo-random code generator instead of a PROM to generate the hopping sequence. Such an approach would reduce the number of possible hopping sequences, but would facilitate "tuning" the receiver to any permissible hopping sequence. A very small amount of added logic could be used to restrict the transmitter to a particular hopping sequence, if this is deemed necessary from a regulatory viewpoint.

4.1.5 Factors Affecting Performance

In conventional FM mobile radio systems, performance is limited primarily by a combination of noise, man-made interference and fading. In a slow FH system, some of these factors have different manifestations. In addition, other factors, such as synchronization errors, can degrade performance. The following is a general description of these factors. Specific performance predictions will be presented in Section 4.1.9.

4.1.5.1 Noise

For the purposes of this report, the term "noise" includes receiver front-end noise, natural background noise, and certain types of incidental man-made noise, such as automotive ignition noise. Manufacturers of land mobile communication equipment commonly specify receiver noise in terms of sensitivity rather than noise figure or noise temperature. For receivers operating in the 800-900 MHz region, typical sensitivities range from 0.35 to 0.5 microvolts (across 50 ohms) by the 20 dB quieting method or 0.2 to 0.35 microvolt (across 50 ohms) by the 12 dB SINAD* method. These sensitivities correspond roughly to noise figures in the range of 9 to 14 dB (see Reference 91). Lower noise figures are obviously feasible but probably at increased cost (92). Even at the sensitivities of current equipment, however, the overall system noise is often dominated by sources external to the receiver. Man-made background noise in the vicinity of 1 GHz is typically on the order of 10 microvolts per meter in a 10 kHz bandwidth (93,94). At lower frequencies, the level of man-made noise increases.

The impact of wideband background noise, whether man-made or natural, will be approximately the same for slow FH and for conventional narrowband FM systems having comparable parameters.

4.1.5.2 Interference

Interference from narrowband communication signals can include co-channel and adjacent channel interference, intermodulation products, harmonics, spurious emissions and interference due to spurious receiver responses. In addition, the hypothetical frequency plan (Figure 11) shows an allocation in the 915 MHz ISM band; in this case interference from ISM devices can be anticipated (see Section 3.4.2). Based on measured data (79,80) applied with an assumed propagation loss of $1/R^3$ to $1/R^4$ and a 15 kHz i.f. bandwidth, it appears that a single 915 MHz microwave oven can produce a 3 dB degradation in receiver sensitivity over a maximum range of about four to twelve miles. This estimate is based on a typical receiver sensitivity of 0.35 microvolts (EIA SINAD).

Finally, the FH system will suffer self-interference from an aggregate of FH signals. Figure 16 illustrates the impact of two types of interference on a slow FH system. The horizontal axis represents time; the vertical axis represents the interference level of the output of the i.f. filter. In Figure 16a, a hypothetical single narrowband interferer is received near the center of one of the FH channels. Interference pulses, having a duration of one hop, are produced whenever the receiver hops to

* SINAD = (Signal + Noise + Distortion)/(Noise + Distortion)

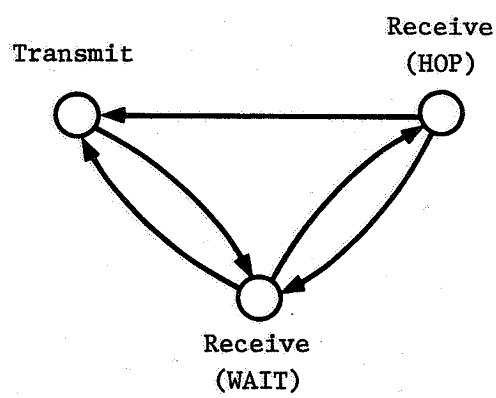

FIGURE 15
SIMPLIFIED STATE DIAGRAM FOR F.H. TRANSCEIVER

the channel occupied by the narrowband interferer. It is assumed that the interferer is not so strong that it is visible through the stopband of the i.f. filter. Figure 16b shows the analogous result for interference from a single FH signal. In this case, the interference pulses are of random duration (with a maximum duration of one hop and an average duration of 1/2 hop) because the clocks that control hopping in the receiver and in the interfering transmitter are not synchronized.

The types of interference depicted by Figure 16 can be described as co-channel interference. As in conventional narrowband systems, adjacent-channel interference will also occur whenever the interference in one or more of the (hopping) adjacent channels is sufficiently strong. Interference that is two or more channels removed from the intended signal can also occur. The approach described here provides no protection against wideband interference, since, in this case, each FH channel is subject to interference.

4.1.5.3 Propagation Factors: Fading and Shadowing

The characteristics of land mobile propagation have been extensively studied and documented (see, for example, References 46-49 and 85). One of the outstanding characteristics of this type of channel is the usual lack of a direct line-of-sight between the base station and the mobile with which it is communicating. Under this condition, the link is characterized by multiple reflected or diffracted paths. The transmission coefficient of the channel is modeled as a narrowband Gaussian stochastic process (95). As a result, the envelope of a CW signal transmitted over this type of channel is Rayleigh-distributed at the receiver input. The time variations of the fading envelope are a function of the vehicle speed, measured in wavelengths per unit time. For example, at 30 miles per hour a 900 MHz carrier experiences 20 dB fades (relative to the mean-square level) at an average rate of about 11 per second. The average (20 dB) fade duration is about 1.0 ms.

This model may be excessively pessimistic when a direct path exists between transmitter and receiver. Under such conditions, a Ricean fading model (which provides for one specular component plus a diffuse component) may be more appropriate.

A particular manifestation of the fading environment as it relates to slow frequency hopping is that the de-hopped signal level will abruptly change whenever the frequency difference between successive hops exceeds the coherence bandwidth of the channel. The coherence bandwidths of land mobile radio channels

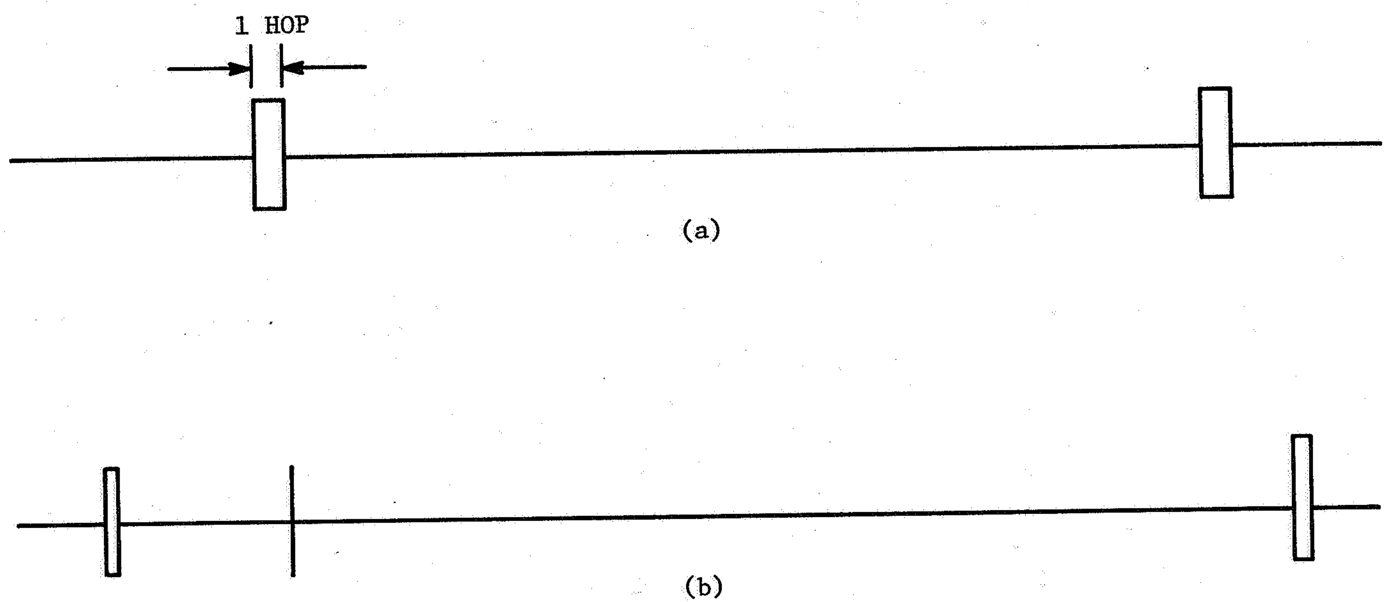

FIGURE 16
INTERFERENCE IN A FH SYSTEM

vary from 25 kHz to 500 kHz (95), so this can be expected to happen often in a FH system occupying several megahertz of bandwidth. This feature may actually be desirable, since it limits the length of any fade, even at low vehicle speeds. Thus, a measure of frequency diversity is provided.

4.1.5.4 Synchronization Errors

Synchronization errors can degrade the performance of this FH system in two ways. Figure 17 illustrates the effect of minor synchronization timing errors. Since the signal will not be "seen" when the transmitter and receiver are tuned to different channels, this type of timing error causes the level of the de-hopped i.f. signal to drop to zero periodically. These "glitches" occur at the hopping rate. Neglecting the rise and fall times of the transient waveform associated with the i.f. filter, the fraction of the signal that is lost in this way is equal to the timing error expressed as a fraction of the hop duration. Unless the system is provided with a very fast carrier squelch, the receiver's audio output will contain noise or interference during these "glitches." Depending on the hopping rate, the subjective effect of these periodic audio noise bursts can be quite distracting. For this reason, some type of fast audio blanking or squelch may be required.

A more serious type of degration will occur when the receiver experiences interference over such extended periods of time that it gives up on trying to maintain synchronization and goes into the "wait" mode, listening for a signal on the first frequency of the hopping sequence. The receiver may also simply fail to acquire the intended signal when it appears. However, by the time the interference reaches such a severe condition, the audio output of the receiver will probably have become unintelligible anyway. Further investigation is needed to confirm this.

4.1.5.5 Intrinsic FH Self-Noise

The approach described here does not require that either the transmitter or receiver maintain phase coherence between hops. Thus, even in the absence of synchronization errors, external noise, and interference, the audio output of the receiver will contain periodic "clicks" that correspond to the phase discontinuities between hops. Schreiber (10) has shown that the signal-to-noise power ratio attributable to this phenomenon is approximately $3\beta^2 (f_m \tau /2)$ for low hopping rates. In this expression, which is in good agreement with the exact result for $f_m \tau > 2$, the following notation has been used.

β = modulation index
f_m = information bandwidth
τ = interval between hops = 1/(hopping rate)

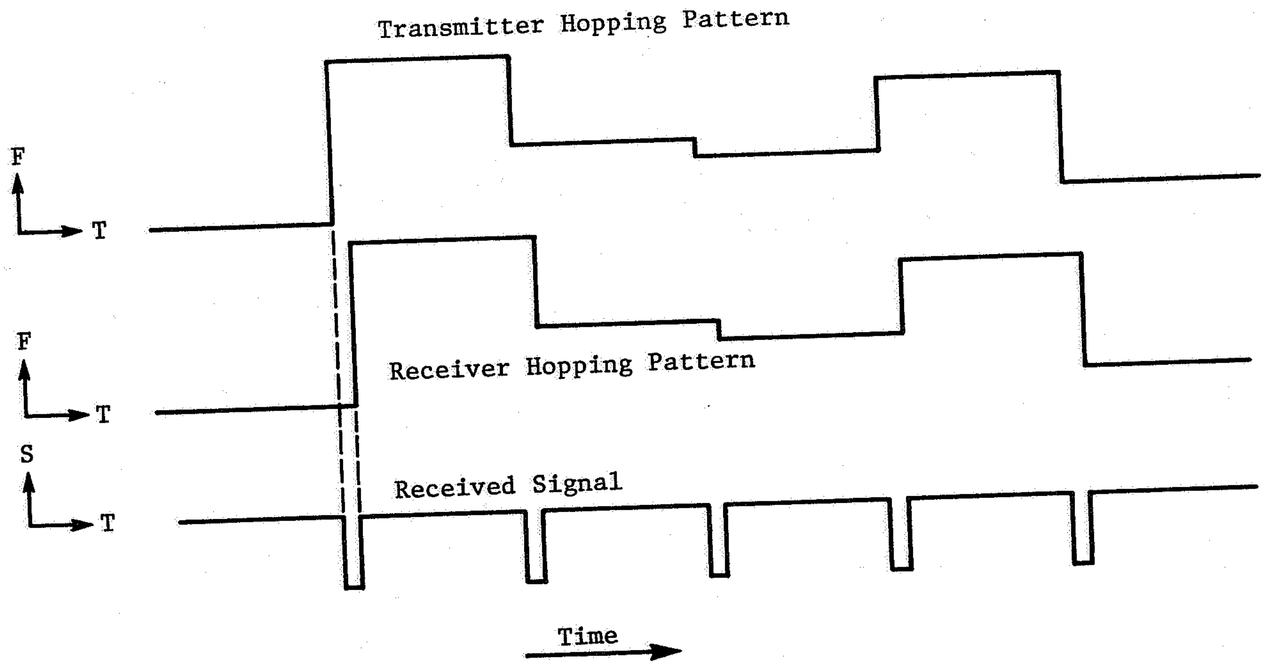

FIGURE 17
SIMPLIFIED TIMING DIAGRAM SHOWING THE IMPACT OF SYNCHRONIZATION ERRORS

For the type of system described here audio SNRs on the order of 25-35 dB are attainable. However, there is currently no information available on the effect of this type of noise on intelligibility or subjective voice quality.

4.1.5.6 Random FM and Harmonic Distortion

Random FM is an effect caused by random variations in the complex channel transmission coefficient. It produces an irreduceable amount of low frequency audio noise at the discriminator output (85). For the type of system under consideration here, random FM limits the audio SNR to about 26dB, for vehicle speeds of 55 miles per hour. At lower speeds, the achievable SNR increases.

Harmonic distortion results primarily from nonlinearities in the transmitter and receiver audio sections and from bandwidth limitations in the receiver's i.f. filter. In conventional systems, total harmonic distortion (measured from transmitter input to receiver output) is less than 15 percent (97). The effects of random FM and harmonic distortion should be about the same for FH systems and conventional systems.

4.1.6 Address Code Design

The problem of constructing frequency-hopping multiple access codes having low mutual interference properties has been considered by Yates (98) and by Einarsson (99). No attempt will be made to incorporate specific codes into this example. However, the problem of address code design may be constrained by implementation considerations. Specifically, if the receiver must be able to select any valid user address code (by key pad command, for example) then it may be desirable to use linear feedback shift register generators to control the hopping sequences. If the transmitters and receivers all use fixed codes, such constraints do not exist for hopping sequences of reasonable length.

4.1.7 Degree of Privacy

One of the potential motivations for considering this approach is the provision of privacy from unauthorized listeners. The protection thus provided is not absolute, since slow FH transmissions can be de-hopped by using a sophisticated, frequency-agile receiver or by using a very wideband receiver at close range to the transmitting antenna. Both of these interception techniques can be made more difficult by increasing the number of channels or the hopping rate, or both. However, for the listener with a coventional narrowband FM receiver (or a frequency-hopping receiver with a different code) the degree of privacy should be high. Miller and Licklider (96) have demonstrated that, when speech is interrupted on an average of at least once per second, and the interruptions have a duty factor of at least 90 percent, the result is virtually unintelligible. This implies that a slow FH system having a hopping rate greater than one per second and at least 10 channels should provide privacy from unsophisticated listeners.

4.1.8 System Parameters

Typical parameters for the type of slow frequency hopping system described here are presented in Table 2. Obviously, these figures do not reflect the result of any in-depth tradeoff study. They are presented for the purpose of illustration only.

One important pair of parameters not listed in Table 2 is the attenuation of adjacent channel and next-adjacent channel interference relative to co-channel interference. These parameters depend on the modulation index, the channel spacing, the frequency tolerances, the design of the receiver's i.f. filter and the amount of "splatter" generated by the hopping process. In the initial simulation runs (Appendix C), 10 dB adjacent channel attenuation and 30 dB next-adjacent channel attentuation were assumed. (Signals more than two channels away on a particular hop are assumed to produce no interference on that hop.) This is representative of what can be achieved using the least expensive monolithic crystal or creamic filters, with channel spacing of 15 to 20 kHz. The parameters were later changed to 60 and 80 dB, respectively, to show the performance of more sophisticated crystal filters and 25 kHz channel spacing. A small number of runs were performed with 200 dB attenuation of adjacent and next-adjacent channel interference to slow an upper limit on the gains that could be achieved in this area.

4.1.9 Peformance and Spectrum Efficiency

In virtually all spread spectrum multiple access systems, there is a tradeoff between communication performance and spectrum efficiency. Indeed, this tradeoff is not limited to spread spectrum multiple access, but is inherent in virtually any approach to sharing the spectrum in which the possibility of mutual interference exists. If high levels of interference are acceptable, then many users can operate simultaneously. If the tolerable level of interference is reduced to improve performance, then, for a given system configuration, the number of simultaneous users must be reduced. The issue of cost is also involved, since increased system complexity can allow improvements in communication performance for a given level of interference, or improvements in spectrum efficiency for a given level of communication performance.

TABLE 2

TYPICAL SYSTEM PARAMETERS

Transmitters	
Modulation:	Narrowband FM, with or without FH
Peak Deviation:	5.0 kHz
Audio Passband:	300 - 3000 Hz
Output Power:	35 Watts
Channel Spacing:	25 kHz
Spurious Outputs:	-75 dB
Hopping Rate:	20/s
Antennas	
Base:	Omnidirectional, 9 dB gain, vertical polarization
Mobile:	Omnidirectional, 6 dB gain, vertical polarization
Receiver	
Sensitivity:	0.35 microvolts (EIA SINAD)
Selectivity:	-80 dB (EIA SINAD)
Intermodulation:	-75 dB (EIA SINAD)
Spurious and Image Rejection:	-100 dB

In order to facilitate the exploration of these tradeoffs for the slow FH technique, two Monte Carlo simulations were developed. The first provides estimates of the statistics of the r.f. signal-to-interference ratio. The second provides estimates of the intelligibility of speech resulting from a slow FH system. Both simulations are described in detail in Appendix C.

The signal-to-interference simulation was a developmental effort that was later supplanted by the intelligibility simulation. Both simulations model the service area as a circular region having a radius of 30 km. In all case, the computed levels of performance refer to a receiver located in the center of this area. The transmitter with which this receiver is in communication (or with which communication is being attempted) is separated from the receiver by a randomly-selected distance. The distribution of distances has a density $p(X) = X/X_o^2 \exp(-X/X_o)$, where $2X_o$ is the mean communication distance. Interfering transmitters have randomly-selected distances (to the receiver) that correspond to a uniform average number of interferers per unit area. Both of these distributions are obviously simplifications. The signal-to-interference simulation measures the probability of attaining a given r.f. signal-to-interference ratio. The intelligibility simulation measures the probability of attaining a given articulation score. In both simulations, the Longley-Rice propagation model (46-48) is used to estimate the mean and variance of the short-term-average signal and interference levels at the receiver input.

The simulations indicate that it is probably feasible to provide adequate communication performance with slow frequency-hopping FM. The spectrum efficiency of this approach, in terms of the maximum number of active users per channel over a given geographic area, depends on several factors, including the mean communication distance and the degree of adjacent channel interference. Under favorable conditions, the slow FH technique may be able to accommodate at least as many users per channel as conventional systems having uniform channel loadings. With less favorable conditions, conventional systems having uniform channel loadings will provide better spectrum efficiency than slow frequency hopping for a given level of performance.

The estimated maximum number of active users per channel (in a geographic area having a radius of 30 km), is presented in Table 3 for various operating conditions, based on the results of the intelligibility simulation. The first column represents the attenuation of adjacent and next-adjacent channel interference relative to co-channel interference. The second column is the mean distance over which communication is attempted. The third

TABLE 3

ESTIMATED MAXIMUM NUMBER OF USERS PER CHANNEL
FOR INTERFERENCE-LIMITED OPERATION

Splatter, dB Attenuation	Mean Communication Distance, km	Type of Link	Estimated Maximum Number of Users Per Channel
10/30	8.0	MB	0.3
10/30	4.0	MB	0.3
200/200	4.0	MB	1.0
60/80	16.0	MB	0.5
60/80	8.0	MB	0.6
60/80	4.0	MB	0.9
60/80	2.0	MB	1.4
60/80	4.0	MM	0.9
60/80	2.0	MM	1.4

column shows whether the link is mobile-to-base (MB) or mobile-to-mobile without repeaters (MM). The final column represents the estimated maximum number of active users per channel, based on the assumed requirement for a 90 percent chance of achieving a word articulation score of 0.75.

Two major results are evident:

(1) The degree of adjacent channel and next-adjacent channel interference can have a substantial effect on performance and spectrum efficiency. The simulation runs that are representative of the least expensive implementations show relatively poor results. Significant improvements in spectrum utilization could accrue if these minimal implementations are averted in favor of commercial-quality communication equipment with 25 kHz channel spacings.

(2) The maximum number of users per channel is a function of the mean communication distance. This relationship is shown in Figure 18 for mobile-to-base operation and 60/80 dB adjacent/next-adjacent channel attenuation. Obviously, short communication distances correspond to high signal-to-interference ratios. Thus, more users can be accommodated per channel for short range communication than for long range communication. This is true for conventional narrowband communication systems as well as spread spectrum communication systems. For conventional narrowband systems, however, the maximum number of users per channel at any given time is limited to integer values. (If two conventional users can be accommodated on one channel over a given geographic area, than two hundred conventional users can be accommodated on one hundred channels over the same area, assuming that all channels are equally loaded. The number of conventional users on a particular channel at a particular time can only be 1, 2, 3,...). Simulation runs with one channel indicate that if two conventional users are to share a single channel, the mean communication distance must be less than about 2 km for the parameters used here. This is based on an assumed requirement that the signal-to-co-channel-interference ratio be above 6 dB for at least 90 percent of the transmitter locations. Adjacent channel interference is not included. The relatively low degree of frequency reuse in this case is related to the fact that the transmitters are deployed randomly throughout the service area. (Higher degrees of frequency reuse are possible in cellular systems, but in such systems each channel is available over as little as 14 percent of the service area.)

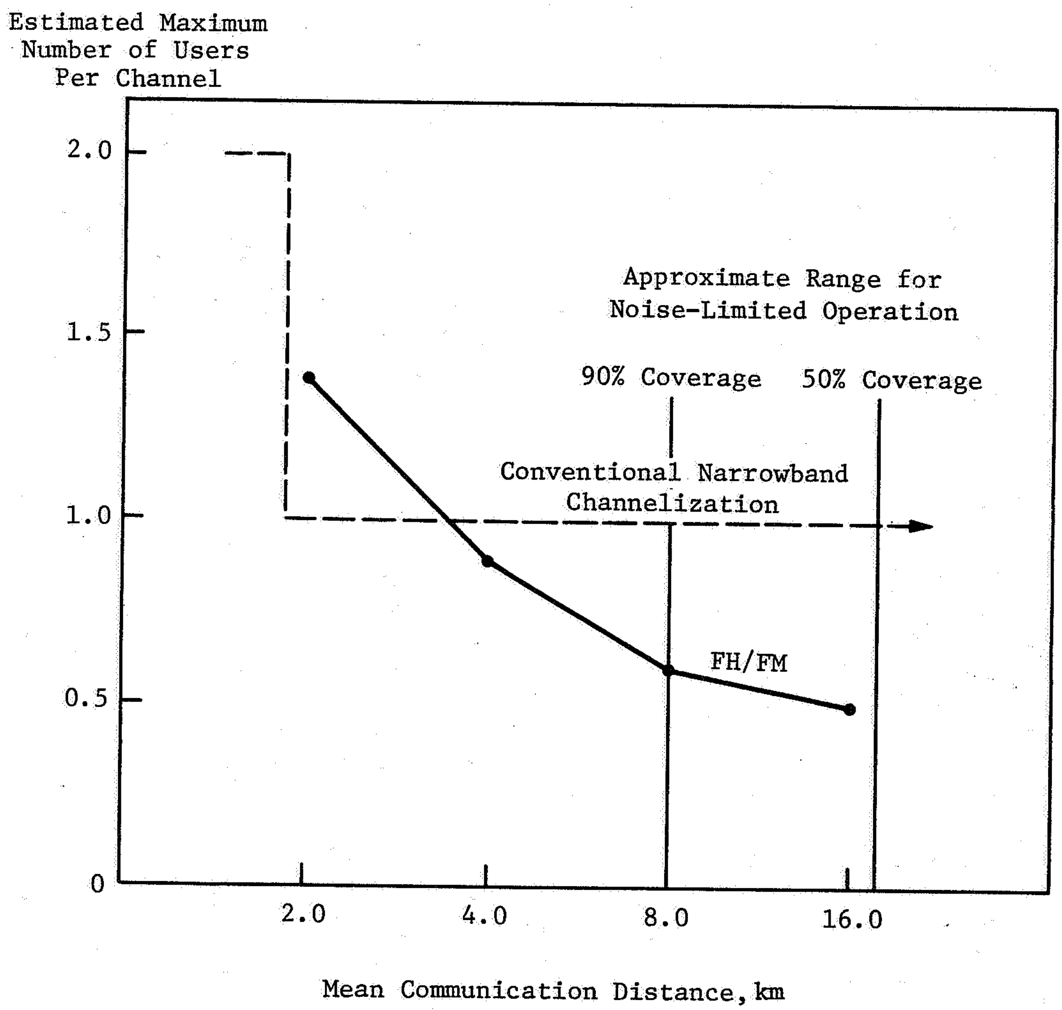

FIGURE 18
MAXIMUM SPECTRUM UTILIZATION AS A FUNCTION OF COMMUNICATION DISTANCE FOR INTERFERENCE-LIMITED SYSTEMS

It can be seen that, for the set of simulation parameters used here, the spectrum efficiency of the slow frequency hopping approach, in terms of the maximum number of simultaneous calls per channel, is comparable to or above the corresponding level for conventional narrowband systems for mean communication ranges of about 2 to 4 km. At shorter ranges, two or more calls per channel can be accommodated with conventional narrowband systems. At longer ranges, the degree of interference suffered by a slow frequency hopping system limits the number of users for a given level of performance.

Some caution needs to be exercised in the interpretation of Figure 18 and the results on which it is based. Specifically, the implicit comparison of the slow frequency hopping approach to conventional FM techniques in terms of spectrum utilization is credible only if both techniques have the same channel spacing and the same voice quality. It is always possible to trade voice quality for improved spectrum utilization in both techniques by simply reducing the channel spacing (with or without a concomitant reduction in the modulation index). This issue can probably be resolved only by experiments which would be beyond the scope of this study.

In the absence of interference, the communication range will be limited by Rayleigh fading and background noise. As shown in Appendix C, this range will be limited to about 19.3 km (12.1 miles) for 50 percent coverage or 8.3 km (5.2 miles) for 90 percent coverage, even under favorable conditions. In urban areas, the maximum range will be substantially below these values. In the presence of combined interference, fading, and background noise, the maximum range will be further reduced.

In summary, it is clear that the slow frequency hopping approach and the conventional narrowband approach have different characteristics and may be suited to different applications. Whether the predicted performance and spectrum efficiency of the slow frequency hopping approach are compatible with the needs of various user groups remains to be seen.

4.1.10 Repeater Operation and Full Duplex Operation

The feasibility of a slow FH voice communication system does not depend on the use of separate bands for mobile-to-base and base-to-mobile transmissions. Half-duplex operation can be achieved using a common band for both link directions, as in the current 27 MHz citizens band. However, if the uplink and downlink bands are split, the possibility of repeater operation and full duplex operation is preserved. Each of these functions may be beyond the economic reach of the average CB hobbyist, but they may find other applications. Indeed, some types of service may not be practical without the use of repeaters.

Repeaters receive on the mobile-to-base frequency and simultaneously transmit on the base-to-mobile frequency, thus providing for extended-range mobile-to-mobile communication. In a frequency hopping system, it is envisioned that a repeator would consist of one or more FH receivers and FH transmitters connected back-to-back and sharing a common antenna through a duplexer*. Other arrangements (such as retransmitting the entire mobile-to-base band with a wideband linear repeater) are obviously possible, but are probably not as efficient. For any repeater arrangement, it may be necessary to include a digital code in the beginning of each mobile transmission addressed to the repeater. This would provide for immediate activation of the repeater in response to mobile transmissions addressed to it, while allowing the repeater to ignore other transmissions.

The use of repeaters to extend communication performance would degrade spectrum efficiency, since a single mobile user occupies an uplink and a downlink simultaneously. This is also true for conventional systems.

Full duplex operation would allow two users in mutual communication to talk simultaneously, telephone-style, at the expense of increased equipment cost and decreased spectrum efficiency. Again, each full-duplex link requires two simultaneously-active transmitters instead of one.

4.2 Sensor Systems Using Homodyne Detection

One of the significant practical disadvantages of spread spectrum systems is the requirement for acquisition and synchronization of the spread spectrum waveform. In applications where the transmitter and receiver (for a given link) are co-located, this requirement may be mitigated or averted. Particular examples of such applications (where conventional narrowband signals have been used to date) are police radar and commercial microwave intrusion detectors. Both types of system have the basic configuration shown in Figure 19. A C.W. Signal is radiated, and reflected returns are detected by using a small fraction of the transmitted signal as a local oscillator signal. The result is an audio waveform having a frequency equal to the Doppler shift of the object from which the reflection was produced. Reflections from stationary objects result in d.c. components, which are ignored.

---

* The ability of a duplexer to accommodate frequency hopping will depend on the separation between the transmit and receive bands and on the bandwidth over which the transmitter is hopped. The bandwidths of current commercially available duplexers range from a few hundred kilohertz or less to a few megahertz.

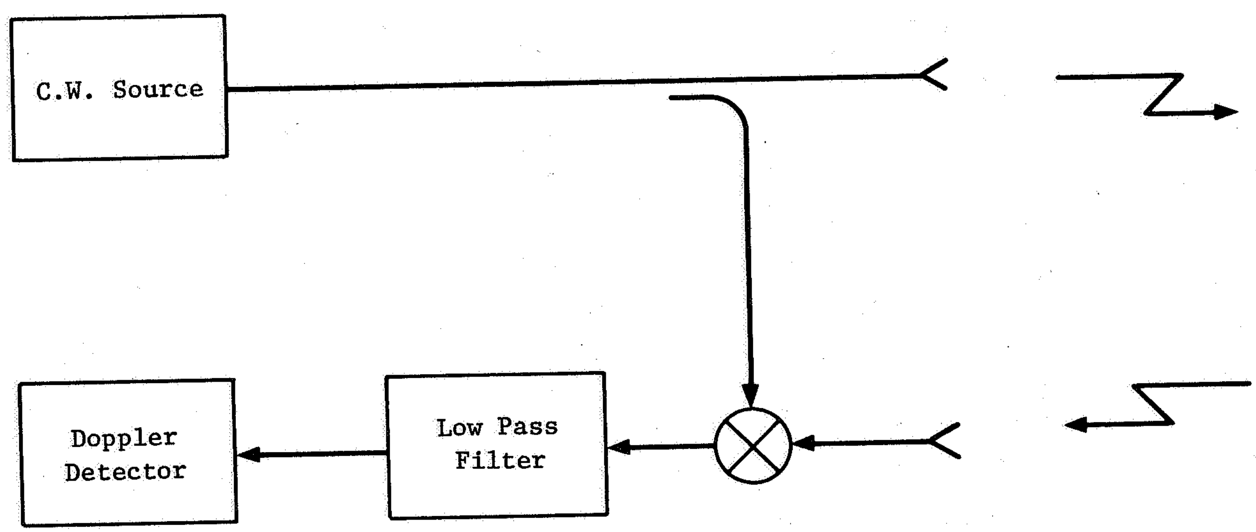

FIGURE 19
CONVENTIONAL HOMODYNE SYSTEM

The minimum assigned bandwidth required by this type of system depends almost entirely on the frequency tolerance of the oscillator used to produce the transmitted signal.

Microwave intrusion detectors provide surveillance of an area or volume around the sensor or in a particular direction relative to the sensor. They normally operate under Part 15 of the Rules and Regulations.

One disadvantage of this type of intrusion detector is its susceptibility to false alarms resulting from reflections from unprotected areas (adjoining rooms or buildings, for example) and from radio frequency interference (106). In personnel detection radar designed for military applications, these limitations have been mitigated by applying wideband modulation to the transmitted signal (107).

Figure 20 illustrates a hypothetical application of such techniques to a commercial intrusion detector. This configuration differs from the one shown in Figure 19 in that provision has been made for frequency-hopping the transmitted signal at a fast rate. In addition, the reference signal has been delayed by an amount Δ. The frequency difference between adjacent hops is assumed to be large with respect to the maximum Doppler shift, but small with respect to the transmitter's center frequency.

On any particular hop, the transmitted signal is:

$$\cos(\omega_o t + k\omega_m t) \qquad o < t < \tau$$
$$k = -n, -n + 1, \ldots 0.\ 1, \ldots . n$$

where τ is the duration of one hop and ωm is the angular frequency difference between adjacent hopping "channels." For a specular reflector at a range r, the received signal is of the form:

$$\cos[\omega_o(t-2r/c) + k\omega_m(t-2r/c)]$$
$$= \cos[\omega_o t + k\omega_m t - (2r/c)(\omega_o + k\omega_m)],\ 2r/c < t < \tau + 2r/c$$

The reference signal is proportional to

$$\cos[\omega_o(t-\Delta) + k\omega_m(t-\Delta)],\ \Delta < t < \tau + \Delta$$

The output of the low pass filter is then

$$\cos[\omega_o(\Delta-2r/c) + k\omega_m(\Delta-2r/c)],\ \max(2r/c,\Delta) < t < \tau + \min(2r/c,\Delta)$$

$$\approx \cos[\omega_o(\Delta-2r/c)] \quad \text{for } n\omega_m << \omega_o$$

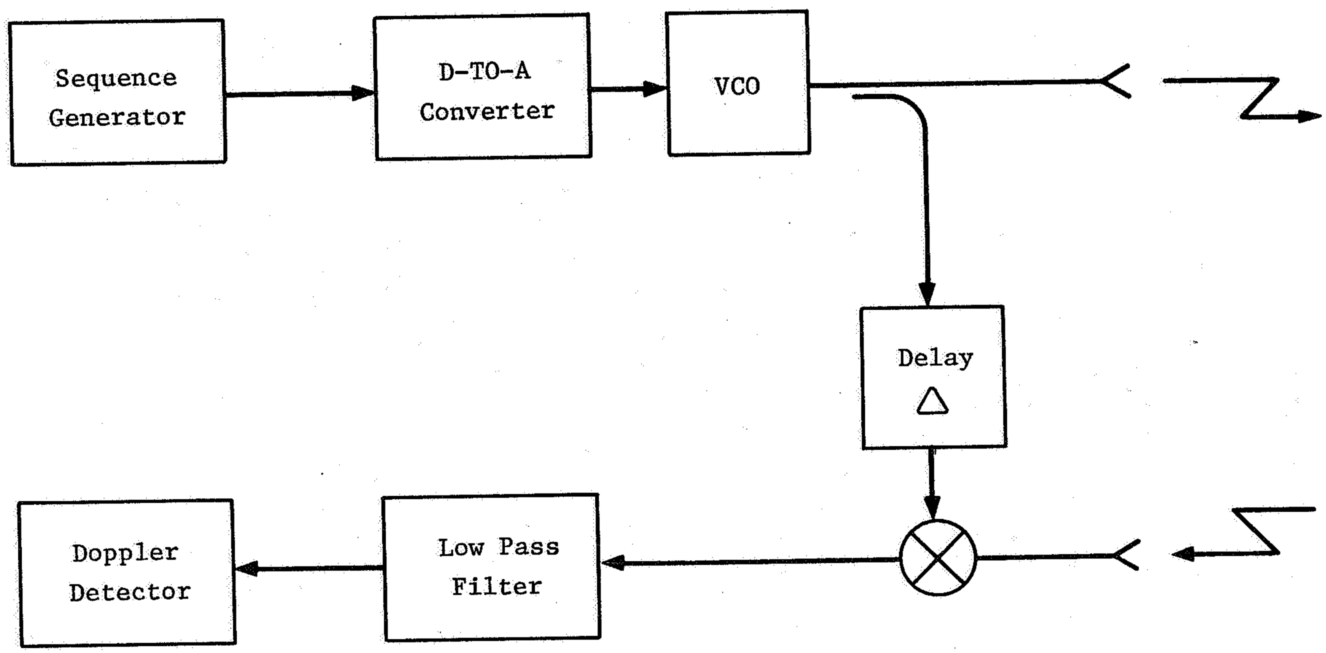

FIGURE 20
RANGE-LIMITED, INTERFERENCE-RESISTANT HOMODYNE SYSTEM (FH VERSION)

For $|\Delta - 2r/c| << \tau$, the output of the low pass filter is a sinusoid at the Doppler frequency, $\omega_d = 2\dot{r}\omega_o/c$. For $\tau < |\Delta - 2r/c|$, the received signal and the reference signal differ by a multiple of ω_m, so the resulting sinusoid is above the cutoff of the low pass filter. The receiver is insensitive to such returns. For $0 < |\Delta - 2r/c| < \tau$, the output of the mixer consists of the samples of a sinsuoid at the doppler frequency interrupted by samples of sinsuoids at frequencies above the cutoff of the low pass filter. Since the sample rate is equal to the hopping rate (which is much greater than $2|f_d|$), the output of the low pass filter is a sinsuoid at the Doppler frequency having an amplitude proportional to $1-|\Delta - 2r/c|/\tau$. The receiver also exhibits reduced sensitivity to narrowband interference or interference from other sensors, since such sources produce only short pulses at the mixer output.

Example

Let $\Delta = 100$ ns, $f_H = 1/\tau = 5.0$ MHz.

The maximum sensitivity of the sensor to returns of a fixed amplitude occurs at $r = c\Delta/2 = 50$ ft. At ranges between zero and 50 ft., the sensitivity increases monotonically with increasing range for returns of a given amplitude. For ranges greater than 50 ft., the sensitivity monotonically decreases out to a maximum range of $c(\tau + \Delta)/2 = 150$ ft.

Analogous examples can be produced for intrusion detectors based on direct sequence techniques.

The extension to police radar is obvious. The FCC has received numerous complaints about jamming of these devices by motorists (113). The addition of spread spectrum capabilities would appear to be an economically viable long-term approach to this problem.

5. SPREAD SPECTRUM DEVELOPMENT AND IMPLEMENTATION: COSTS AND RISKS

In this section, a number of potential negative implications of spread spectrum technology will be addressed. These include: the risk of increased interference to conventional users, the difficulty of monitoring and the economic burden on the spread spectrum user.

5.1 The Risk of Increased Interference

Current users of the spectrum are likely to be concerned about potential interference from spread spectrum signals. Obviously, it is no more possible to give general a priori assurances on this issue with spread spectrum than with conventional modulation techniques. Proposed modes of operation will have to be evaluated individually to assess their potential for interference. However, it is reasonable to expect that well-conceived spread spectrum applications will be able to evolve with acceptably low levels of degradation to existing services.

5.1.1 In-Band Interference

The most severe type of interference will occur when conventional receivers share a band with spread spectrum transmitters. For continuous direct sequence emissions, the fraction of the received spread spectrum power falling into the receiver passband may be as high as B_{if}/R_c, where B_{if} is the i.f. bandwidth and R_c is the spread spectrum chip rate. For example, if B_{if} = 10 kHz and R_c = 10 MHz, the isolation between the interferer and the receiver is only 30 dB. This assumes that the receiver is tuned to the center frequency of the spread spectrum signal, that B_{if} is large with respect to the code repetition rate (but small with respect to the chip rate, R_c) and that the interfering signal is not so strong as to cause nonlinear amplification prior to the i.f. filter. Such a low degree of isolation would clearly be unacceptable for some applications. On the other hand, it might be acceptable for satellite applications in which the received power levels are constrained to narrow range or for terrestrial applications involving very low spread spectrum power levels or very low spread spectrum duty cycles, for example.

In-band interference from frequency hopping signals can take several forms, depending on the system parameters and the ratio of signal power to interference power (65). When the FH interference power at the receiver input is not greatly larger than the power of the intended signal (so that interference occurs only when the spread spectrum "hops into" the i.f. passband of the victim receiver) and the hopping rate is low compared to the i.f. bandwidth, interference will take the form of short pulses. During each pulse, the signal-to-interference ratio at the i.f. output will be equal to the signal-to-interference ratio at the receiver input. However, for a single FH interferer using N channels equally, this interference will occur for only a fraction of the time equal to 1/N. (A particular FH signal hopping over 100 channels will cause interference to a particular narrowband receiver only one percent of the time under such conditions.)

When the interfering FH signal power greatly exceeds the power of the intended signal, or when fast hopping rates are used, interference may occur even when the FH signal is not centered within the i.f. passband. The interference can result from sidelobes (i.e., "splatter") produced by the hopping process or from sidelobes of the modulation process used to impose the transmitted information on the FH signal. In addition, intermittent interference can occur as a result of intermodulation products generated by the interaction of the FH signal with other signals in the early receiver stages.

5.1.2 Out-of-Band Interference

Like conventional signals, spread spectrum signals can cause interference to services operating in adjacent bands. For direct sequence and fast frequency hopping signals, the major risk is likely to be from the sidelobes of the spread spectrum modulation. Sidelobes can be minimized by shaping the transmitted pulses and by using continuous-phase modulation. However, these measures are not without economic consequences and they may, in some cases, degrade the communication performance of the spread spectrum system.

For well-designed slow FH systems, the hopping sidelobes can be low with respect to the sidelobes produced by the information process, particularly when FM is used. Such systems may produce no more out-of-band interference than conventional signals.

5.2 Monitoring and Enforcement

Since the Commission is charged with the enforcement of its Rules and Regulations, the issue of whether to allow the use of emissions that are difficult to monitor must be addressed. This issue appears to have a number of major facets, including the ability to monitor message content and operational procedures, the ability to monitor technical characteristics such as radiated power, modulation parameters, and frequency stability, and the ability to collect statistical data on spectrum occupancy.

To the extent that any radio system provides privacy from unauthorized listeners, that system will be more difficult to monitor for message content and operational procedures than "clear voice" systems. This does not mean that such monitoring will be impossible; only that it will be more costly, more time-consuming, or less effective than it otherwise would be. The problem is not new, nor is it in any way unique to spread spectrum. (See, for example, FCC Dockets 18108, 18261, and 21142, which all relate to digital communication.) Neither is it limited to systems designed to provide privacy. Clear text digital transmissions are generally more expensive to monitor than voice transmissions simply because of the cost of the terminal equipment.

The problem of monitoring may be exacerbated, however, for spread spectrum systems in which the number of possible user codes is large. As a hypothetical example, consider a 1.2 MHz band that is channelized at 25 kHz increments. There are $1200/25 = 48$ conventional channels to be monitored. However, if this bandwidth is used by a frequency hopping system having the same channel spacing, the number of possible user codes is $48! = 1.24 \times 10^{61}$, assuming that each code (i.e., hopping sequence) uses every channel once and only once. (The number of possible codes increases if this assumption is relaxed or removed.) Although it can be argued that constraints on mutual interference would prevent all of these codes from being available for use, the point is that the number of possible user codes is so large that the probability of finding an active user on a randomly-selected code is negligibly small. Similar examples can be produced for direct sequence signals.

The situation just described could make any type of monitoring activity impractical if no remedial measures were available. Rules and regulations would be simply unenforceable. Unlicensed operations could grow out of control. Fortunately, however, the Commission has a number of options that could be used to significantly mitigate or eliminate this problem. For example, the Commission could:

1. Use its regulatory powers to allow the sale of equipment having a capability for multiple user codes only in cases where a demonstrated need for such a capability exists. Restrict the number of codes according to demonstrated need.

2. Require that user codes be assigned on a permanent basis and not be subject to modification by users.

3. Require the registration of user code assignments with the FCC.

4. Authorize spread spectrum emissions only for services in which enforcement problems are manageable without extensive monitoring.

5. Authorize only spread spectrum techniques that can be "decoded" using a conventional wideband receiver at short range, such as frequency hopping with AM*, frequency-hopped on-off keying, or chirped pulse modulation.

6. Employ specialized monitoring equipment that would allow a user code to be identified by observing the signal over a sufficient period of time, along with spread spectrum receivers having selectable user codes. Allow only techniques for which such decoding is practical.

In identifying these options, no recommendation is made as to their suitability for specific cases. The intent is simply to list a number of rather obvious regulatory approaches that should be considered. Various combinations of these measures may be appropriate for particular applications.

5.3 Development and Implementation Costs

The ultimate utility of spread spectrum techniques in non-Government applications will depend strongly on economic factors. The majority of all spread spectrum equipment that has been designed and produced to date has been intended for military and aerospace applications. The relatively high cost of this equipment (typically at least ten to twenty thousand dollars per user terminal) often reflects requirements for features such as:

- resistance to intentional jamming and "spoofing,"
- a high degree of information security,
- interoperability with other equipment,
- survivability and operability in extreme environments,
- power, weight, and volume restrictions, and
- very high reliability.

Although some non-Government applications may require one or more of these features, the standards for commercial equipment are not usually as severe as the standards for military and aerospace equipment. For this reason, it is likely that commercial spread spectrum equipment, if produced in sufficient quantities, could be sold for prices significantly below the price of comparable military or aerospace equipment.

---

* i.e., double sideband AM with carrier: A3 emission with frequency hopping.

Because potential non-Government markets for spread spectrum communication equipment tend to be specialized, poorly defined, or poorly recognized, efforts to design and develop such equipment on a commercial basis have been limited.

The development of reliable, absolute cost estimates is usually an expensive, high-risk endeavor. Prices ultimately reflect such factors as the market size and market share as perceived by individual producers, the degree of competition in the market, unforeseen development or production problems, inflation, and other relatively unpredictable factors. The detailed development of such absolute price or cost estimates would be beyond the scope of this study. Nevertheless, it is possible to provide a coarse ranking of the likely relative costs of several of the spread spectrum concepts that have been described in this report.

The results of this ranking are summarized in Table 4. The basis of this table will be discussed in the following paragraphs. The six techniques or system configurations that are included in the table have been grouped into categories of high, moderate, and low cost. The order of the entries within any of these categories is arbitrary except as noted below. Each of the six techniques was placed in its respective category on the basis of the likely increase in commercial price relative to the price of comparable conventional narrowband equipment, not on the basis of absolute price level. This increase includes the impact of development costs, where applicable. Production within the next five years is assumed.

The "high" cost category includes configurations that seem likely to sell for at least 1.8 times the price of comparable commercial narrowband equipment. The corresponding factor for "moderate cost" configurations is 1.2 to 1.8, and for "low cost" configurations, 1.2 or less. The definition of "comparable" commercial narrowband equipment clearly requires some subjective judgment. Also, "low cost" configurations could become more expensive over the long run due to unforeseen requirements; "high cost" configurations could become less expensive over the long run in the event of unforeseen technological breakthroughs. Before taking even the coarse categorization of Table 4 as a solid ranking, such risks need to be taken into account.

The basis for the cost/price ranking of each of the six configurations included in Table 4 will now be discussed.

5.3.1 Fast Frequency Hopping/DPSK

The specific FFH/DPSK technique under consideration here is described in Section 3.2.2.1. It is potentially one of the most

TABLE 4

ESTIMATED RELATIVE COSTS OF SIX SPREAD SPECTRUM CONFIGURATIONS

SYSTEM CONFIGURAITON	POTENTIAL APPLICATIONS	STATE OF DEVELOPMENT
	HIGH COST	
Fast Frequency Hopping/DPSK	High Capacity Land Mobile	Concept
Fast Frequency Hopping/FSK	High Capacity Land Mobile	Concept
	MODERATE COST	
Direct Sequence Modems	Voice/Data Communications	Fully Developed and Tested for Satellite Applications
SAMSARS	Maritime Distress Alerting	Engineering Models have been Developed and partially tested
	LOW COST	
Slow FH/NBFM	Personal Radio Service Private Land Mobile Radio	Commercial applications untested
Homodyne Detectors	Intrusion Alarms Police Radar	Tested for military applications

expensive spread spectrum techniques ever suggested for use under the regulatory aegis of the FCC. The factors contributing to this conclusion include the following.

- State of Development. Although a significant number of analyses and simulations have been performed, no equipment appears to have been developed, even on an experimental basis. In view of the uniqueness of the configuration, development costs could be significant.

- The Need for Digitized Speech. Digital speech techniques vary in cost, in voice quality and in susceptibility to transmission errors. However, the least expensive digital speech techniques providing reasonable voice quality at 30 kbits/s are likely to be more expensive than conventional analog transmission techniques.

- The Need for Coherent Frequency Synthesis. The frequency synthesizer (or other waveform generator) incorporated in the transmitter must be capable of hopping rates on the order of 10^5/s. Moreover, the phase of the transmitted waveform must remain coherent not just from one hop to another, but over at least one entire FH code cycle consisting, typically, of 16 to 32 hops. This is because the demodulation technique requires a phase comparison between chips on successive code cycles.

- The Need for Dynamic Control of the Transmitter Power. This feature is needed to combat the "near-far" problem.

- The Number and Type of Components Required for the Demodulator. Several potential demodulator designs have been suggested (24). Each of these designs requires, at a minimum, multiple bandpass filters and multiple product detectors. In typical designs, the number of each of these components might be on the order of 16 to 32. In addition, the demodulator requires this number of analog delay lines or, alternatively, a single tapped delay line having a delay-bandwidth product that is currently near or beyond the state of the art for charge-coupled devices (CCDs) and surface acoustic wave (SAW) devices. Furthermore, a yet-to-be-developed technique for rapidly and economically changing the FH address code is required if the equipment is to be used in cellular mobile radio systems.

5.3.2 Fast Frequency Hopping/Multilevel FSK

The specific FFH/FSK technique considered here was described in Section 3.2.2.2. The potential cost implications are similar to those just described for FFH/DPSK, except that coherent frequency synthesis is not required, and the demodulator configuration is significantly different. The first of these factors represents a cost advantage for FFH/FSK. The second probably does not, since the FFH/FSK demodulation technique described earlier would require, in effect, that a spectrum analysis of the transmission band be performed on each hop. It has been suggested that this function could be performed with charge coupled devices (CCDs) using the chirp-Z transform algorithm (101) but the details of the demodulator design do not appear to have been investigated.

It is not clear whether dynamic control of the transmitter power would be required with this approach. The obviation of this requirement, together with a successful CCD demodulator implementation, could bring this technique into a lower cost category.

5.3.3 Direct Sequence Modems

The type of equipment configuration to which this section refers is not as general as the title suggests. Examples of the specific types of equipment that belong in this category are described in detail in References 108 and 109. The use of suppressed clock pulse duration modulation (SCPDM) provides for voice transmission without digitization. The processing gain for voice is on the order of 21 dB. Digital data can be accommodated at rates up to 2.4 kbits/s. This configuration provides privacy, multiple access and discrete addressing (up to 2048 user codes). Similar equipment was used for the maritime VHF tests described in Section 3.3.3.

The current prices of these modems*, in small quantities, could put them in the "high cost" category. However, the relatively advanced state of development and the basic design both suggest that there is significant potential for price reductions in a commercial market environment, should one develop.

5.3.4 SAMSARS

The SAMSARS (Satellite Aided Maritime Search and Rescue System) concept was briefly described in Section 3.2.4. Direct sequence techniques are applied to low-rate digital data in order to provide for multiple access and resistance to unintentional

* About $20,000, depending on the features required.

interference. The SAMSARS concept has been tested via satellite, although not from a floating buoy, as an operational deployment would require (59).

The total cost of SAMSARS will include the cost of special-purpose receivers at each of at least three existing satellite ground stations. The way in which the ground station costs will be passed on to users has not yet been determined. Indeed, no reliable estimates of the ground station costs has been developed, although the cost of each receiver (including commercial development, testing, and installation) is believed to be on the order of $250,000. There are similar uncertainties in the ground station costs for alternative narrowband satellite distress signaling techniques, although the ground station costs for the narrowband techniques should be lower, all other things being equal.

In 1977, the Intergovernmental Maritime Consultative Organization (IMCO) Subcommittee on Radiocommunications estimated the cost of various Emergency Position-Indicating Radio Beacons (EPIRBs) that would signal via geosynchronous maritime communication satellite (110). The estimate was $1500 for a 1.6 GHz spread spectrum EPIRB and $1275 for a 1.6 GHz narrowband EPIRB. Although the basis of these estimates is not clear, their ratio (1.18) would place the spread spectrum EPIRB in the "low cost" category. If the cost of the EPIRB alone is considered, this is probably a legitimate conclusion. However, on the basis of overall system costs, the SAMSARS concept can reasonably be placed in the "moderate cost" category.

5.3.5 Slow FH/FM

This concept, as it might apply to non-Government mobile radio applications, was described in detail in Section 4.1. It could be implemented as a relatively simple modification of existing transceiver designs.

Slow frequency hopping has been applied in a number of military designs, including SINCGARS-V (see Section 2.3.2). Target prices for the SINCGARS-V transceivers are comparable to the current price range for commercial VHF transceivers for private land mobile applications (111, 112).

5.3.6 Homodyne Detector Applications

As described in Section 4.2, this concept is also intended as a simple modification of existing designs. The current prices for commercial microwave intrusion sensors (in small quantities) range from several hundred dollars to a thousand dollars or more. In order to qualify for the "low-cost" category, the cost increment for the improved versions must be no more than 50 to 200 dollars. It seems quite reasonable to expect that this target could be attained, considering the small amount of circuitry that must be added.

5.3.7 Cost Trends

For much of the technology that is used in the implementation of spread spectrum systems, costs have been decreasing due to the use of new technology and concomitant increases in productivity. Particular examples are frequency synthesizers (now used in citizens band "walkie-talkies") microprocessors (used in a wide range of consumer applications) and SAW devices (now used in consumer-grade television receivers).

6. CONCLUSIONS

6.1 General Comments

Spread spectrum techniques cover a wide variety of potential Government and non-Government applications. These techniques have been in use long enough to produce some advocates and some opponents. In any particular application, arguments can be made for and against the use of spread spectrum. Advocates and opponents alike will be able to find information in this report to support their respective cases, since the purpose of this report is not to defend individual viewpoints -- whether general or specific -- but to present a reasonably balanced view of the potential benefits, costs, and risks of spread spectrum communications as they might be applied under the FCC's regulatory domain.

The preparation of this report covered a time span of roughly nine months. The amount of original work that can be performed in such an interval is obviously limited. In addition, common sense dictates that the length of the final report not be expanded to an unreasonable size by the inclusion of large quantities of material that are already available in open technical literature. For these reasons, it was necessary focus on limited subject areas, possibly to the detriment of others. A certain amount judgment is involved in this process, and the particular areas covered in this report are likely to interest some readers more than others.

Spread spectrum systems have characteristics that differ both qualitatively and quantitatively from the characteristics of conventional narrowband systems. The costs, risks, and benefits of particular applications can be intelligently assessed only on a case-by-case basis. Provision for such assessments already exist within the FCC's rulemaking process. Under this process, decisions are affected by evaluations performed by the FCC staff and by comments from users of the spectrum who have an interest in such decisions. The structure of the rulemaking process provides for the introduction of new technologies, but tends to discourage or inhibit reckless use of the spectrum under ill-conceived implementations.

Many potential spread spectrum applications are likely to be economically unattractive. This factor seems to have been ignored or minimized in previous studies of non-Government spread spectrum applications. Economically unattractive technologies are not likely to have significant constituencies among potential user groups and are thus unlikely to become the subject of FCC regulatory proceedings.

Other potential spread spectrum applications may be economically feasible, but may make poor use of the spectrum resources that they would require. The commission should then be prepared to determine, on a case-by-case basis, whether the benefits provided by such applications justify their inefficient use of spectrum resources. Many of the potential benefits of spread spectrum technology can be achieved with more spectrum-efficient narrowband technologies, but a cost tradeoff is likely to be involved. In some cases, it is possible that particular services, functions, benefits, or levels of performance can be provided by spread spectrum technology at a lower cost than with conventional narrowband technology. However, examples of such cases are not easy to find. In other cases, it is possible that particular services, function, benefits, or levels of performance can only be provided by spread spectrum or other wideband techniques.

In certain applications, spread spectrum techniques can make more efficient use of the spectrum than the usual implementations of narrowband techniques. Such applications tend to be easier to identify when the information bandwidth per user is low and the operating frequency is high. They also tend to result when constraints are placed on the achievable spectrum efficiency of the narrowband technique with which spread spectrum is being compared. Examples of such constraints include the need for guard bands, the existence of interference, and the use of specific modulation techniques. Comparisons of alternative modulation techniques or multiple access techniques must be carefully defined if the results as to be meaningful.

With these introductory comments, the following conclusions are submitted on the use of spread spectrum techniques for non-Government applications.

6.1.1 Performance and Efficiency of Spectrum Utilization

A tradeoff exists between communication performance (or radiolocation performance) and efficiency of spectrum utilization in virtually all spread spectrum systems. In the absence of interference and multipath, spread spectrum systems can achieve about the same levels of performance attainable with corresponding narrowband systems. However, such single-link spread spectrum systems make very inefficient use of the spectrum. As additional spread spectrum links are added in the same bandwidth, spectrum utilization improves, but higher signal energy is required to maintain a given performance level. The continued addition of user links will eventually increase mutual interference enough to prevent the attainability of any fixed level of performance, even with infinite signal energy. Very high levels of spectrum utilization can be achieved, but typically at very low levels of performance. This general concept is presented in quantitative form in Appendix A.

A second aspect of the tradeoff between performance and spectrum efficiency is that communication techniques yielding high performance in background noise alone exhibit higher levels of spectrum utilization (for a given performance level) than do less-efficient techniques, when applied in a multiple-access environment. Thus, digital CDMA systems using powerful forward error correction techniques can provide better spectrum efficiencies than corresponding systems with only simple coding.

In multiple access systems where all signals are received with equal average power levels, spread spectrum multiple access techniques suffer a theoretical disadvantage in their maximum attainable spectrum efficiency relative to CDMA or TDMA. The disadvantage can be severe when high levels of performance are required. This statement holds even if error correction coding or M-ary signaling are allowed. Again, Appendix A provides a quantitative background for this conclusion.

Common sense dictates that the utmost care be used in drawing conclusions about practical systems based on idealized theoretical models. In particular, the work presented in Appendix A may be directly applicable only to certain geosynchronous satellite links and similar applications where the received signals have relatively uniform power levels and the frequency tolerances (including Doppler shifts) are narrow. In such applications, spread spectrum multiple access techniques typically require more bandwidth and higher power per user than do TDMA or FDMA techniques.

In applications where received signal powers vary over a wide range, or where frequency tolerances are significant (relative to the information rate per user), the tradeoff between CDMA and FDMA must be viewed in a different light. Comparisons of this type must be formulated carefully if the results are to be meaningful. It is often (if not always) possible to produce comparative examples in which practical implementations of spread spectrum systems exhibit better spectrum utilization than the conventional frequency-channalized approach if the performance levels of the two systems being compared are not carefully defined.

A number of previous studies have compared the spectrum efficiency of spread spectrum with the spectrum efficiency of conventional FM systems for land mobile communication systems. In some of these comparisons, the spread spectrum approach appears to be equivalent to or somewhat superior to conventional FM systems on the basis of efficiency of spectrum utilization. In other comparisons, conventional FM techniques exhibit somewhat better performance. Such exercises call into question not only the voice quality for the approaches being compared but also the cost of the equipment, since the higher cost of spread spectrum equipment opens the door to other narrowband approaches (like amplitude-companded SSB) that may be more efficient in their use of the spectrum than narrowband FM.

One area in which spread spectrum techniques probably make better use of the spectrum than FDMA techniques is in the transmission of low-rate data at microwave frequencies when substantial frequency tolerances (due to oscillator tolerances or Doppler shifts, or both) are involved. This topic is explored in further detail in Appendix B. But again, economic factors are involved. In some of these cases TDMA may prove to be the most effective approach from the viewpoint of spectrum utilization, if its typically higher cost (with respect to FDMA) is acceptable. In the case of simple homodyne detector applications (see Section 4.2) spread spectrum techniques may be able to provide very high spectrum efficiencies at low cost.

Another area in which spread spectrum techniques may be able to improve the utilization of the spectrum is in cases where use can be made of ISM bands that are relatively unsuitable for applications requiring guaranteed high levels of performance. Indeed, since users of the ISM bands are not nominally protected from interfrence, it can be argued that any productive use of these bands frees other spectrum resources that are needed by applications requiring protection from interference.

6.2 Costs and Risks

6.2.1 Costs

Most spread spectrum systems can be viewed as conventional narrowband systems to which a higher level of complexity has been added for the purpose of achieving particular design characteristics. This added level of complexity naturally entails increased cost. In some cases, the cost increase may be relatively small. A preliminary assessment of the relative costs of various spread spectrum applications was provided in Section 5.3.

6.2.2 The Risk of Increased Interference

Current users of the spectrum are likely to be concerned about the risk of interference from new spread spectrum systems. This will be particularly true if attempts are made to "overlay" spread spectrum signals on bands that are now used for conventional signals in the same geographic area. Such "overlays" may be more acceptable in certain ISM bands than in bands where users are now protected from interference.

Even if "dedicated" bands are set aside for spread spectrum users, these users will suffer interference from one another. But the impact of such interference is minimized by the inherent interference resistance of the spread spectrum approach. The interference potential of any implementation must be assessed on the basis of the proposed system parameters.

6.2.3 Monitoring and Enforcement

Spread spectrum techniques can be used to produce systems in which each user has a unique code that underlies his transmitted signal. It can be difficult or impossible to demodulate a particular signal without knowledge of its "user code," and the number of possible user codes can be astronomical. If the FCC licenses such systems, it will face roughly the same type and degree of risk that it faces with conventional secure voice and secure data systems. Some possible measures for mitigating these risks are listed in Section 5.2.

6.3 Potential Applications

Although it is obviously not possible to predict specific applications that the FCC will be asked to license, it may be useful to identify the potential applications that have been proposed to date. Of these, potential land mobile applications using fast frequency hopping have dominated in the area subject to FCC licensing. Although a significant amount of analysis has been performed in this case, little or no experimental activity is evident, and implementations in the near-term seem unlikely, except possibly on an experimental basis. The prospects for the realization of such systems is weakened by their potentially high cost, although future technological breakthroughs could negate this factor.

In the maritime mobile area, direct sequence spread spectrum signaling is being proposed for emergency signaling via satellite. This concept has been partially field-tested, and could be implemented in the late 1980's. Direct sequence signaling has also been examined for possible use in terrestrial VHF maritime mobile applications, but the well-known vulnerability of direct sequence techniques to nonuniform signal levels (i.e., the "near-far" problem) will work against their realization in any mobile application that does not involve relay from geosynchronous satellites or dynamic control of transmitter powers.*

* However, hybrid techniques involving direct sequence techniques are entirely possible in such applications.

Aside from these better-known potential applications, the use of spread spectrum may be proposed in areas that have not previously received such attention. Near-term applications are likely to involve low-cost techniques like slow frequency hopping, the modification of homodyne sensors to provide a spread spectrum capability, chirp radar or chirp radiolocation systems.

6.4 Acknowledgements

The number of people who contributed to this study is so large that a complete list would be impractically long. Preliminary discussions on a possible study of non-Government uses of spread spectrum techniques were first held between Mr. Jack Robinson of the FCC's of the Plans and Policies and Dr. Feisal Keblawi of MITRE, after they met at a short course on the subject. A proposal was subsequently prepared and, after coordination with the FCC's Chief Scientist, Dr. Stephen Lukasik, a contract was awarded. Dr. Michael Marcus of the FCC's Office of Science and Technology was named as COTR.

Besides the individuals named above, the following people contributed to the study or provided useful cooperation in one form or another.

Yarsolav (Bob) Kaminsky	MITRE
Joe Fee	MITRE (currently with FAA)
Jim Morrell	MITRE
George Steelman	MITRE
Art Smith	MITRE
Mary Ann Davis	MITRE
Bill Zeiner	MITRE
Bob Noyer	MITRE
Richard Horn	MITRE
Paul Ebert	COMSAT Corporation
Aaron Weinberg	Stanford Telecommunications
Richard Tell	EPA/Las Vegas
George Cooper	Purdue University
Ray W. Nettleton	Michigan State University
Jerry Silverman	Magnavox
Charles Cahn	Magnavox
Paul Goodman	Bell Laboratories/Holmdel
Thijs de Haas	DoC, NTIA/Boulder
Carl S. Mathews	DoC, Maritime Administration
John Juroshek	DoC, NTIA/Boulder
Samuel Musa	DoD
Don J. Torrieri	DoD

7. REFERENCES

1. R. C. Dixon, Spread Spectrum Systems, Wiley, 1976.

2. Donald D. Neuman, "JTIDS--An Integrated Navigation and Identification System--and Its Potential for Air Traffic Control."

3. A. L. Covitt and D. D. Neuman, "Band Sharing--A Case Study," National Telecommunications Conference, Washington, DC, November 1979.

4. F. H. Dickson, "Packet Radio Communications," Collins Radio Co., AD-786155, 30 September 1974.

5. J. Inst. of Navigation, Vol. 25, No. 2, Summer 1978, Special Issue on GPS.

6. Final Report, Multipath/Modulation Study for the Tracking and Data Relay Satellite System, Contract No. NAS5-10744, The Magnavox Company, Government and Industrial Systems Division.

7. B. G. Glazer, "Spread Spectrum Concepts--A Tutorial," Proceedings of the 1973 Symposium on Spread Spectrum Communications," Volume 1, pp. 5-8, March 1973.

8. R. L. Harris, "Introduction to Spread Spectrum Techniques," in Spread Spectrum Communications, Advisory Group for Aerospace Research and Development, AD-766914, July 1973.

9. William F. Utlaut, "Spread Spectrum Principles and Possible Application to Spectrum Utilization and Allocation," IEEE Communications Society Magazine, September 1978.

10. Heinz H. Schreiber, "Self-Noise of Frequency-Hopping Signals," IEEE Trans. Comm. Tech., Vol. COM-17, No. 5, October 1969.

11. D. H. Hamsher, Communication System Engineering Handbook, Chapter 18, McGraw-Hill, 1967.

12. Edward G. Magill, et al., "Charge-Coupled Device Pseudo-Noise Matched Filter Design," Proc. IEEE, Vol. 67, No. 1, January 1979.

13. Ronald D. Haggarty, et al., "SAW Processors vs. Competing Technologies for Spread Spectrum Data Link Applications," IEEE Ultrasonics Symposium, 26 December 1978.

14. Carlo H. Sequin and Michael F. Tompsett, Charge Transfer Devices, Academic Press, 1975.

15. Donald M. Grieco, "Inherent Signal-to-Noise Ratio Limitations of Charge-Coupled Device Pseudo-Noise Matched Filters," IEEE Trans. Comm., Vol. COM-28, No. 5, May 1980.

16. Andrew J. Viterbi, "Spread Spectrum Communications--Myths and Realities," IEEE Communications Magazine, May 1979.

17. Andrew R. Cohen, et al., "A New Coding Technique for Asynchronous Multiple Access Communication," IEEE Trans. Comm. Tech., Vol. COM-19, No. 5, October 1971.

18. J. P. Costas, "Poisson, Shannon, and the Radio Amateur," Proc. IRE, December 1959.

19. George R. Cooper and Ray W. Nettleton, "Spectral Efficiency in Cellular Land-Mobile Communications--A Spread-Spectrum Approach," PB293264, 1978.

20. G. R. Cooper and R. W. Nettleton, "Spread Spectrum in a Fading, Interference-Limited Environment," International Communications Conference, June 10-13, 1979.

21. D. P. Grybos, et al., "Probability of Error Performance of the Spread Spectrum Mobile Communications Receiver in a Non-Rayleigh Fading Environment," 29th IEEE Vehicular Technology Conference, March 27-30, 1979.

22. G. R. Cooper, et al., "Cellular Land Mobile Radio: Why Spread Spectrum," IEEE Communications Magazine, Vol. 17, No. 2, March 1979.

23. G. R. Cooper and R. W. Nettleton, "A Spread-Spectrum Technique for High-Capacity Mobile Communications," IEEE Trans. Vehicular Technology, Vol. VT-27, No. 4, November 1978.

24. D. P. Grybos and G. R. Cooper, "A Receiver Feasibility Study for the Spread Spectrum High Capacity Mobile Radio Systems," 28th IEEE Vehicular Technology Conference, March 22-24, 1978.

25. Charles Cahn (Magnavox), Private Communication, December 10, 1979.

26. Jerry Silverman (Magnavox), Private Communication, December 10, 1979.

27. Del Norte Technology, Inc., Comments on FCC Notice of Proposed Rulemaking PR Docket 80-9, "Police Surveillance."

28. IEEE Trans. Comm., Vol. COM-25, No. 8, August 1977 (Special Issue on Spread Spectrum).

29. Samuel A. Musa and Wasyl Wasylkiwskyj, "Co-Channel Interference of Spread Spectrum Systems in a Multiple User Environment," IEEE Trans. Comm., Vol. COM-26, No. 10, October 1978.

30. Leslie A. Berry, "Spectrum Metrics and Spectrum Efficiency: Proposed Definitions," IEEE Trans. Electromagnetic Compat., Vol. EMC-19, No. 3, August 1977.

31. R. P. Gifford, "EMC Revisited--1966," IEEE Trans. Electromagnetic Compat., Vol. EMC-8, No. 3, September 1966.

32. K. Powers, "Diversity of Broadcasting," Science and Technology, April 1968.

33. L. C. Tillotson, et al. "Efficient Use of the Radio Spectrum and Bandwidth Expansion," Proc. IEEE, Vol. 61, No. 4, April 1973.

34. D. R. Ewing and L. A. Berry, "Metrics for Spectrum-Space Usage," Office of Telecommunications Report 73-24, U.S. Department of Commerce, Nov. 1973.

35. O. L. Luk'yanova, "Characteristics of the Utilization of the Radio-Frequency Spectrum by a Particular System," Telecommunications and Radio Engineering, Vol. 28/29, No. 4, April 1974.

36. C. Colavito, "On the Efficiency of the Radio Frequency Spectrum Utilization in Fixed and Mobile Communication Systems," Alta Frequenza, Vol. XLIII, No. 9, September 1974.

37. "The Effect of Modulation Characteristics on the Efficiency of Use of the Geostationary Orbit," Documents of the XIIIth Plenary Assembly, ITV, CCIR Report 559, Geneva, 1974.

38. D. N. Hatfield, "Measures of Spectral Efficiency in Land Mobile Radio," IEEE Vehicular Technology Conference, 1975.

39. Scott H. Cameron, "Efficiency of Spread Spectrum Signaling with Respect to Spectrum Utilization," 19th IEEE Electromagnetic Compatibility Symposium, August 1977.

40. "Mobile Multiple Access Study, Final Report," TRW Defense and Space Systems Group, NASA Contract No. NAS5-23454, August 1977.

41. R. C. Sommer, "Asynchronously Multiplexed Channel Capacity," Proc. IEEE, Vol. 54, January 1966.

42. Lucky, Salz, and Weldon, Principles of Data Communication, Ch. 8 and 9, McGraw-Hill, 1968.

43. Frank de Jager and Cornelis B. Dekker, "Tamed Frequency Modulation, A Novel Method to Achieve Spectrum Efficiency in Digital Transmission," IEEE Trans. Comm., Vol. COM-26, No. 5, May 1978.

44. Michael B. Pursley, "Performance Evaluation for Phase-Coded Spread-Spectrum Multiple-Access Communication--Part 1: System Analysis," IEEE Trans. Comm., Vol. COM-25, No. 8, August 1977.

45. Leslie A. Berry and E. J. Haakinson, "Spectrum Efficiency for Multiple Independent Spread Spectrum Land Mobile Radio Systems," NTIA Report 78/11, PB-291539, November 1978.

46. Anita G. Longley, "Location Variability of Transmission Loss--Land Mobile and Broadcast Systems," Office of Telecommunications, Report No. 76-87, May 1976.

47. A. G. Longley and P. L. Rice," Prediction of Tropospheric Radio Transmission Loss Over Irregular Terrain, A Computer Method--1968," ESSA Technical Report ERL 79-ITS67, AD-676874, July 1968.

48. A. G. Longley and P. L. Rice, "Comparison of Propagation Measurements with Predicted Values in the 20 to 10,000 MHz Range," ESSA Technical Report ERL 148-ITS 97, January 1970.

49. Anita G. Longley, "Radio Propagation in Urban Areas," 28th IEEE Vehicular Technology Conference, March 22-24, 1978.

50. George R. Cooper and Raymond W. Nettleton, "A Spread Spectrum Technique for High Capacity Mobile Communications," 27th IEEE Vehicular Technology Conference, March 16-18, 1977.

51. Raymond W. Nettleton and George R. Cooper, "Error Performance of a Spread-Spectrum Mobile Communications System in a Rapidly-Fading Environment," National Telecommunications Conference, 1977.

52. Raymond W. Nettleton and George R. Cooper, "Mutual Interference in Cellular LMR Systems: Narrowband and Broadband Techniques Compared," Midcon 1977 Proceedings, November 8-10, 1977.

53. Gary D. Ott, "Vehicle Location in Cellular Mobile Radio Systems," IEEE Trans. Veh. Technol., Vol. VT-26, No. 1, February 1977.

54. James J. Mikulsky, "Mobile Radio: To Spread or Not to Spread," IEEE Communications Magazine, Vol. 17, No. 4, July 1979.

55. Paul S. Henry, "Spectrum Efficiency of a Frequency-Hopped DPSK Mobile Radio System," 29th IEEE Vehicular Technology Conference, 1979.

56. A. Weinberg, et al., "A Novel Concept for a Satellite-Based Maritime Search and Rescue System," IEEE Trans. Veh. Technol., Vol. VT-26, No. 3 (Special Joint Issue on Maritime Communications), August 1977.

57. A. Weinberg, "Characteristics and Performance Aspects of a Concept for Satellite Based Maritime Search and Rescue," The MITRE Corporation, Report No. M77-101, September 1977.

58. W. Scales, "Development of an Experimental Spread Spectrum Maritime Communications System: Phase 1," The MITRE Corporation, Report No. MTR-79W00217, July 1979.

59. J. J. Fee, et al., "Development of an Experimental Spread Spectrum Maritime Communications System: Phase II," The MITRE Corporation, Report No. MTR-80W00052, November 1980.

60. "Operational Concept, Technical Characteristics and Test Results of the Satellite-Aided Maritime Search and Rescue System (SAMSARS)," U.S. CCIR Document USSG 8/12C, 21 May 1980.

61. SARSAT System Summary, April 1980.

62. "European Communications Experiments in L Band with ATS-6," European Space Research Organization, Volumes 6 and 11.

63. Larry D. Reed, "Land Mobile Spectrum Utilization, Los Angeles and San Diego, CA," FCC Report No. PRB/RDL 79-02, November 1979.

64. "Spectrum Requirements for the Civil Maritime and Aeronautical Services," Report to the FCC by the UHF Task Force, Appendix A, FCC Order No. 1820-2 (30E25.13), November 1978.

65. Paul Newhouse, "Procedures for Analyzing Interference Caused by Spread-Spectrum Signals," ITT Research Institute Report No. ESD-TR-77-033, AD A056911, February 1978.

66. Don J. Torrieri, "Frequency-Hopping in a Jamming Environment," AD A067299, December 1978.

67. "Japanese EPIRB Experiment Plan," CCIR XVth Plenary Assembly, Japan No. 8/7/5, June 1980.

68. George R. Cooper, "Multi-Service Aspects of Spread-Spectrum Mobile Communication Systems," MIDCON '78, Paper No. 28.2, Dallas, Tx, December 12-14, 1978.

69. T. Dvorak, "Compatibility of Spread-Spectrum Signals with Narrow-band-FM Receivers in VHF Mobile Networks," IEEE Electromagnetic Compatibility Symposium, p. 19-23, June 20-27, 1978.

70. J. R. Juroshek, "A Compatibility Analysis of Spread-Spectrum and FM Land Mobile Radio Systems," NTIA Report 79-23, August 1979.

71. John R. Juroshek, "A Preliminary Estimate of the Effects of Spread-Spectrum Interference on TV," NTIA Report 76-8, PB-286623, June 1978.

72. R. F. Ormondroyd, "Spread Spectrum Communication in the Land Mobile Service," IEE Conference on Communication Equipment and Systems, Birmingham, England, April 4-7, 1978.

73. D. C. Coll and N. Zervos, "The Use of Spread Spectrum Modulation for the Co-Channel Transmission of Data and Television," National Telecommunications Conference, December 27-29, 1979.

74. "EMC Analysis of JTIDS in the 960-1215 MHz Band," Report No. OT 78-140, March 1978.

75. L. R. Welch, "Lower Bounds on the Minimum Cross Correlation of Signals," IEEE Trans. Inf. Theory, May 1974.

76. Dilip V. Sarwate, "Bounds on Cross Correlation and Autocorrelation of Sequences," IEEE Trans. Inf. Theory, Vol. IT-25, No. 6, November 1979.

77. R. D. Yates and G. R. Cooper, "Design of Large Signal Sets with Good Aperiodic Correlation Properties," Technical Report TR-EE 66-13, Purdue University, September 1966.

78. IEEE Proceedings, Vol. 62, No. 1, (Special Issue on Industrial, Scientific, and Medical Applications of Microwaves), January 1974.

79. Richard A. Tell, "Field-Strength Measurements of Microwave-Oven Leakage aat 915 MHz," IEEE Trans. Electromagnetic Compat., Vol. EMC-20, No. 2, May 1978.

80. J. M. Durante, "Building Penetration Loss at 900 MHz," 1973 IEEE Vehicular Technology Conference, Cleveland, Ohio, December 3-5, 1973.

81. M. I. Skolnik, Radar Handbook, McGraw-Hill, p. 24-17, 1970.

82. Del Norte Technology, Inc., Petition for Rulemaking, General Docket 80-135 (non-Government radiolocation in the 420-450 MHz band).

83. A. M. McCalmont, "Communications Security Devices (CSD)--Techniques, Constraints, and Selection," 28th IEEE Vehicular Technology Conference, Denver, Colorado, March 22-24, 1978.

84. Thijs De Haas, et al., "Maritime Digital Selective Calling System Technical Description," The MITRE Corporation, Report No. MTR-7217, Rev. 1, June 1976.

85. William C. Jakes, Jr., (Ed.), Microwave Mobile Communications, Wiley, 1974.

86. W. Scales, "Urban Field Tests of Four Vehicle Location Techniques," The MITRE Corporation, Report No. MTR-6399, Rev. 1, April 1973.

87. H. Helgert and R. L. Pickholtz, "Low Rate Coding for Spread Spectrum," National Telecommunications Conference, Washington, DC, November 27-29, 1979.

88. Bruce P. Gibbs, "Results of the NASA/MARAD L-Band Satellite Navigation Experiment," EASCON-77 (also reprinted in AESS Newsletter, July 1978).

89. Ray W. Nettleton, "Traffic Loads in a Spread-Spectrum Cellular Land-Mobile System," 1979 MIDCON, November 6-8, 1979.

90. Ray W. Nettleton, "Traffic Theory and Interference Management for a Spread Spectrum Cellular Mobile Radio System," International Conference on Communications, Seattle, June 8-12, 1980.

91. Donald H. Hamsher (Ed.), Communication System Engineering Handbook, P. 17-14, McGraw-Hill, 1967.

92. Merril I. Skolnik, Radar Handbook, p. 5-10 and 5-11, McGraw-Hill, 1970.

93. Edward N. Skomal, "The Range and Frequency Dependence of VHF-UHF Man-Made Radio Noise in and Above Metropolitan Areas," IEEE Trans. Vehicular Technol., Vol. VT-19, No. 2, May 1970.

94. Reference Data for Radio Engineers, Fourth edition, p. 763, ITT, 1967.

95. Michael J. Gans, "A Power-Spectral Theory of Propagation in the Mobile-Radio Environment," IEEE Trans. Vehicular Technol., Vol. VT-21, No. 1, February 1972.

96. George A. Miller and J. C. R. Licklider, "The Intelligibility of Interrupted Speech," J. Acoustical Soc. of America, Vol. 22, No. 2, March 1950.

97. "Minimum Standard for Land-Mobile Communication Systems Using FM or PM in the 25-470 MC Frequency Spectrum," EIA Std. RS-237, August 1960.

98. R. P. Yates, "Design of Large Signal Sets with Good Aperiodic Correlation Properties," Ph.D. Thesis, Purdue University, 1966.

99. Goran Einarsson, "Address Assignment for a Time-Frequency-Coded Spread Spectrum System," Bell Laboratories, Crawford Hill Laboratory, Holmdel, NJ, 07733.

100. James L. Flanagan, et al., "Speech Coding," IEEE Trans. Comm., Vol. COM-27, No. 4, April 1979.

101. A. J. Viterbi, "A Processing Satellite Transponder for Multiple Access by Low-Rate Users," Digital Satellite Communications Conference, Montreal, October 1978.

102. David J. Goodman, et al., "Frequency Hopped Multilevel FSK for Mobile Radio," 1980 International Zurich Seminar on Digital Communications.

103. Uzi Timor, "Improved Decoding Scheme for Frequency Hopped Multilevel FSK System," Bell Laboratories, Crawford Hill Laboratory, Holmdel, NJ.

104. Goran Einarsson, "Address Assignment for a Time-Frequency-Coded Spread Spectrum System," Bell Laboratories, Crawford Hill Laboratory, Holmdel, NJ.

105. B. G. Haskell, "Computer Simulation Results on Frequency Hopped MFSK Mobile Radio--Noiseless Case," Bell Laboratories, Crawford Hill Laboratory, Holmdel, NJ.

106. "Selection and Application Guide to Commercial Intrusion Alarm Systems," NBS Special Publication 480-14, August 1979.

107. U. A. Frank, et al., "The Two-Pound Radar," RCA Eng., Vol. 13, No. 2, pp. 52-54, August-September 1967.

108. Leonard Jacobson, et al., "Highly Efficient Voice Modulation for Low C/N_0 Communications Channels with Hard Limiting Repeaters," IEEE Trans. Comm., Vol. COM-21, No. 2, February 1973.

109. "MX-170C for Satellite and Radio Communications," Report No. R-5362, Magnavox, Advanced Products Division, Torrance, CA, May 1976.

110. "Operational Requirements for EPIRBs and Survival Craft Equipment," COMXVIII/WP.8, IMCO Subcommittee on Radiocommunications, 1977.

111. R. D. Boyle, "SINCGARS V--Secure Battlefield Communications for the 1980s," International Defense Review, pp. 1155-1157, 6/1977.

112. "JTIDS/TIES Consolidate Tactical Communications," EW Review, September/October 1977.

113. "More Complaints Received About Speed-Radar Jammers," Signal, p. 52, July 1980.

114. William E. Gordon and Lewis M. Duncan, "SPS Impacts on the Upper Atmosphere," Aeronautics and Astronautics, July/August 1980.

APPENDIX A:

SPECTRUM EFFICIENCY OF SSMA

Summary

The spectrum efficiencies of several spread spectrum multiple access (SSMA) techniques are examined. Spectrum efficiency, for the purpose of this paper, is defined as total network information throughput per unit bandwidth. Results are presented for fading and non-fading channels. Under conditions of equal received power levels, direct sequence techniques can out perform frequency-hopping techniques. This conclusion is reversed when one of the interferers is much more powerful than the desired signal. The spectrum efficiencies attainable with SSMA are generally low, although SSMA may provide better spectrum efficiency than FDMA or TDMA under certain specialized conditions.

A.1 INTRODUCTION

The spectrum efficiency of particular Spread Spectrum Multiple Access (SSMA) networks has been the subject of a number of recent papers. For example, Cooper and Nettleton have analyzed the performance and spectrum efficiency of a unique frequency-hopping system using DPSK (A1-A5). Henry (A6) has performed an independent analysis of the same basic approach (although with different parameters) that includes spectrum efficiency computations and comparisons. Frequency hopping with multilevel FSK was analyzed by Goodman and others (A7).

At a more general level, numerous analyses of SSMA performance have been published, but usually without reference to spectrum efficiency (A8-A13). Two notable exceptions are Costas (A14) and Cameron (A15). In some cases, it may not be possible to derive a simple, closed-form expression for SSMA spectrum efficiency in terms of the network performance parameters. In others, such a relationship follows indirectly from the results of the performance analysis.

Spread spectrum systems are of several basic types. In direct sequence (DS) systems, each information symbol is encoded as a pseudo-random sequence that is superimposed on the carrier phase. In frequency-hopping (FH) systems, the transmit and receiver carrier frequencies are switched at regular intervals according to a pseudo-random pattern. Each method has advantages and disadvantages, independent of the multiple access application. Other relative advantages and disadvantages relate directly to multiple access capabilities.

The purpose of this appendix is to relate SSMA communication performance (in terms of error rate) to spectrum efficiency. It can be argued that such relations should exist for any SSMA system. First, let spectrum efficiency be initially defined as

$$\eta = \frac{MR}{B} = \frac{Mq}{BT}$$

where

- M = number of simultaneous SSMA links
- R = information rate per link
- B = total system r.f. bandwidth
- q = information content of one symbol (bits)
- T = symbol duration

Although other measures of spectrum efficiency have been suggested (A16), this one has been used frequently (A5,A6). By this definition, η is approximately the network information throughput per unit bandwidth for systems with low error rates. (The definition will later be modified to reflect the loss of information due to transmission errors.)

As an elementary observation, it is clear that if M exceeds the number of the dimensions in the signal space, 2BT, then there will necessarily be mutual interference between the various links, even if all of the links have the same symbol timing and the relative carrier phases can be perfectly controlled. Furthermore, as $M/2BT = \eta/2q$ increases, the average distance between members of any signal set in the space defined by 2BT dimensions must decrease. Thus, a relationship between spectrum efficiency (as defined above) and communication performance would be expected in every case.

As a first step in established relations between performance and spectrum efficiency, consider the general case in which each of M distinct DS signals occupies an r.f. bandwidth equal to twice the one-sided bandwidth, W, of its baseband code. In the absence of phase synchronization between the M carriers, the effective dimensionality of the signal space is then 2WT = BT. Assume further that the M-1 unwanted signals can be represented as a composite waveform characterized as an independent noise process, x(t). Then, for equal signal energies, the one-sided power density of x(t) is approximately

$$S_o = \frac{(M-1)E}{BT} \tag{1}$$

where E is the energy of a transmitted symbol. The effective total noise power density is

$$N'_o = N_o + S_o \tag{2}$$

where N_o is the background noise power density. Solving equation (2) for S_o and substituting into equation (1) yields

$$N'_o - N_o = \frac{(M-1)E}{BT}$$

which implies

$$\eta = q\left[\frac{N'_o}{E} - \frac{N_o}{E} + \frac{1}{BT}\right] \tag{3}$$

The required error rate, for any given data modulation technique, determines the required E/N_0'. Also, E/N_0 is just the signal-to-noise ratio without interference, so both of these parameters are presumably known. Thus, equation (3) relates spectrum efficiency to communication performance through E/N_0'. Note that for large WT, η can approach one only if E/N_0 is large.

As Mazo recently pointed out (A8), this model must be used carefully, because it embodys the tacit assumption that the M codes are more or less uniformly distributed over the available signal space. In particular, if the M codes in use at any given time are only a subset of a larger number, L, of assigned user codes, then the error rate computed from E/N_0 (whether using a Gaussain noise model or not) will not be a guaranteed error rate for any M users, but rather an average error rate over the ensemble of all possible subsets of M codes out of the total population L.

Note also that, while equation (3) implies that η is a discrete function of the variables on the right, η can, in fact, only take on discrete values, corresponding to particular values of M and N_0'. However, it will later become evident that, for large M, the values of η are so close together that equation (3) can, for some purposes, be considered to be continuous.

There are several other noteworthy aspects to equation (3). Since it contains a term inversely proportional to BT, a disadvantage is implied for spread spectrum systems. For narrowband systems, $BT \approx 2$ ($WT \approx 1$) and $\eta = 0.5$ for binary signals in the absence of interference. (Note that η can exceed unity if M-ary narrowband signals are used.) However, narrowband systems are not resistant to co-channel interference and must generally be multiplexed by frequency. The requirement for guard bands limits the spectrum efficiency of Frequency Division Multiple Access (FDMA) to values typically smaller than unity, although FDMA may, in many cases, offer spectrum efficiency superior to that of SSMA. For large SSMA systems, BT must be large also, and in this case η is seen to be relatively independent of BT. Also, the use of error-correcting codes to increase BT can substantially lower the required E/N_0' for a given error rate, thus yielding a net increase in spectrum efficiency.

A.2 DIRECT SEQUENCE

A.2.1 Synchronous Direct Sequence

Since it is usually impractical to synchronize transmitters at to within one code "chip," (spread spectrum code element) this case is primarily of theoretical interest. It provides an upper bound on uncoded binary systems, and was treated by Mazo (A8) in a recent analysis.

The case to be considered here involves the use of M binary phase-shift keyed (PSK) signals of the form

$$x_i(t) = b_i c_i(t)\cos(\omega_c t + \phi_i) \qquad 0 \leq t \leq T$$

where $b_i = \pm 1$ and all variables are real. The code sequence is

$$c_i(t) = \sum_{j=0}^{K-1} a_j S(t-j\Delta)$$

where again $a_j = \pm 1$. Also, s(t) is the basic "chip" waveform of length Δ. The carrier phase of the desired signal is taken to be zero; the phases of the other signals are assumed to be uniformly distributed over $(0,2\pi)$. The effective dimensionality of the baseband signal space is $K = T/\Delta$. Note that all signals are assumed to be mutually synchronized except for carrier phase.

A.2.1.1 Noise-Limited Case

First, let us examine the case where the maximum number of simultaneous links, M, is equal to the number of codes that have been assigned. That is, all possible users are allowed to access the network simultaneously. Clearly, if $M \leq K$, the M codes can be chosen to be orthogonal, so interference does not result. The maximum spectrum efficiency is then

$$\eta = \frac{M}{2WT} = \frac{M}{2K} \leq 0.5$$

where the equality holds for the maximum loading without interference, i.e., M = K.

A.2.1.2 Interference-Limited Case, Restricted Access

In this case, the condition $K < M$ requires explicit consideration of both noise and interference. The M codes in use at any given time are taken to be a subset of a total population of L user codes. For example, there may be L = 1000 potential users, of which only M = 500 are allowed simultaneous access.

The receiver correlates the incoming signal against $c_i \cos(\omega_c t)$ (where ϕ_1 has been taken to be zero), and uses the sign of the result as an estimate of b_i. The test statistic against which the bit decisions are made is proportional to

$$y_1 = b_1 + \sum_{i=2}^{M} b_i \rho_{1i} \cos\phi_i + n_1 \tag{4}$$

where ρ_{ij} is the normalized cross-correlation

$$\rho_{ij} = \frac{1}{E}\int_0^T c_i(t)c_j(t)dt$$

and $n_1 = \int_0^T n(t)c_1(t)\cos\omega_c t dt$ is the noise component at the correlator output. The predetection SNR (power ratio) is

$$\frac{2E}{N_o'} = \frac{2}{\sum_{i=2}^{M} \rho_{1i}^2 + N_o/E} \tag{5}$$

Because of the presence of ρ_{1i} in equations (4) and (5), the performance depends on the particular set of codes in use, as well as the number of links in use. However, for any intelligently-selected large set of codes, there are bounds on ρ_{ij}^2. As a lower bound, the Welch bound on inner products (A19) requires that

$$\rho_{max}^2 > \frac{1}{L-1}\left(\frac{L}{K} - 1\right)$$

where ρ_{max}^2 is the maximum ρ_{ij}^2 for a set of L codes, each consisting of K binary elements.

The Welch bound does not provide a direct constraint on the predetection SNR, except for the very weak bound

$$\frac{2E}{N_o'} \leq \frac{2}{\rho_{max}^2 + N_o/E}$$

and even this holds only for the code subsets containing ρ_{max}. It is therefore useful to consider the ensemble of all subsets of M codes out of the population L. The ensemble average of ρ_{ij}^2 will then be close to the arithemetic average of ρ_{ij}^2 for a given set of codes as long as M is sufficiently large. The "typical" predetection SNR is then

$$\frac{2E}{N_o'} \approx \frac{2}{(M-1)\overline{\rho^2} + N_o/E} \tag{6}$$

where $\overline{\rho^2}$ is the ensemble average of ρ_{ij}^2.

The spectrum efficiency is then

$$\eta = \frac{M}{2K} \approx \frac{1}{2K}\left(\frac{\beta}{\overline{\rho^2}} + 1\right)$$

where the notation $\beta = N_o'/E - N_o/E$ has been introduced for brevity.

It happens that the Welch bound on inner products (A19) was derived by placing a bound on $\Sigma\rho_{ij}^2$, where the summation is taken over all i,j. The result is that $\overline{\rho^2}$ is subject to the same bound as ρ_{max}^2. The impact of this on spectrum efficiency is that

$$\eta \leq \frac{1}{2}\left[\frac{\beta(L-1)}{L-K} + \frac{1}{K}\right]$$

It will now be shown that this bound can be achieved, at least for integer values of L/K. Consider the following strategy for assigning M codes out of a population L. Let the first K codes be orthogonal ($\rho_{ij} = 0$). These same codes are then uniformly re-assigned for the remaining L-K codewords. Because of the re-use of codewords, $\rho_{max}^2 = 1$. Also,

$$\overline{\rho^2} = 0 \cdot \Pr\text{ (2 randomly selected codewords are different)}$$

$$+ 1 \cdot \Pr\text{ (2 randomly selected codewords are identical)}$$

If we consider only values of L that are integer multiples of K, then the number of times that each of the K codewords is assigned is L/K. Then it is easily shown that

$$\overline{\rho 2} = K\left[\frac{\frac{L}{K}\left(\frac{L}{K} - 1\right)}{L(L-1)}\right] = \frac{1}{L-1}\left(\frac{L}{K} - 1\right) \qquad K < L$$

which is precisely the lower bound on $\overline{\rho^2}$ that was derived above. Since the bound is clearly attainable, at least in principle, the achievable spectrum efficiency is

$$\eta = \frac{1}{2}\left[\frac{\beta(L-1)}{L-K} + \frac{1}{K}\right] \approx \frac{\beta}{2}\left(\frac{L/K}{L/K-1}\right) \qquad 1 << K << L$$

This expression seems to suggest that the spectrum efficiency can be increased to any desired value by operating at low signal-to-noise ratios (which correspond to high values of β).

The problem is that in defining η, a low error rate has been assumed. At sufficiently low SNRs, η is high, but the amount of information being conveyed is small. This deficiency can be corrected by defining

$$\eta' = \eta(1 - H_L) \tag{7}$$

where H_L is the average loss of information per transmitted bit. That is, for binary signaling,

$$\begin{aligned} H_L &= -P_e \log_2 P_e - P_c \log_2 P_c \\ &= -P_e \log_2 P_e - (1-P_e)\log_2(1-P_e) \end{aligned} \tag{8}$$

This modification does not account for the impact of error correction coding (if used) or repeated transmissions in an ARQ system, but it does allow a more accurate information-theoretic view of spectrum utilization.

The bit error probability, P_e, can be easily computed if the interference term $\Sigma b_i \rho_{1i} \cos\phi_i$ in equation (4) is taken to be a normally distributed random variable, as suggested by the central limit theorem and the condition of equal received power levels. Then

$$P_e = Q\left(\sqrt{\frac{2E}{N_o'}}\right)$$

where $Q(x) = \frac{1}{\sqrt{2\pi}} \int_x^\infty e^{-t^2/2} dt$

The normal approximation to the bit error rate can be used to plot spectrum efficiency (η') in terms of error probability. The curves $\eta'(P_e)$ are shown in Figure A1 for representative values of L/K and E/N_o. It can be seen that the spectrum efficiency rises rapidly as the error rate is allowed to rise above 10^{-3}. The curves extend only to bit error rates of 0.10, since this approaches the useful limit for practical systems. However, the curves continue to rise to a maximum theoretical efficiency at error rates of about 0.4. At such high error rates, the upper family of curves (L/K = 2) peaks at about 0.88, while the lower family of curves (L/K = ∞) peaks at half that value. It is reasonable to question whether such high efficiencies could ever be attained in a practical system because the overhead associated with error detection and correction is not included in η'.

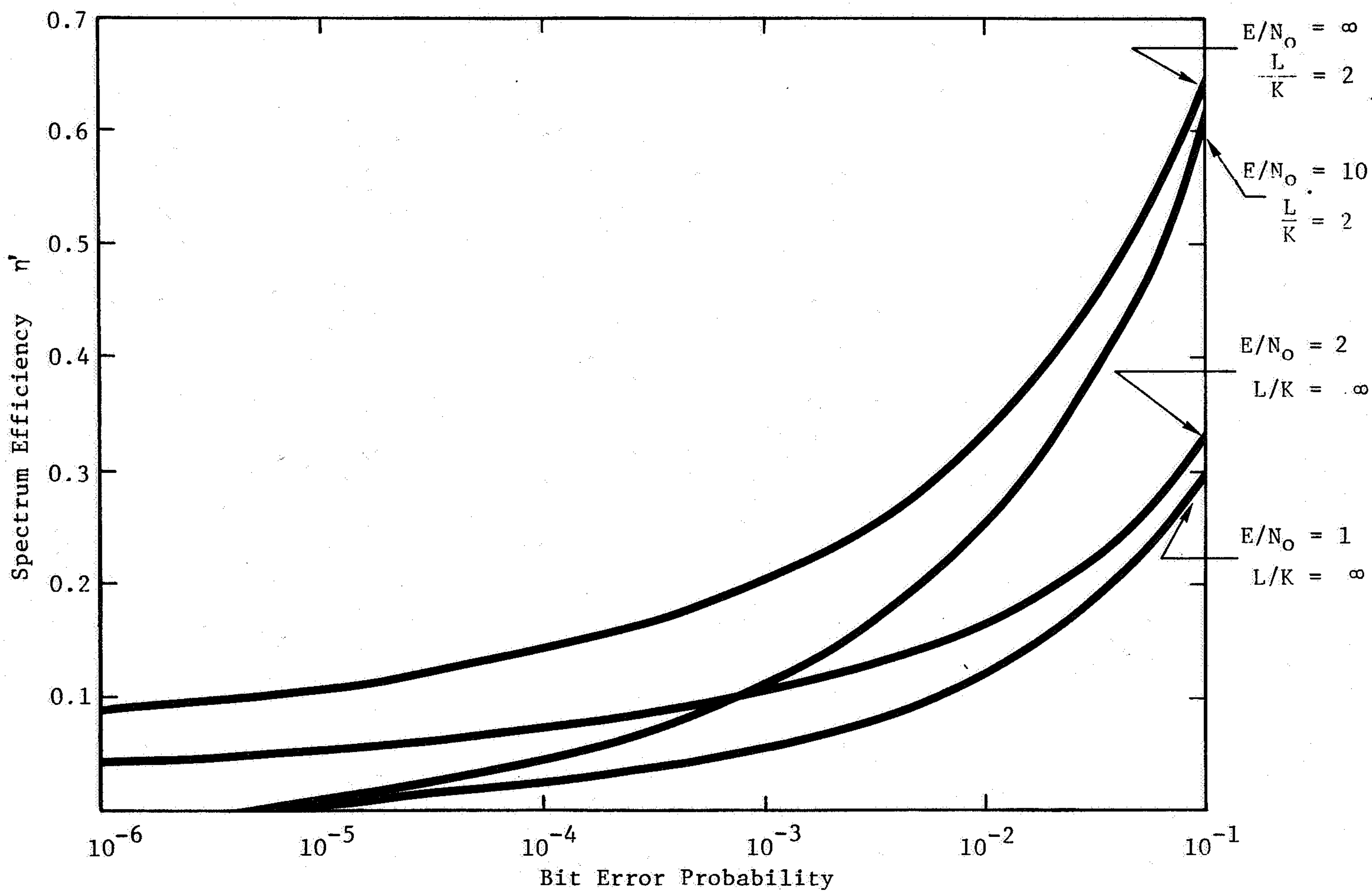

FIGURE A1
SPECTRUM EFFICIENCY OF SYNCHRONOUS DIRECT SEQUENCE SSMA: NUMBER OF ASSIGNED CODES GREATER THAN NUMBER OF SIMULTANEOUS USERS

In using Figure A1, it is important to keep in mind that P_e is computed using the mean squared cross correlation for the code population L. It is not a guaranteed error rate for any particular subset of M codes, but rather an indication of performance for an "average" set of M codes.

A.2.1.3 Interference-Limited Case, Unrestricted Access

In some cases, it may not be possible to limit access to the network. For this reason, it is interesting to examine the results when all assigned user codes are simultaneously active, i.e., when L = M. Using the arguments put forth in the previous section, the mean squared cross correlation between codes is subject to the bound

$$\overline{\rho^2} \geq \frac{1}{M-1}\left(\frac{M}{K} - 1\right)$$

where, as shown earlier, the equality is attainable (for integer values of M/K, at least) using simple (though not necessarily practical) code assignment strategies. Using equation (6),

$$\frac{2E}{N_o'} \approx \frac{2}{\left(\frac{M}{K} - 1\right) + N_o/E}$$

From which

$$\eta = \frac{M}{2K} = (\beta+1)/2$$

where, again,

$$\beta = N_o'/E - N_o/E$$

As before, $\eta' = \eta(1-H_L)$ can be computed based on a normal approximation to the interference-plus-noise. This is shown in Figure A2 as a function of the bit error rate for $E/N_o = 10$ and $E/N_o = \infty$.

It is interesting to note that η approaches 0.5 from above as β approaches zero. The condition $\beta = 0$ corresponds to an absence of interference, and this is a condition that can occur only for $M \leq K$. As noted under the "noise-limited case," the spectrum efficiency for M = K is precisely 0.5. For the "restricted access case," M < L. Under these conditions, a lack of interference cannot be guaranteed for K = M, since there is a good chance that the M codes selected at random out of the

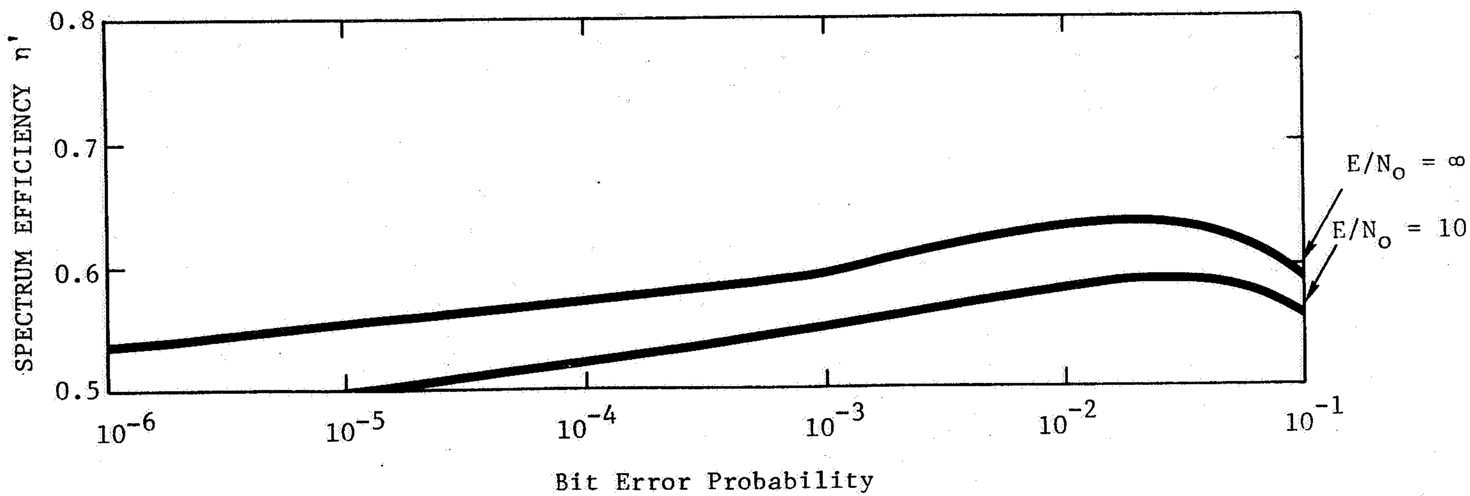

FIGURE A2
SPECTRUM EFFICIENCY OF SYNCHRONOUS DIRECT SEQUENCE SSMA: NUMBER OF CODES = NUMBER OF USERS

population L will not be orthogonal. Clearly, increasing the number of potential users, L, over the maximum number of simultaneous users, M, is not always beneficial.

A second interesting feature of Figure A2 is that, unlike Figure 1, the maxima occur within a potentially practical range of error rates. Specifically, the maxima occur for error rates of about 0.02.

A.2.2 Asynchronous Direct Sequence

In most cases, it is not practical to establish precise timing between all terminals in the network. Indeed, one of the advantages of SSMA with respect to time division multiple access (TDMA) is that such network synchronization is not required for SSMA.

As before, consider the case where M users are active out of a potential user population L. However, in this case, any interfering code is displaced from the desired code by a random time delay. In general, neither bit transitions nor the chip transitions of the various codes are alligned. The impact of the non-aligned bit transitions will be examined first.

There are K-1 ways in which the bit transition of a particular interfering signal can fall within the interval $0 \leq t \leq T$, assuming the chip transitions are aligned. Furthermore, there are 8 possible values of correlator output for each of these positions, depending on the signs of the two interfering bit segments and the "desired" bit. Thus, for two codes, the number of possible cross correlations is (K-1)8 when the bit transitions are not simultaneous. If the bit transitions are simultaneous, then there are just 4 possible cross correlations, since only one interfering bit occurs during the "desired" bit. The total number of possible cross correlations is then $(K-1)8 + 4 = 8K - 4$. Each of these cross correlations is an inner product between two binary vectors, so the total set of $L(L-1)(8K-4)$ such correlations is subject to the Welch inner product bound.

$$\rho^2_{max} \geq \frac{1}{L(L-1)(8K-4)} \left[\frac{L(L-1)(8K-4)}{K} - 1 \right] \approx \frac{1}{K}$$

As noted earlier, this bound also applies to $\overline{\rho^2}$. Also 1/K is just the mean squared cross correlation between random bit streams when the chip transitions between the two sequences correspond. Thus, the bound on ρ^2 should be approachable for $1 << K$ and $L << 2^K$.

As noted in (A9), the impact of nonaligned bit transitions is to reduce the mean squared cross correlation of the random sequences from 1/K to 2K/3. Thus, the mean predetection SNR is

$$\frac{2E}{N_o'} \approx \frac{2}{(M-1)\overline{\rho^2} + N_o/E} = \left[\frac{M-1}{3K} + \frac{N_o}{2E}\right]^{-1}$$

from which

$$\eta = \frac{M}{2K} = \frac{3}{4}\left(\frac{N_o'}{E} - \frac{N_o}{E}\right) + \frac{1}{2K} \approx \frac{3\beta}{4}$$

As before, the efficiency, modified to include the loss of information in the channel, is $\eta' = \eta\ (1-H_L)$. As shown in Figure A3, the efficiency rises with increasing error rate within the range of "practical" error rates. The maxima of η, for the two values of E/N_o plotted in Figure 3, occur at error rates of about 0.4, yielding a theoretical maximum efficiency of about 0.66.

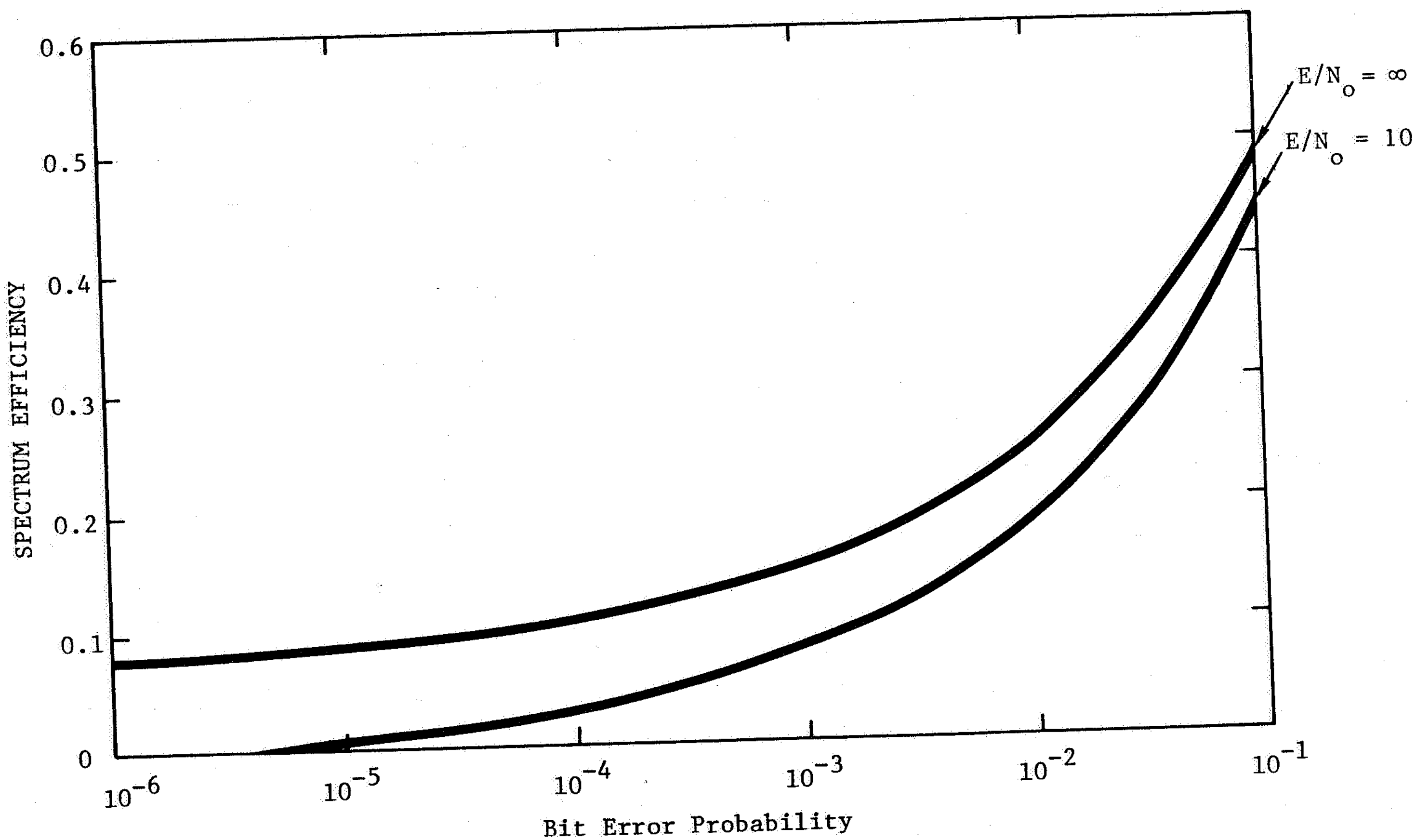

FIGURE A3
SPECTRUM EFFICIENCY OF ASYNCHRONOUS DIRECT SEQUENCE SSMA

A.3 FREQUENCY HOPPING

Frequency hopping (FH) systems operate by simultaneously changing the transmit and receive carrier frequencies at intervals throughout each message. This process can be visualized with the aid of Figure A4, which is a frequency versus-time plot for a single frequency-hopping link. Ideally, multiple links can operate with no mutual interference as long as their frequency-time plots never overlap. This is obviously a simplification, since some adjacent-channel interference will inevitably occur in practical systems.

Because of the difficulty of maintaining the phase coherence of the carrier from one hop to another, frequency shift keying (FSK) is often used in FH systems. For the purpose of this analysis, each FH transmitter sends one bit per hop using binary FSK. The receiver uses noncoherent detection to recover the transmitted signal. All signal levels are assumed to be equal at the receiver input. The effect of spectral "splatter" is not considered; for the case of equal signal levels, this simplification can be expected to have a minor impact. Also, for simplicity, the number of assigned hopping patterns is taken to be equal to the maximum number of simultaneous users.

The spectrum efficiency of a FH system, using the definition of equation (7), is

$$\eta' = \eta(1 - H_L) = \frac{M}{2N}(1 - H_L) \tag{9}$$

where $N = B/2R$ is taken to be the number of sub-channels, each sub-channel containing a mark frequency and a space frequency. In systems where the sub-channel spacing is greater (or less) than 2R, equation (9) must be adjusted accordingly.

A.3.1 Synchronous FH

If the hopping of the various links can be synchronized to a master timing source, then interference between links can be averted for $M \leq N$ by assigning nonoverlapping time-frequency sequences to the M users. Again, this condition is of primarily theoretical interest. The limiting value of η for this case is $\eta\ (M=N) = 0.5$. Then the network throughput per unit bandwidth is

$$\begin{aligned}\eta'(M=N) &= \eta(M=N)(1-H_L)\\ &= 0.5\left[1 + P_e \log_2 P_e + (1-P_e)\log_2(1-P_e)\right]\end{aligned}$$

where $P_e = 0.5 \exp(-E/2N_o)$

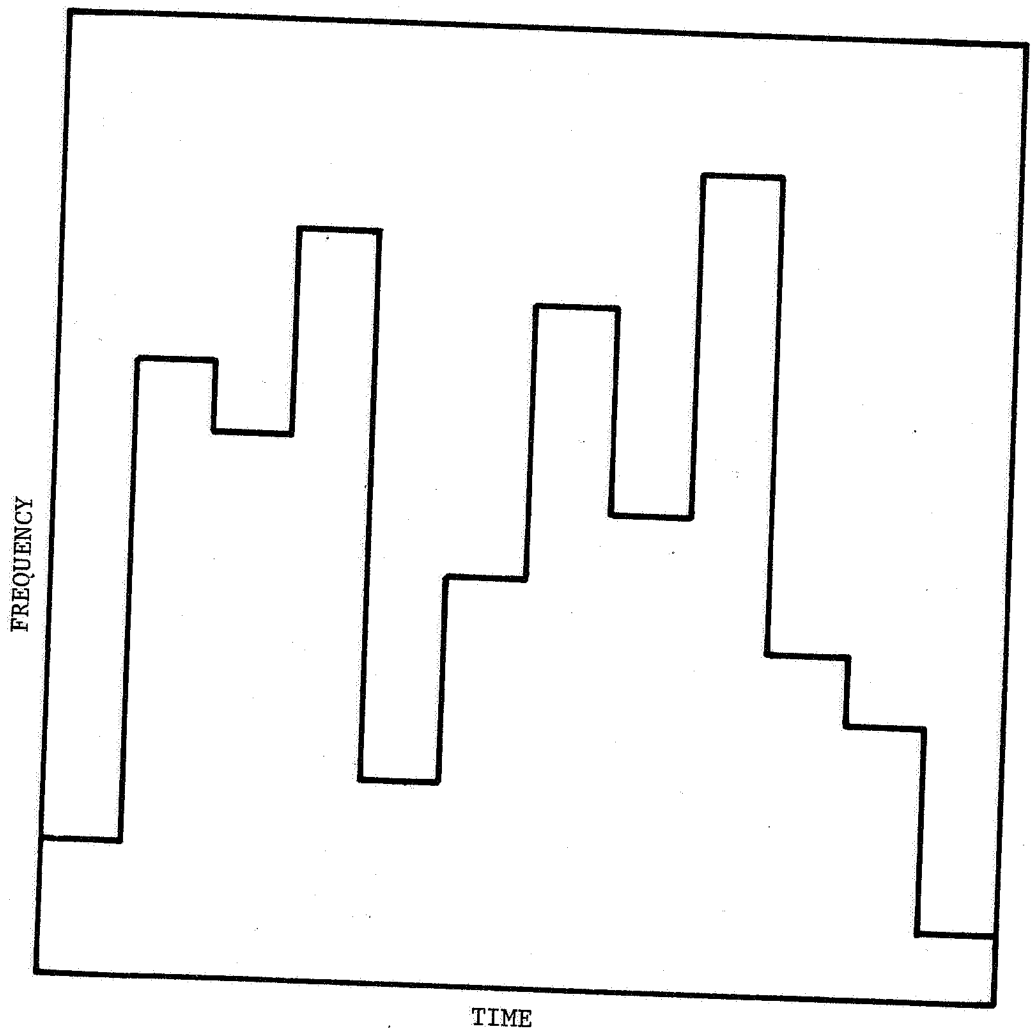

FIGURE A4
TYPICAL FREQUENCY HOPPING SEQUENCE

Note that of the N^N possible time-frequency trajectories, only N have been used for the first M=N users. For $N < M \leq 2N$ users, interference-free operation can no longer be guaranteed, but it is still possible to assign a unique hopping pattern to each user such that 2(M-N) signals suffer interference on each hop, with all signals suffering interference for an equal fraction of any long message.

This can be accomplished, in principle, by constructing the M codes such that on each hop, the 2(M-N) "unlucky" signals are selected at random out of the population M. The remaining 2N-M signals are assigned the "preferred" "noncolliding" sub-channels. Although this may not be a practical design procedure, it demonstrates that an interference probability (per signal) of 2(M-N)/M is attainable in principle. Furthermore, by this procedure no "collision" involves more than two signals. The average error probability is then

$$P_e = (2N-M)P_o/M + 2(M-N)P_1/M \tag{10}$$

$$\approx \left(\frac{2N-M}{M}\right) .5\exp\left(-E/2N_o\right) + \left[\frac{2(M-N)}{M}\right] 0.25$$

where P_i is the bit error probability with i interferers.

Given M, N, and E/N_o, the values of P_e and η' can be determined from equations (8) through (10). The upper set of curves in Figure A5 show η' as a function of P_e for N = 100. It can be seen that the curves for $E/N_o = 10$ and $E/N_o = \infty$ are essentially the same for $0.015 < P_e$. Note that meaningful values of the curves exist only for integral values of M, as shown by the staircase function under the $E/N_o = 10$ curve. The smooth curves actually represent the envelopes of discrete functions.

A.3.2 Asynchronous FH

As noted earlier, it is usually not practical to establish mutual synchronization between all active links in the system. For asynchronous FH systems, the bit error probability has been shown to be (A12).

$$P_e = \sum_{j=0}^{J} \left(1 - \frac{b}{N}\right)^{J-j} \left(\frac{b}{N}\right)^j \binom{J}{j} P_j \tag{11}$$

where J = M-1 = number of potential interferers

b = transmit duty factor

and P_j = probability of error with j interferers.

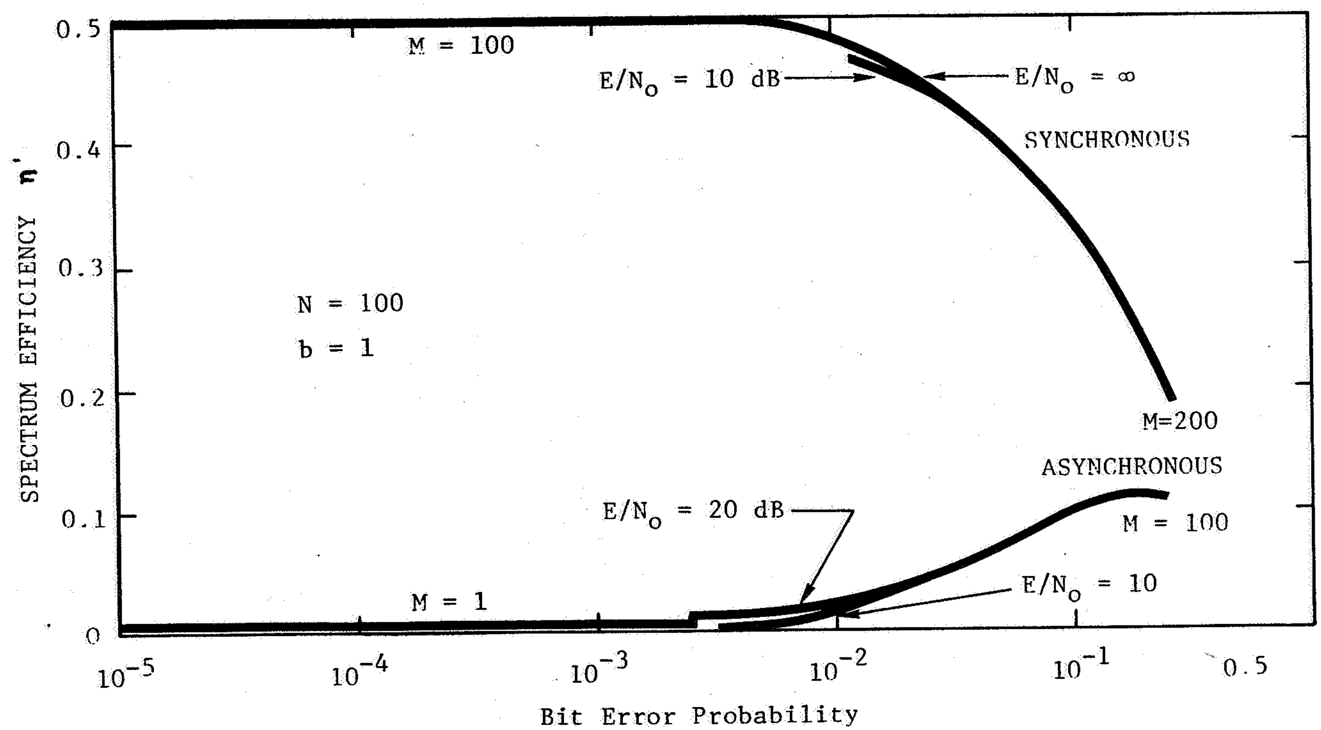

FIGURE A5
SPECTRUM EFFICIENCY OF FREQUENCY HOPPING SSMA
SPECTRUM EFFICIENCY

Equations (8), (9), and (11) thus provide for the evaluation of P_e and η' given M, N, b, and E/N_o. The computation is simplified by taking P_e to be 0.5 for $2 \leq j$. This is reasonable approximation since binary FSK alone is not a particularly interference-resistant technique. The error rate for two or more interferers (each having the same power as the desired signal) can be expected to rapidly approach 0.5.

The lower curves in Figure A5 show the spectrum efficiency (η) of asynchronous FH for E/N_o of 100 (i.e., 20 dB) and 10. Again, the variation in η' due to changes in E/N_o is negligible except at low error rates. The maximum value of η' is about 0.115 for error rates in the vicinity of 0.18.

A.4 PERFORMANCE OF ASYNCHRONOUS DS AND FH IN FADING

A.4.1 Direct Sequence

Fading can be caused by a wide variety of mechanisms. In some cases, direct sequence systems provide protection against multipath-induced fading by "resolving" the individual paths. This strategy can effectively convert a fading channel into a non-fading channel, but it requires a minimum bandwidth that is inversely proportional to the difference between the two "most nearly equal" path lengths.

Gardner and Orr, in reference (A10), treated the performance of a DS SSMA system in which a slow Rayleigh fading envelope is applied as an independent multiplicitive constant to each signal. Their analysis was extended by Hanlon and Gardner (10) to include Rician fading and Rayleigh fading in which the various fading envelopes are correlated. These two performances analyses can be used to provide simple, closed form approximations for spectrum efficiency under restricted conditions.

Specifically, Hanlon and Gardner show that the irreducible error rate (i.e., the error rate in the absence of background noise) for a DS SSMA system using coherent PSK through a channel with independent Rician fading is approximately

$$P_e(N_o=0) \approx \frac{(M-1)(1+C_1)\exp(-c_1)}{6K}$$

where C_1 is the ratio specular to diffuse power. (The corresponding expression for Rayleigh independent fading is obtained by seting $C_1=0$.) This yields directly

$$\eta' = \frac{M}{2K}(1-H_L) \approx \frac{3P_e(1-H_L)}{(1+C_1)\exp(-C_1)} \qquad \begin{matrix} M << L \\ C_1 < 2 \end{matrix}$$

Since this expression is based on the error rate in the absence of background noise, it represents an upper bound on η' for the stated conditions. Figure A6 shows $\eta'(P_e)$ for $C_1=0$ (Rayleigh fading) and for $C_1=1$ (specular component and diffuse component having equal power). As might be expected, the spectrum efficiency in Rayleigh fading channels is substantially reduced from the corresponding performance under non-fading conditions. The addition of a significant specular component does not appear to substantially alter this situation. These results obviously do not apply when the spread spectrum waveform is used to resolve the multipath signal into individual non-fading components.

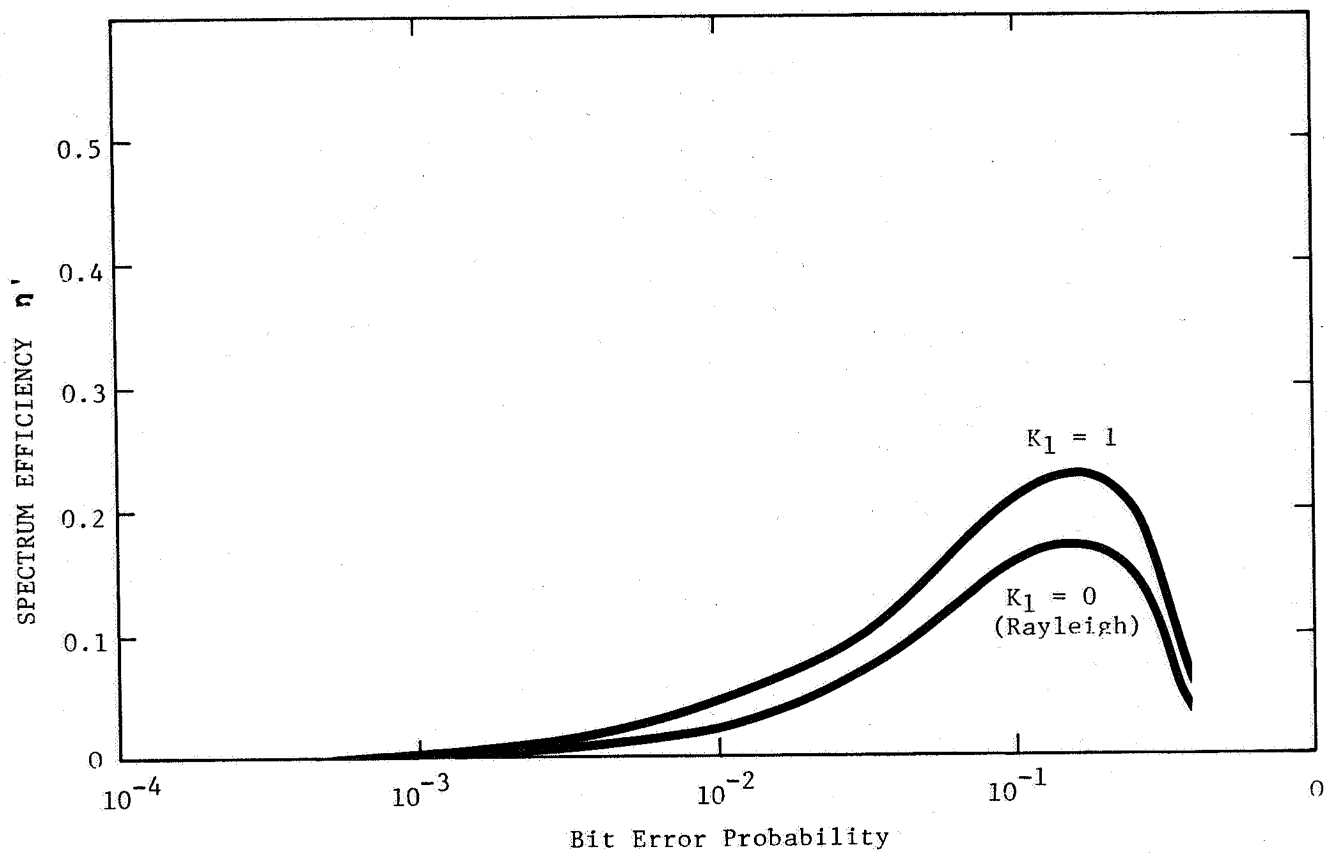

FIGURE A6
SPECTRUM EFFICIENCY OF ASYNCHRONOUS DIRECT SEQUENCE SSMA WITH FADING

A.4.2 Frequency Hopping

The error rate for most one-bit-per-hop binary FH systems can be derived from equation (11) by substituting the proper expression for P_j, the probability of error given that j interferers are present. For slow, flat Rayleigh fading,

$$P_o = \frac{1}{2 + \overline{E}/N_o}$$

where $\overline{E}$ is the mean received energy per signal. Also, since an interfering signal can appear either in the mark or the space channel (with equal probability),

$$P_1 = 0.25 \left[1 + \frac{1}{1 + \overline{E}/N_o} \right]$$

As before, P_j is taken to be 0.5 for $2 \leq j$, since even P_1 is greater than 0.25. Using this approximation with equations (9 and 11) yield the $\eta'(P_e)$ curves shown in Figure 7. At high values of E/N_o, the fading and non-fading FH curves differ only slightly. Under these conditions P_e is determined primarily by interference, which yields similar values of P_1 for the fading and non-fading cases. At lower values of E/N_o, however, noise induced errors begin to have a substantial impact, as shown by the lower curve in Figure A7.

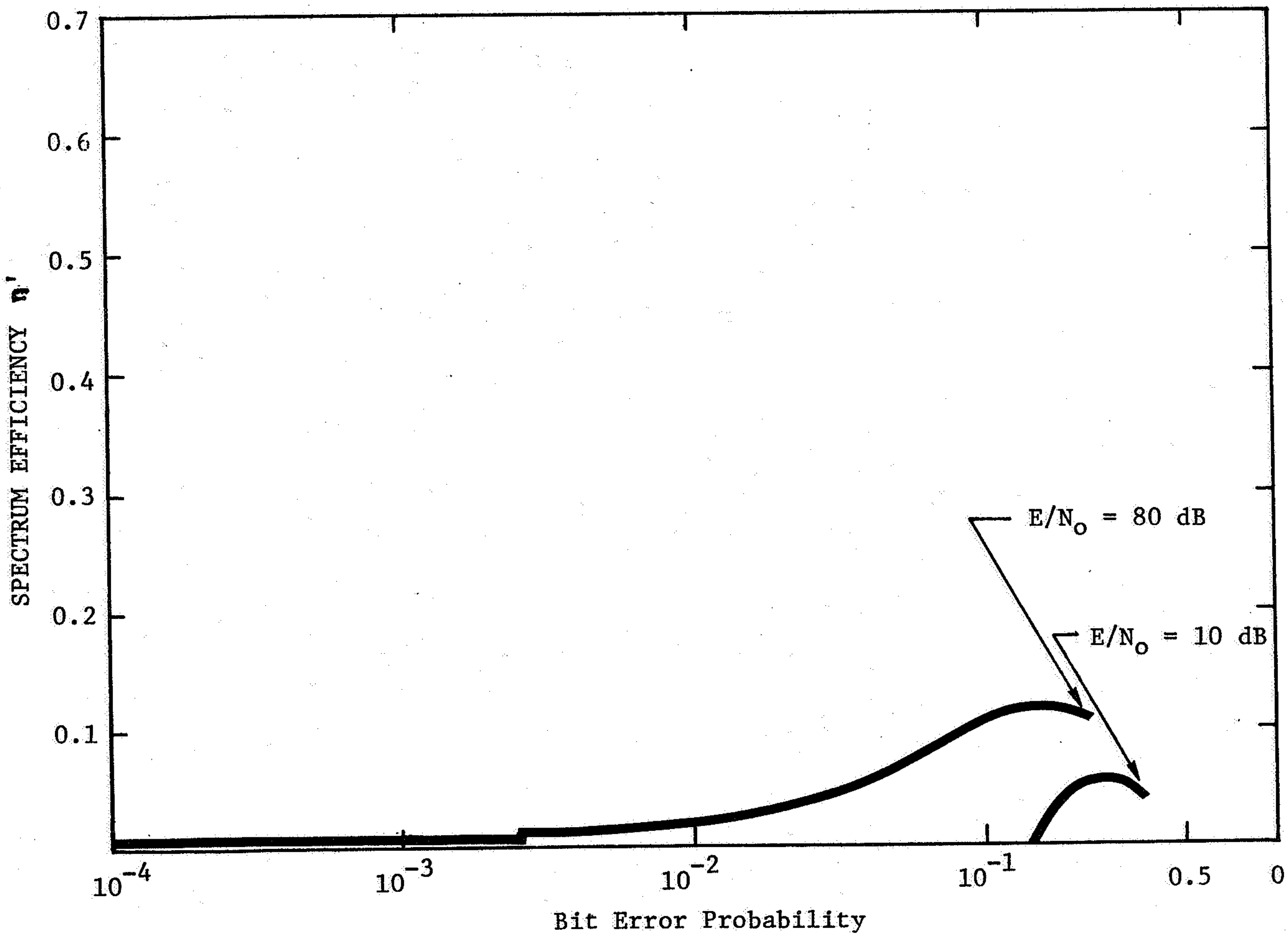

FIGURE A7
SPECTRUM EFFICIENCY ASYNCHRONOUS FREQUENCY HOPPING IN RAYLEIGH FADING

A.5 Error-Correction Coding: Fundamental Limitations

It can be argued that the results presented above are unduely restrictive because they are limited to binary signaling without error correction. It is clear that error correction coding can increase the spectrum efficiency of SSMA networks relative to the networks analyzed in the preceding sections. However, if error correction coding is allowed on SSMA networks it must also be allowed on the FDMA or TDMA networks which serve as a basis of comparison.

The Shannon capacity for channels with additive white Gaussain noise is

$$C = B_c \log_2 \left(1 + \frac{P}{N_o B_c}\right)$$

where B_c is the r.f. channel bandwidth allowed to each user and P is the received signal power. For FDMA, $B_c = B/M$ and P is the average received power per user. For TDMA, $B_c = B$ and P is M times the average power for one user. For SSMA, B_c=B, P is the average power, and N_o becomes N_o', as defined earlier.

Thus the total network channel capacities for FDMA, TDMA, and SSMA are

FDMA: $C_T = M(B/M)\log_2(1 + MP/N_oB)$

TDMA: $C_T = B\log_2(1 + MP_{AV}/N_oB)$

SSMA: $C_T = MB\log_2 \left[1 + \frac{P}{N_oB+(M-1)P}\right] = MB\log_2 \left[1 + \left(\frac{N_oB}{P} + M-1\right)^{-1}\right]$

Obviously, the first two expressions are the same. The best achievable spectrum efficiencies at arbitrarily low error rates are

FDMA: $\eta = MC/B = C_T/B = \log_2(1+MA)$

TDMA: $\eta = MC/B = C_T/B = \log_2(1+MA)$

SSMA: $\eta = MC/B = C_T/B = M\log_2 [1 + (A^{-1} + M-1)^{-1}]$

$\approx M/(M-1)$ for $1 << A$

≈ 1 for $1 \leq A$ and $1 << M$

where $A = P/N_oB$ has been used for brevity. The major result here is that the maximum spectrum efficiencies for FDMA and TDMA increase monotonically with M, while the maximum spectrum efficiency for SSMA decreases asymptotically to 1.0 for increasing M. This is basically due to the assumed lack of self-interference in FDMA and TDMA systems. The channel capacities for such systems can theoretically be raised to any desired level by increasing the transmitter power. Increasing the transmitter power in SSMA systems quickly results in a point of diminishing returns; the system becomes interference-limited. Moreover, increasing the number of SSMA users decreases the effective SNR for each user. For FDMA or TDMA systems, the SNR per user increase as the number of users increases (assuming that the average power per user and noise power density remain fixed), since adding users either decreases the bandwidth per user (in FDMA) or increases the ratio of peak to average power (in TDMA).

When the number of active users, M, is not constant, the average network channel capacity can be computed as

$$C_T = \sum_{M=0}^{\infty} C_T(M)\ P(M)$$

where P(M) is the probability that M users are active. Under these conditions it can be shown that SSMA can yield a higher average network capacity (for a fixed bandwidth, B) when the average number of active users is low (A17, A18). This is due to the assumption that even when few users are active, each FDMA user can occupy only 1/M of the system bandwidth; each TDMA user can occupy the total system bandwidth only 1/M of the time. If this constraint is removed (by trunking the available TDMA or FDMA channels, for example), then it is easily shown that the average SSMA channel capacity is always below the average TDMA or FDMA capacity.

A.6 DISCUSSION

The preceding derivations are designed to provide some basic constraints on spectrum efficiency under the stated conditions. In Sections A.1 through A.4, the assumption of a normal predetection distribution of noise-plus-interference was invoked in order to relate spectrum efficiency to error rate.

The spectrum efficiency of synchronous networks is primarily of theoretical interest. Most practical systems provide synchronization only between pairs of terminals that are communicating with one another. The results given above for synchronous networks indicate that direct sequence SSMA provides better spectrum efficiency than frequency hopping SSMA for the stated conditions. However, part of this apparent difference in performance is simply due to the difference in communication performance between coherently demodulated PSK and noncoherently demodulated FSK. Furthermore, it is easy to find examples in which frequency hopping with noncoherent binary FSK outperforms direct sequence with coherent binary PSK.

Consider, for example, the case in which the received power levels are no longer equal. Specifically, let

$$E_1 = E_2 = E_3 = \ldots E_{M-1} = E_M/\alpha \, , \; 1 \ll \alpha$$

It is easily shown, starting from equation (4) that the spectrum efficiency for asynchronous DS is reduced from $\eta \approx 3\beta/4$ to

$$\eta \approx \left[\frac{M}{M-2+\alpha}\right]\left(\frac{3\beta}{4}\right)$$
$$\approx \frac{M}{\alpha}\left(\frac{3\beta}{4}\right)$$

In many situations, it is not unusual to find 40 dB differences between the weakest and the strongest signal. Taking the hypothetical case of $M = 10$, $\alpha = 10^4$, and $N_o = 0$, and requiring $\beta = 1$ (a generously high value) we find

$$\eta \approx 7.5 \times 10^{-4}$$

For the slow frequency hopping approach treated above, the impact of the single strong interferer is restricted to a single sub-channel (although in extreme cases, splatter from adjacent sub-channels would also have to be considered). If $N = 100$, then a receiver not synchronized to the strong interferer "sees" it only one percent of the time. The error rate due to the strong interference alone is therefore on the order of 5×10^{-3}, which has only a second order effect when the total error rate is on the order of .02, for example, as in the case of 8 interferers of equal power in a 100 channel FH system. The spectrum efficiency for FH is then approximately as shown in Figure A9 for $P_e = 0.02$, or $\eta' \approx 0.05$. Although this efficiency may seem quite low, it is almost two orders of magnitude higher than the efficiency of asynchronous DS under the same conditions.

Regardless of spread spectrum modulation technique, it is clear that the spectrum efficiency of any given SSMA system will improve with decreasing signal-to-interference ratio until the point is reached at which so many errors are being made that the loss of information throughput overwhelms any gain in the number of links operating in a given bandwidth. For binary signals without forward error correction, this point corresponds to a relatively high error rate. The fundamental requirement for practical spectrum-efficient SSMA systems is the ability to operate at low signal-to-interference ratios. However, expanding the system bandwidth in order to achieve this goal by conventional spread spectrum techniques is not typically effective in terms of spectrum efficiency, since the permissible number of users is roughly proportional to the bandwidth, all other things being equal. What is needed, then, is an improvement in basic communication efficiency relative to uncoded binary signaling and the usual spread spectrum techniques, by themselves, do not provide this, since the required E_b/N_o for a given error rate in white Gaussain noise is not changed by the addition of spread spectrum modulation.

Any discussion of theoretical spectrum efficiency needs to be tempered by caveats on practical limitations. For spread spectrum systems, the attainable performance can be limited by the requirement for fast acquisition and reliable synchronization with economically feasible implementations. TDMA and FDMA systems also have practiced limitations. TDMA requires timing gaps to allow for finite propagation delays and can be expensive to implement. FDMA systems are generally easy to implement but, in practice, are ususally not free of interference. In addition, their spectrum efficiencies are limited by the guard bands required between channels. Examples have been produced in which hypothetical spread spectrum systems provide better spectrum utilization than conventional frequency-channelized approaches (see Reference A7 and Appendix B). The outcomes of such comparisons inevitably depend on the particular characteristics and parameters of the systems being compared. These characteristics and parameters, in turn, reflect the potential cost and comunication performance of the two systems. In practical situations, it may not be meaningful to compare SSMA with other techniques without specifying particular practical designs as part of the comparison process. A loss of generality in such specific comparisons is inevitable, but this is the price of improving the reliability of the result.

A.7 REFERENCES

A1. G. R. Cooper and R. W. Nettleton, "A Spread Spectrum Technique for High Capacity Mobile Communications," Conference Record of the 27th Vehicular Technology Conference, Orlando, Florida, March 16-18, 1977, pp. 98-103.

A2. R. W. Nettleton and G. R. Cooper, "Mutual Interference in Cellular LMR Systems Narrowband and Broadband Techniques Compared," Conference Record of MIDCON/77, Chicago, Illinois, November 8-10, 1977.

A3. R. W. Nettleton and G. R. Cooper, "Error Performance of a Spread-Spectrum Mobile Communications System in a Rapidly-Fading Environment," Conference Record of National Telecommunications Conference, Los Angeles, California, December 5-7, 1977.

A4. G. R. Cooper, "Multi-Service Aspects of Spread-Spectrum Mobile Communications System," Conference Record of MIDCON/77, Dallas, Texas, December 12-14, 1978.

A5. G. R. Cooper and R. W. Nettleton, "A Spread-Spectrum Technique for High-Capacity Mobile Communications," IEEE Transactions on Vehicular Technology, Vol. VT-27, No. 4, November 1978.

A6. P. S. Henry, "Spectrum Efficiency of a Frequency-Hopped-DPSK Mobile Radio System," Conference Record of the 30th IEEE Vehicular Technology Conference, 1977, pp. 7-12.

A7. D. J. Goodman, et al., "Frequency Hopped Multilevel FSK for Mobile Radio," 1980 International Zurich Seminar on Digital Communications.

A8. J. E. Mazo, "Some Theoretical Observations on Spread Spectrum Communications," Bell System Technical Journal, Vol. 58, No. 9, November 1979, pp. 2013-2023.

A9. M. B. Pursley, "Performance Evaluation for Phase-Coded Spread Spectrum Multiple-Access Communication--Part 1: System Analysis," IEEE Trans. Comm., Vol. COM-25, No. 8, August 1977, pp. 795-799.

A10. C. S. Gardner and J. A. Orr, "Fading Effects on the Performance of a Spread Spectrum Multiple Access Communication System," IEEE Trans. Comm., Vol. COPM-27, No. 1, January 1979, pp. 143-149.

A11. R. C. Hanlon and C. S. Gardner, "Error Performance of Direct Sequence Spread Spectrum on Nonselective Fading Channels," IEEE Trans. Comm., Vol. COM-27, No. 11, November 1979, pp. 1696-1700.

A12. S. A. Musa and W. Wasylkiwskyj, "Co-Channel Interference of Spread Spectrum Systems in a Multiple User Environment," IEEE Trans. Comm., Vol. COM-26, No. 10, October 1978, pp. 1405-1413.

A13. D. J. Toerrieri, Simultaneous Mutual Interference and Jamming in a Frequency Hopping Network, DARCOM Countermeasures/Counter-Countermeasures Office, DARCOM CM/CCM 79-6, October 1979.

A14. J. P. Costas, "Poisson, Shannon, and the Radio Amateur," Proc. IEEE, December 1959.

A15. S. H. Cameron, "The Efficiency of Spread Spectrum Signaling with Respect to Spectrum Utilization," 19th IEEE Electromagnetic Compatibility Symposium, August 1977.

A16. L. A. Berry, "Spectrum Metrics and Spectrum Efficiency: Proposed Definitions," IEEE Trans. EMC, Vol. EMC-19, No. 3, August 1977, pp. 254-260.

A17. R. C. Sommer, "Asynchronously Multiplexed Channel Capacity," Proc. IEEE, January 1966.

A18. R. C. Sommer, "The Noisy Asynchronously Multiplexed Channel," IEEE Trans. Information Theory, January 1967.

A19. L. R. Welch, "Lower Bounds on the Minimum Cross Correlation of Signals," IEEE Trans. Information Theory, May 1974.

APPENDIX B

SPECTRUM EFFICIENCY OF FDMA AND SSMA NETWORKS:

THE IMPACT OF FREQUENCY TOLERANCES AT LOW INFORMATION RATES

B.1 INTRODUCTION

In Frequency Division Multiple Access, guard bands are used to separate adjacent frequency assignments. If all transmitters and receivers are synchronized to a single carrier source, the guard bands need only compensate for the finite roll-off characteristics of the channel filters. In the more usual situation, however, each transmitter/receiver has its own frequency source. In this case, each guard band must be expanded to provide for the frequency tolerances.

With Spread Spectrum Multiple Access (SSMA), users are separated by codes, so only a single pair of guard bands at the edges of the SSMA channel is required. When frequency tolerances are considered, SSMA can surpass FDMA in spectal efficiency, i.e., in the number of users per unit bandwidth for a fixed set of system parameters. This statement can be translated to quantitative terms as follows.

B.2 FDMA NOTATION

Let
f_i = assigned frequency of the i^{th} carrier
f_c = center frequency of the FDMA band
d = relative frequency tolerance of the transmitters
R = information bandwidth
K_f = bandwidth expansion factor = r.f. bandwidth/R
b_f = required bandwidth per user
b_g = guard bandwidth per channel in the absence of frequency errors

Spectrum Utilization

It is assumed that the FDMA band is contiguous and that its bandwidth, B, is much smaller than its center frequency, f_c. It is also assumed that the frequency errors of the various user terminals are independent, and there is exactly one user per channel. The frequency separation between channels is then:

$$f_{i+1} - f_i = K_f R + f_i d + f_{i+1} d + b_g \approx K_f R + 2f_c d + b_g \qquad (1)$$

That is, the bandwidth per user is just the occupied bandwidth plus the two tolerances plus the minimal guard band, b_g.

B.3 ASYNCHRONOUS DIRECT SEQUENCE SSMA (CDMA)

Notation

Let
- M = number of active users
- R = information bandwidth
- K_s = bandwidth expansion factor = r.f. bandwidth/R
- d = relative frequency tolerance
- N_o = noise power density
- A_i = average output SNR with i active users
- p = 1 for noncoherent detection, 2 for coherent detection
- B = total bandwidth of channel
- b_s = bandwidth per user
- B_e = guard bandwidth at the channel edges for d=0
- n = number of code chips per code period

Spectrum Utilization

In this case, each user has a unique code. Again, the frequency errors of the various user terminals are taken to be independent. All users are assumed to be active simultaneously in the worst case. All signals are assumed to produce equal power levels at the receiver input. Here, the total width of the assigned channel is taken to be:

$$B = K_s R + 2f_c d, + B_e$$

i.e., the occupied bandwidth plus twice the tolerance plus B_e. The bandwidth per user is then:

$$b_s = \frac{K_s R + 2f_c d + B_e}{M} \tag{2}$$

The required bandwidth expansion factor, K_s, is determined by the processing gain required to achieve a given value of mean signal-to-noise ratio. The exact relationship between processing gain and output SNR in a direct sequence SSMA system depends on the specific set of codes and modulation technique being used. However, for a system using binary PSK, in which the codes can be approximated as random binary sequences, Pursley (1) has shown that the mean output SNR (power ratio) is approximately

$$\left[\frac{i-1}{3n} + \frac{N_o}{2E}\right]^{-1}$$

where i is the number of users, n is number of code chips per information bit and coherent detection has been assumed. In general,

$$A_i = \frac{p}{2}\left[\frac{i-1}{3n} + \frac{N_o}{2E}\right]^{-1}$$

solving for the code length, n, yields

$$n = \frac{2\,(i-1)}{3p}\left[\frac{1}{A_i} - \frac{1}{A_1}\right]^{-1}$$

Because the basic phase-shift keying process has a bandwidth expansion factor of 2, the overall bandwidth expansion factor, K_s, is 2n

Thus, given the available output SNR for a single user (A_1), the required output SNR for full system loading (A_m), and the system capacity (M), the required bandwidth expansion factor (K_s) can be estimated. This determines b_s as shown in equation (2).

Break-Even Point

In examining equations (1) and (2), it is clear that both are linear in f_c, d, and R. However, the coefficients of these variables in b_s are smaller than their counterpart in b_f by the factor M. Allowing d and R to remain fixed, the lines $b_s(f_c)$ and $b_f(f_c)$ always intersect at one and only one point for $1 < M$.

Furthermore, since both slopes are positive, $b_s < b_f$ above this point and $b_f < b_s$ below. Points above the break-even point require less bandwidth per user with SSMA; points below the break-even point require less bandwidth per user with FDMA. Analogous results hold if d or R is allowed to vary with all other parameters constant.

The break-even point is readily found by setting $b_f = b_s$ and solving for f_c, d, or R. For example,

$$f_o = \frac{R\,(K_s - K_f M) + B_e - Mb_g}{2d\,(M-1)} \tag{3}$$

Note that positive solutions exist only for $M\,(K_f R + b_g) < K_s R + B_e$, since the opposite condition would imply that the total required bandwidth for FDMA would be larger than the total bandwidth for SSMA even with perfect frequency control.

B.4 SSMA WITH SLOW FREQUENCY HOPPING

For this case, equations (1) through (3) are still valid. However, a modified approach to determining the required K_s may be needed. For slow-hopping systems, the signal-to-interference ratio can change drastically from one moment to the next, as the time-frequency slots used by the desired signal and the interferer briefly "collide." Instead of describing the system in terms of average SNR, it may be more meaningful to describe the probability that interference will occur in a given slot. This is especially significant for the "equal-signal-levels" model, since any interference can momentarily render the channel useless. For a single interferer, the probability of this for any particular "state" of the system is 1/L, where L is the number of channels (shared by the desired signal and the interferer). For an active user population M, the probability of interference is:

$$P = 1 - (1 - \frac{1}{L})^{M-1}$$

The required number of channels is then

$$L = \left[1 - (1 - P)^{(M - 1)^{-1}}\right]^{-1} \simeq \frac{M - 1}{P} \text{ for small } P \qquad (4)$$

The total bandwidth expansion factor, K_s, is then L times the "un-hopped" bandwidth per channel, assuming that the channel are spaced as closely as possible.

It is interesting to apply these results in two hypothetical examples where the information rate is low.

Example 1

Let us use the following parameters to compare FDMA with direct sequence SSMA.

$d = 2 \times 10^{-6}$ ($\pm 0.0002\%$)
$R = 100$ Hz
$M = 100$ Users
$A_1 = 30$ dB
$A_m = 10$ dB
$K_f = 2$
$b_g = 100$ Hz
$p = 2$ (coherent detection)

Then,

$$n = \frac{2(M-1)}{3P}\left[\frac{1}{A_M} - \frac{1}{A_1}\right]^{-1} = 333$$

The spread spectrum signal bandwidth is then $K_sR = 2nR =$ 66.6 kHz. For convenience, we take the combined guard bandwidth at the edges of the spread spectrum band to be equal to the spread spectrum signal bandwidth. Thus, by equation (3) f_o is 261 MHz. The model indicates that at higher frequencies, SSMA is more spectrum-efficient. The total required bandwidth at 261 MHz, by either equation (1) or equation (2), is just over 134 kHz, or about 1 kHz per user.

Example 2

Now consider an asynchronous SSMA system with slow frequency hopping. Let

$$
\begin{aligned}
P &= 0.05 \\
M &= 100 \text{ users} \\
R &= 100 \text{ Hz} \\
d &= 2 \times 10^{-6} \\
K_f &= 2 \\
b_g &= 100 \text{ Hz} \\
B_e &= 4R
\end{aligned}
$$

Then, by equation (4), L = 1930 channels. The "un-hopped" bandwidth is K_fR = 200 Hz, so the total spread spectrum signal bandwidth is 386 kHz. That is, K_s = 386 kHz/100 Hz = 3860.

By equation (3), the break-even frequency is f_o = 900 MHz. The total required bandwidth at this frequency is 390 kHz, or 3.9 kHz per user.

Discussion

It can be seen that the factors favoring SSMA over FDMA are:

(1) Loose frequency tolerances.

(2) Low information rates.

(3) High operating frequencies.

(4) Low marginal cost of transmitter power (relative to the cost of precision oscillators or TDMA), since SSMA typically requires either increased transmitter power or coding gain, or both, to combat the combined effects of background noise and self-interference. This is reflected in the requirement for direct sequence SSMA that $A_m < A_1$. If A_m is an acceptable SNR, then, in a permanently interference-free system, A_1 could be reduced to this level. In an SSMA system, however, the "spread" between A_1 and A_m fixes the relationship between the required processing gain and the number of users.

In comparing SSMA with FDMA, the effect of frequency tolerances can be neglected only if the absolute tolerances are small with respect to the information rate, or if all network terminals share a common frequency standard. Otherwise, ignoring frequency tolerances can bias the tradeoff in favor of FDMA.

B.5 REFERENCES

1. Michael B. Pursley, "Performance Evaluation for Phase-Coded Spread Spectrum Multiple-Access Communication--Part I: System Analysis," IEEE Trans. Comm., Vol. COM-25, No. 8, August 1977.

APPENDIX C

MONTE-CARLO SIMULATION OF LAND MOBILE RADIO WITH SLOW FREQUENCY HOPPING

C.1 OBJECTIVES

The purpose of this simulation is to provide a preliminary assessment of the feasibility of using slow FH for land mobile services. The sensitivity of the model to change in certain parameters is investigated.

C.2 SCOPE

In the development of the simulation models described here, it soon became evident that the specific parameters of the models would depend on the type of service being provided. Because of the potential expense of studying a number of potential types of services in depth, a hypothetical Citizens Band service was selected for simulation. However, the basic models described here are easily modified to embody the characteristics of other services.

C.3 DESCRIPTION OF MODEL

Every meaningful model requires the use of approximations, simplifications, or other constraints. The following description documents these and provides some comments on their impact.

C.3.1 Frequency Hopping

One of the basic features of the model presented here is that all active transmitters are assumed to hop in mutual synchronism. That is, they are assumed to change channels simultaneously. In actual operation, the various transmitters will hop asynchronously. The impact of this simplification is that it results in a predicted average frequency of interruption (by interference) that is too low. However, it also results in a prediction of the average duration of each interruption that is too high by the same factor. So prediction of the average signal-to-interference ratio is unbiased in this respect. However, the other performance parameter predicted in this study -- articulation -- is based on previously-published empirical evidence (see Section C3.4) so it is hard to say to what extent, if any, it may be biased by this simplification.

No attempt is made to model the hopping codes of the various transmitters. Instead, each transmitter is assumed to select a channel at random on each hop. All channels have an equal probability of being selected by any given transmitter on each hop. The hops of the various transmitters are statistically independent. This is expected to be an accurate representation for the "average" selection of M active user codes out of a larger population of K well-designed user codes. For the worst-case selection of M out of K codes, the performance predicted by the simulation may be slightly optimistic. However, most of the simulation runs are based on the use of 100 channels, and this results in an extremely large potential set of user codes. (Approximately $100! = 10^{158}$ codes, if all sequences using each channel once and only once per code cycle are allowed). Thus, the probability of randomly selecting two codes that are highly correlated is small, unless this code set is poorly designed.

Hopping rates of 1 hop per second and 20 hops per second are investigated. Hopping rates lower than 1 hop per second will provide reduced voice privacy and will engender long synchronization lags, which are unacceptable in real-time voice communication systems. Hopping rates much in excess of 20 hops per second may be difficult to implement with inexpensive synthesizers.

C.3.2 Sources of Performance Degradation

The major source of performance degradation is assumed to be interference from frequency-hopping transmitters. Sources of degradation that are neglected include:

- Background noise
- Fading
- Interference from other services
- Random FM
- Harmonic distortion
- "Click" noise resulting from the phase discontinuties between hops
- Receiver synchronization errrors

In addition, the model neglects the possibility of intermodulation products generated by the passage of mutliple FH signals through nonlinear circuits. However, the model does compute the effect of "splatter" from adjacent-channel interfers and from interfers that are two channels removed from the intended signal. Splatter resulting from interference that is more than two channels removed from the intended signal is assumed to have a negligible impact.

The overall impact of these simplifications obviously bias the predicted performance in a positive direction. Thus, when a particular simulaton run predicts poor performance, it can be concluded that the performance of an operational system would be poor. When the run predicts good results, the only conclusion is that good performance <u>might</u> be achievable in an operational system.

C.3.3 Propagation Model

The path losses between transmitters and receiver are computed by the Longley-Rice propagation model (Reference A-1). This model predicts the long-term median transmission loss between points, given parameters such as antenna heights, terrain irregularity, frequency, and distance. The specific parameters used in computing transmission loss are presented in Table C-1.

The values of refractivity, ground permittivity, and ground conductivity shown in Table C-1 are recognized as "typical" values; transmission loss at 915 MHz is relatively insensitive to changes in these parameters, except under conditions of unusual refractivity. Random antenna siting (with respect to the surrounding terrain) is assumed for both mobiles and base stations, on the hypothesis that most base stations would be located in the user's home or office. Antenna heights (above

TABLE C-1

PROPAGATION PARAMETERS

PARAMETER	VALUE
Surface Refractivity	301.
Ground Permittivity	15. esu
Ground Conductivity	0.005 mho/meter
Surface Irregularity	90 meters
Frequency	915 MHz
Antenna Siting	Random (Mobile & Base)
Mobile Antenna Height	2. meters
Base Antenna Height	7. meters

ground) of 2 and 7 meters are assumed for the mobile and base, respectively. The surface irregularity (90m) is typical of hilly terrain.

In order to minimize the cost of running the simulation, a look-up table of transmission loss versus distance was constructed. The table covers distances from 1.0 km to 100. km in logarithmically spaced increments of about two percent. The largest difference in transmission loss between adjacent entries in the table is 1.0 dB. However, the quantization interval for the most frequently encountered distances is less than 0.5 dB.

The geographic variability in path loss for a given transmitter-receiver distance is estimated by Reference A-2 as:

$$\sigma_L = 6 + 0.55\sqrt{d_H/L} - 0.004\,(d_H/L)\ \mathrm{dB}$$

where d_H is the terrain irregularity and L is the wavelength. This value is used to produce a normally-distributed, zero mean adjustment to the median transmission loss.

For transmitter-receiver separations less than 1.0 km, the Longely-Rice model is not applicable. Free-space loss is assumed for this case, which occurs infrequently in most of the simulation runs.

The propagation models described above are based largely on data taken in areas free of urban development. In a recent paper (Reference C-3), Longley suggested modifications for use in urban areas. The suggested correction to the median tranmission loss (in dB) contains a constant term, a frequency-dependent term, and a distance-dependent term. In computing signal-to-interference ratios, only the distance-dependent term is significant, since the other two terms are the same for the signal and the interference. The distance-dependent term has a negative sign, which implies that despite higher overall transmission losses in urban areas, the transmission loss increases with distance at a somewhat lower rate than in undeveloped areas. However, the overall change in the rate of fall-off is small for the range of distances used here.

C.3.4 Antennas And Power Levels

All stations are assumed to have antennas that are omnidirectional in azimuth and of equal gain. Equal transmitter power levels are assumed. Since the system is assumed to be interference-limited, no other assumptions about antenna gains or transmitter powers are required.

C.3.5 Basic Simulation Models

Two basic simulation models have been developed. The simplest computes the statistics of the r.f. signal-to-interference ratio for given statistical distributions of transmitter/receiver distances over a specified coverage area. The second model goes one step further and estimates the intelligibility of the resulting speech based on the pattern of interruptions, using empirical results by Miller and Licklider (C-4).

Both models assume that mobile-to-base transmissions and base-to-mobile transmissions occupy separate, disjoint frequency bands. Primary emphasis is on evaluating the performance of the mobile-to-base links, since, in the absence of repeaters, mobile-to-mobile operations would probably also be conducted in this band, thus making the traffic intensity higher than in the base-to-mobile band. A few simulation runs were also performed for mobile-to-mobile operation, although it is expected that this mode of operation would be of less importance than it is in the 27 MHz citizens band because of the discrete-address nature of the system.

C.3.5.1 Signal-to-Interference Model

A simplified flow chart of the signal-to-interference simulation model is shown in Figure C-1. After the various parameters have been read in, a random trial is conducted to select the distance between the receiver and the transmitter generating the intended signal. Another set of random trials is conducted to select the distances between the receiver and all other active transmitters. Next, a random trial is conducted to select a channel for each active transmitter. The absolute value of the frequency difference (measured in number of channels) between the intended signal and each potential interference is computed and stored. For co-channel interference or interferences within two channels of the intended signal, the path loss is computed. Finally, the signal-to-interference ratio is computed, ignoring those active transmitters that are more than two channels removed from the intended signal.

This process is iterated 200 times, and the distribution of the resulting values of r.f. signal-to-interference ratio is computed and plotted.

The second simulation model estimates a measure of speech intelligibility by a procedure that will be described in the next section. In order to perform these computations it is necessary to first estimate the total fraction of the voice message that is interrupted by interference and the average rate of interruption.

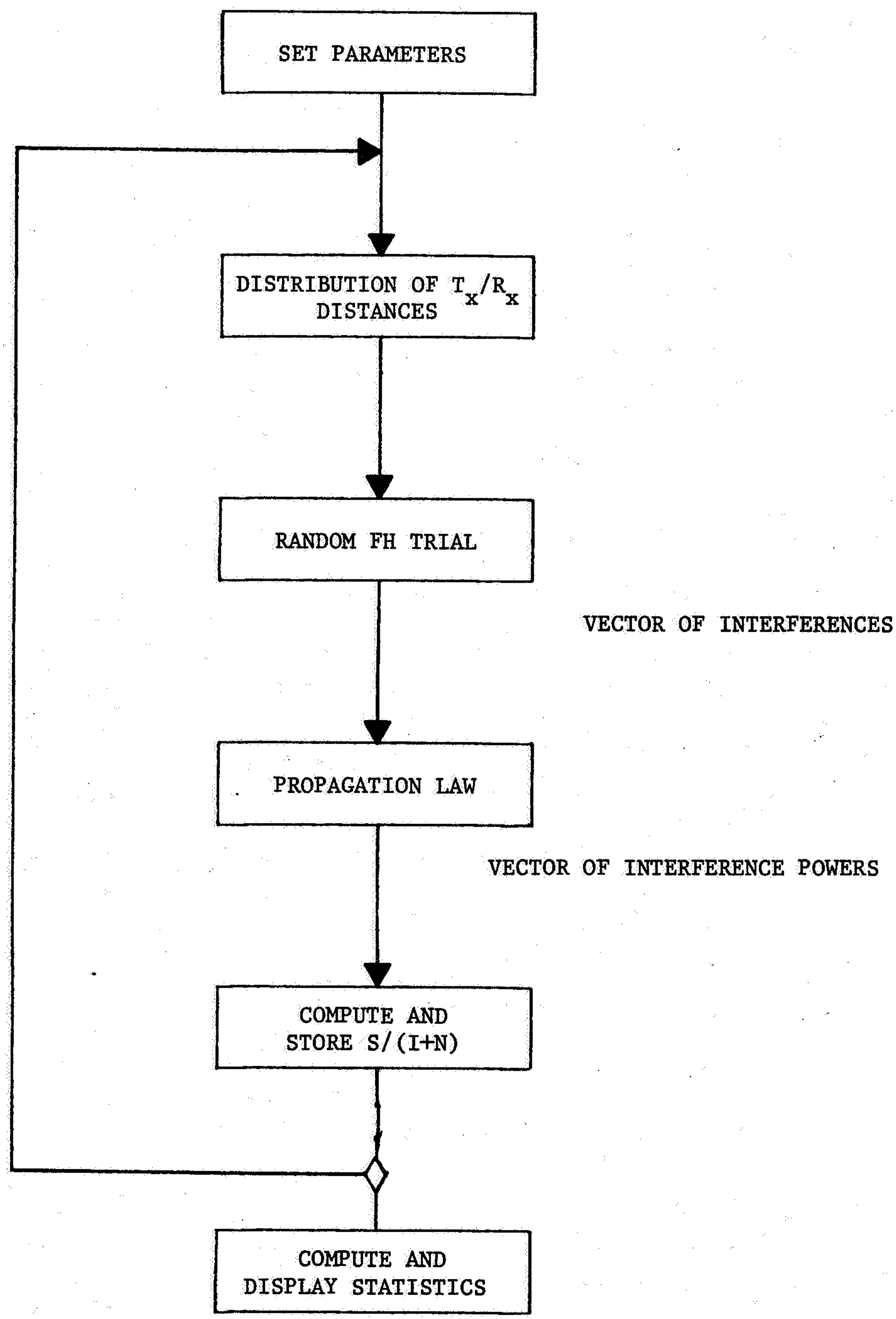

FIGURE C1
SIGNAL-TO-INTERFERENCE SIMULATION: SIMPLIFIED FLOW CHART

Figure C-2 is a simplified flow chart of the second simulation model. It should be viewed as a functional description of the computer program, because some operations deviate slightly from the format of Figure C-2 for reasons of computational efficiency.

After the parameters are read in, all temporary storage arrays are cleared. A random trial is then conducted to assign a channel to each active transmitter. On the first pass through this inner loop, random trials are also perfomed to select transmitter/receiver distances. Path losses are computed and stored so that they will not have to be recomputed on future hopping trials. The r.f. signal-to-interference ratio (C/I) is computed and a counter is incremented if the C/I exceeds 6 dB. For the purpose of estimating intelligibiltiy, the audio SNR is assumed to be infinite for 6 dB < C/I and zero for C/I < 6 dB. This is a simplification based on the threshold/capture characteristics of narrowband FM. It would be a serious oversimplification if most of the observed C/I values were in the region of 6 dB, but because the observed trial values of C/I are spread over a wide range, the approximation should be acceptable.

One hundred such passes are made through the inner loop, with the transmitter/receiver distances and path losses fixed. The result is the number of passes, k, for which C/I < 6 dB. The fraction of hops for which significant interference occurs is then k/100. A count is also kept of the number of interference bursts, i.e., the number of runs of interference-contaminated hops. If this count is L, then the average rate at which interference bursts occur is L/100 times the hopping rate. These two measures - the fraction of the hops on which interference (C/I < 6 dB) occurs and the average number of interference bursts per second are use to compute the articulation, based on procedures to be described in the next section.

This entire process is repeated up to 200 times, with a different randomly-selected set of transmitter/receiver distances on each pass. The distribution of the resulting articulation scores is computed and plotted.

C.3.5.2 Intelligibility Model

The intelligibility of interrupted speech has been studied empirically by Miller and Licklider (Reference C-4). The basic measure of intelligibility used in their study is the Harvard phonetically balanced (PB) articulation test (Reference C-5), which scores the ability of listeners to correctly interpret monosyllable words.

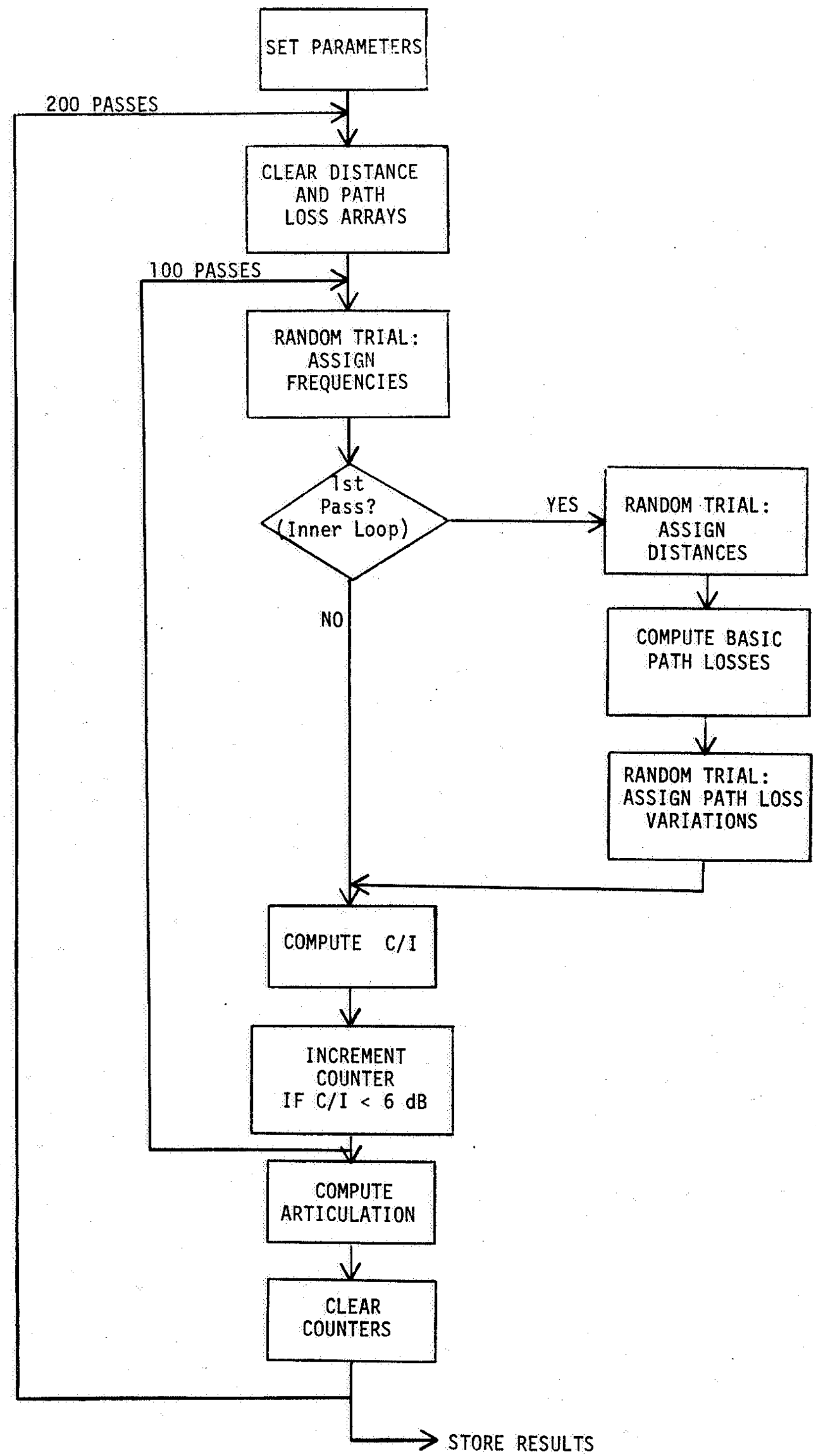

FIGURE C2
SIMPLIFIED FLOW CHART OF INTELLIGIBILITY SIMULATION

Figure C-3 illustrates one result of the Miller and Licklider study. It shows word articulation scores for speech interrupted by noise, as a function of "noise-time fraction," i.e., the duty factor of the noise. The results of interrupting the speech 1, 10, 100 and 1000 times per second are shown. Although these data are based on periodic interruptions, interruptions with irregularly spaced bursts of noise yeilds essentially the same results.

In order to use the results to predict intelligibility, the average rate of interruption and the average "noise-time fraction," or noise duty factor, must be known. Since the Miller and Licklider results are empirical, some form of interpolation is required when the rate of interruption and "noise-time fraction" produced by the simulation do not coincide with the corresponding values used in the empirical intelligibility tests. In order to satisfy this requirement, a table of articulation versus "noise-time fraction" was computed for each of the four rates of interruption shown in Figure C-3. Since each series of 100 hopping trials has only 101 possible outcomes in terms of "noise-time fraction," the size of this lookup table is 4 x 101 = 404 elements. Linear interpolation is used between the empirical data points at each rate of interruption. For a given noise-time fraction and rate of interruption, the articulation is estimated by linearly interpolating between points in the table corresponding to the rates of interruption that bracket the observation. Rates of interruption below one per second are assumed to produce the same articulation scores that result for one interruption per second.

The PB word articulation test used by Miller and Licklider is significantly more demanding than several other scoring methods that have been used (Reference C-6). A PB word articulation score of 75 percent is deemed to be acceptable for "typical" communication purposes.

C.3.6 Distribution of Transmitter-to-Receiver Distances

In all of the simulation runs, the receiver is assumed to be located at the center of a circular region having a radius of 30 km. Interfering transmitters are assumed to have a uniform probability of being within any differential area within this region. In order to account for the fact that mobiles cannot, in most circumstances, approach within a few meters of the base station antenna, an interference-free ring having a radius of 100 meters is created around the receiver. Although these assumptions are obviously simplifications, they are similar to the ones that were used in a recent analysis by Torrieri (Reference C-7).

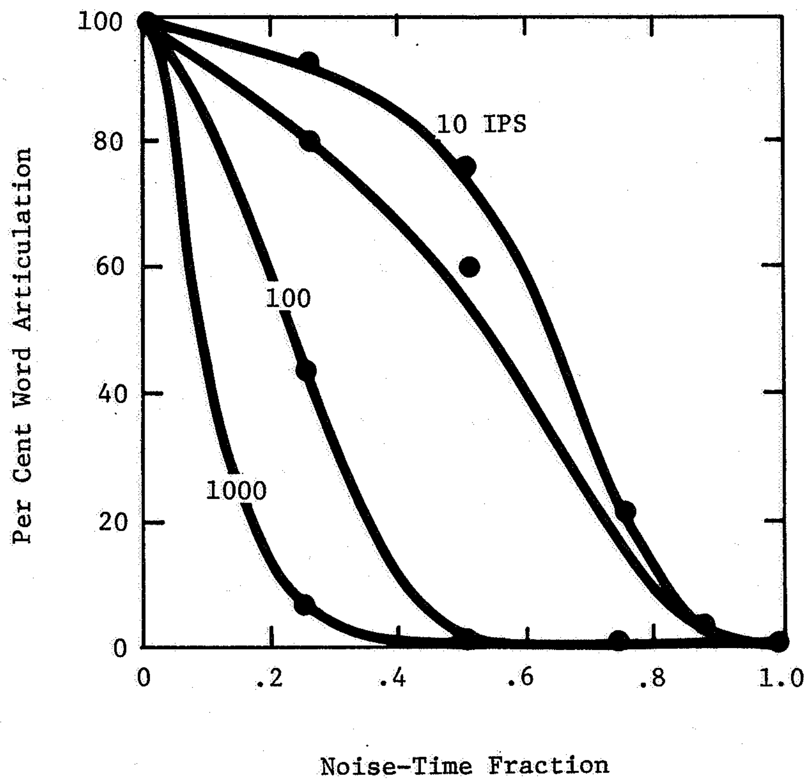

FIGURE C3
EMPIRICAL BASIS OF INTELLIGIBILITY MODEL

The uniform annular geographic distribution of interferers leads to the distribution of transmitter-to-receiver distances shown in Figure C-4 for a receiver in the center of the region.

It is reasonable to expect that the distribution of distances between the receiver and the transmitter with which communication is being attempted will not be in agreement with Figure C-4. In particular, users will soon learn that the range of the system is limited (even in the absence of interference), so it is unreasonable to assume that the longest distances between the receiver and the intended transmitter are the most likely to be attempted. In fact, one would suppose that, after a certain point, the probablity that of user attempting communication over a given distance will decrease monotonically with distance. On the other hand, it is also unlikely that users will often attempt communication over very short ranges because a moving vehicle will quickly leave such a short-range region. These arguments suggest a unimodel distribution communication distances like the one shown in Figure C-5. The functional form of this distribution is:

$$p(x) = (x/X_0^2)e^{-x/x_0}$$

where x is the distance over which communication is being attempted and $2x_o$ is the mean of the distribution.

In an operational scenario, there is likely to be a significant amount of "coupling" between the number of interferers and the mean distance over which communication is attempted. No attempt has been made to model this effect, and further investigation in this area is needed.

C.3.7 Output Format

The output produced by both simulation models is in the form of a cumulative distribution function (CDF) for each simulation run. One model estimates the CDF of the signal-to-interference ratio; the other estimates the CDF of the word articulation. Each type of CDF represents the probability (taken over the ensembles of transmitter locations and frequency hops) that the computed variable falls below any particular value.

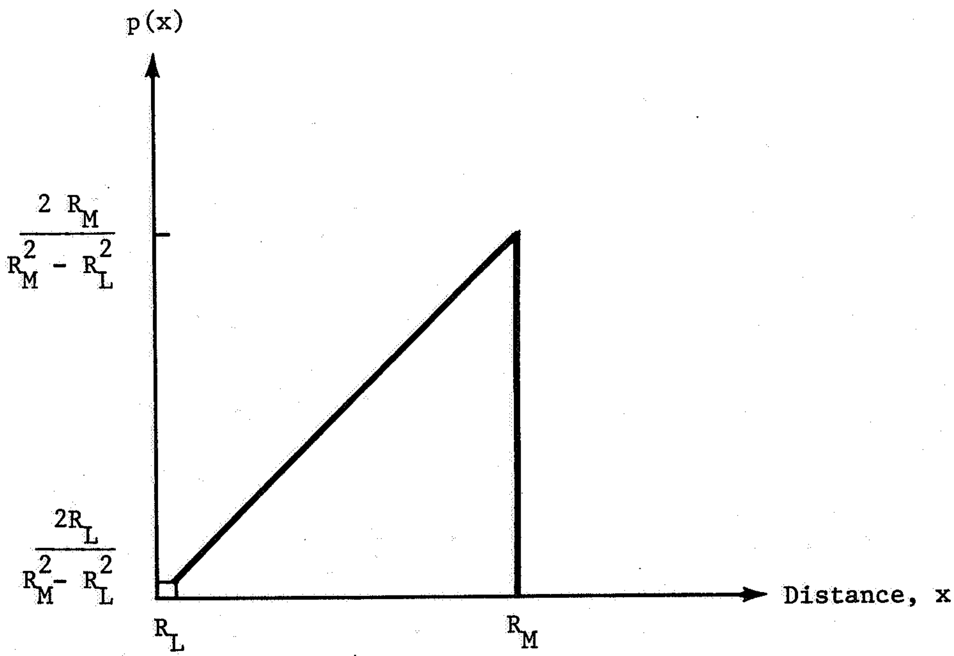

FIGURE C4
RADIAL DENSITY FOR UNIFORM DEPLOYMENT OF INTERFERERS

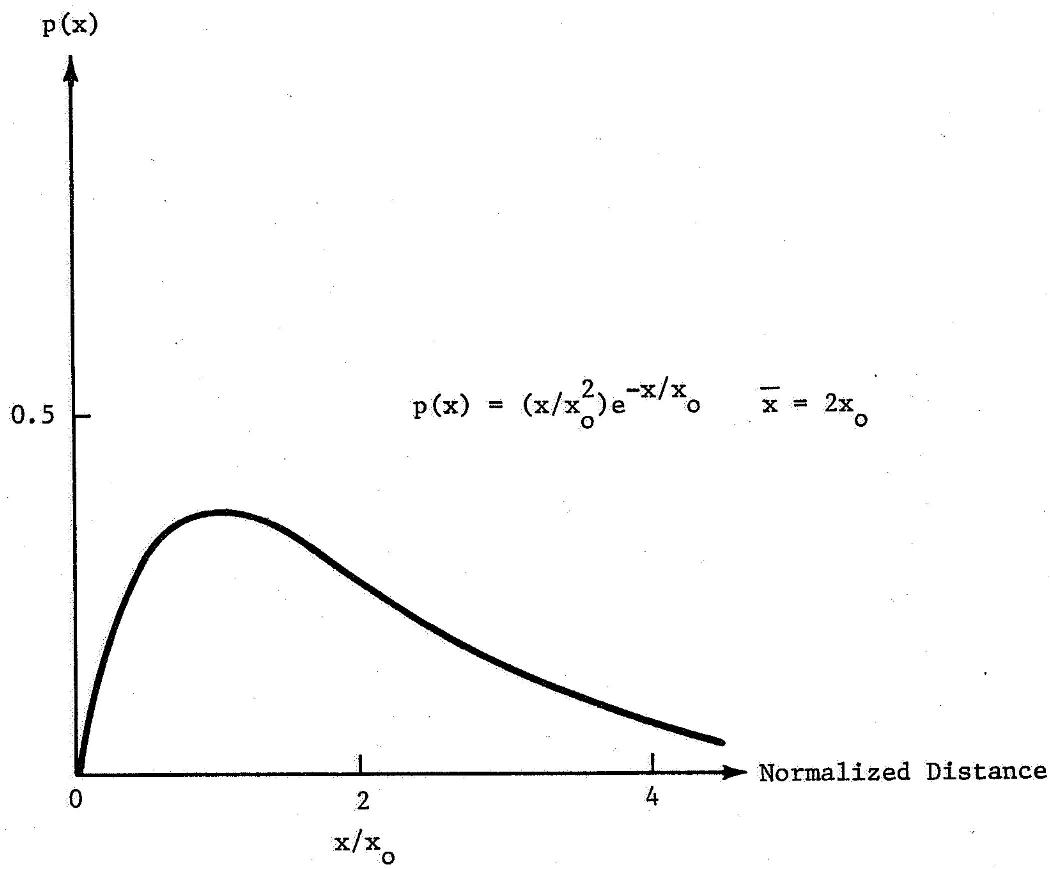

FIGURE C5
DENSITY OF DISTANCE FROM RECEIVER TO INTENDED TRANSMITTER

C.4 SIMULATION RESULTS

C.4.1 Signal-to-Interference Simulation Model

All simulation runs on this model are mobile-to-base runs, with a mean communication distance of 8 km (5 miles). A hopping rate of one per second is used. "Splatter" from adjacent-channel interference is neglected in this preliminary model.

In order to provide a baseline against which to compare the performance of a frequency hopping system, several runs were made for a single channel system. The first run, shown in Figure C-6, provided the CDF of the r.f. signal-to-interference ratio when one interferer is present continuously on the single channel. In order to facilitate interpretation of the CDF, the locations median and the lower decile are marked on the abscissa by a triangle (∇) and a vertical arrow (↑), respectively. This convention will be followed on all of the graphs.

It can be argued, heuristically, that an r.f. signal-to-interference ratio of about 10 dB is required for adequate communications. This rather arbitrary criterion is based on the threshold of a narrowband FM receiver, which is typically a few dB below the 10 dB level. Note that the signal-to-interference ratio varies not only with the various transmitter/receiver distances, but also with time (i.e., with hopping) for any given configuration of transmitter-receiver distances.

It can be seen from Figure C-6 that for a single channel and a single interferer, the 10 dB S/I criterion was met on only about 62 percent of the trials. For two interferers on a single channel (see Figure C-7) this fraction reduces to about 37 percent. For four interferers (see Figure C-8) it is further reduced to about 26 percent. The implication of this is that the degree of frequency reuse (by geographic separation) possible in a conventional narrowband 900 MHz system with the assumed distributions of distance will be limited. However, this issue needs to be addressed in more detail, particularly if the dominant mode of operation in such a system turns out to be mobile-to-mobile (without repeaters).

The results of the signal-to-interference simulation for a 100-channel frequency-hopping are illustrated in Figures C-9 through C-15, for 5, 10, 20, 40, 80, 160 and 200 interferers, respectively. In computing the signal-to-interference ratio on each trial, provision must be made for the case in which no interference occurs, due to a particularly favorable selection of channels on that hop. In this case, a r.f. signal-to-

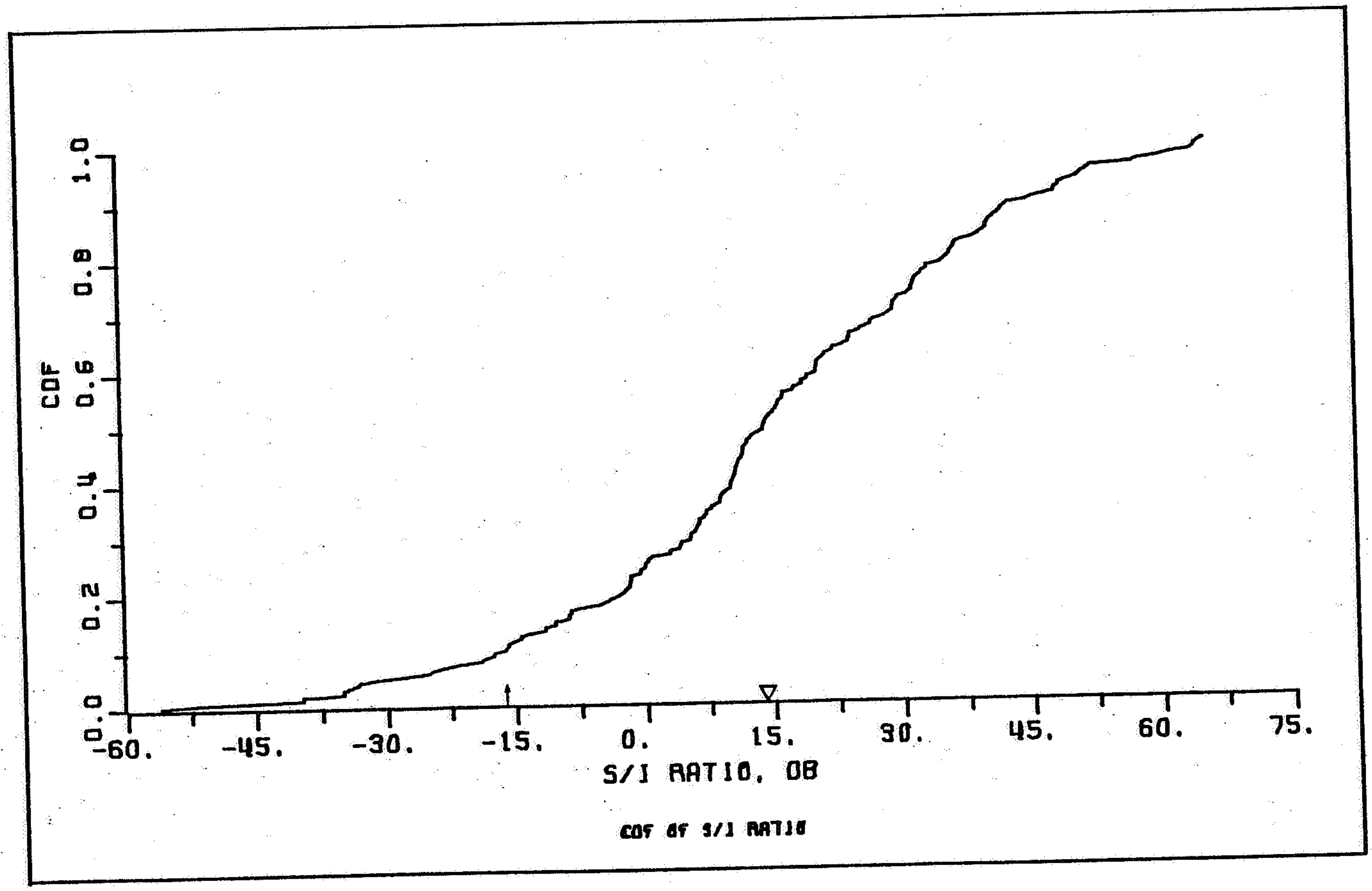

FIGURE C6
CUMULATIVE DISTRIBUTION OF R.F.
SIGNAL-TO-INTERFERENCE RATIO: 1 CHANNEL,
1 INTERFERER

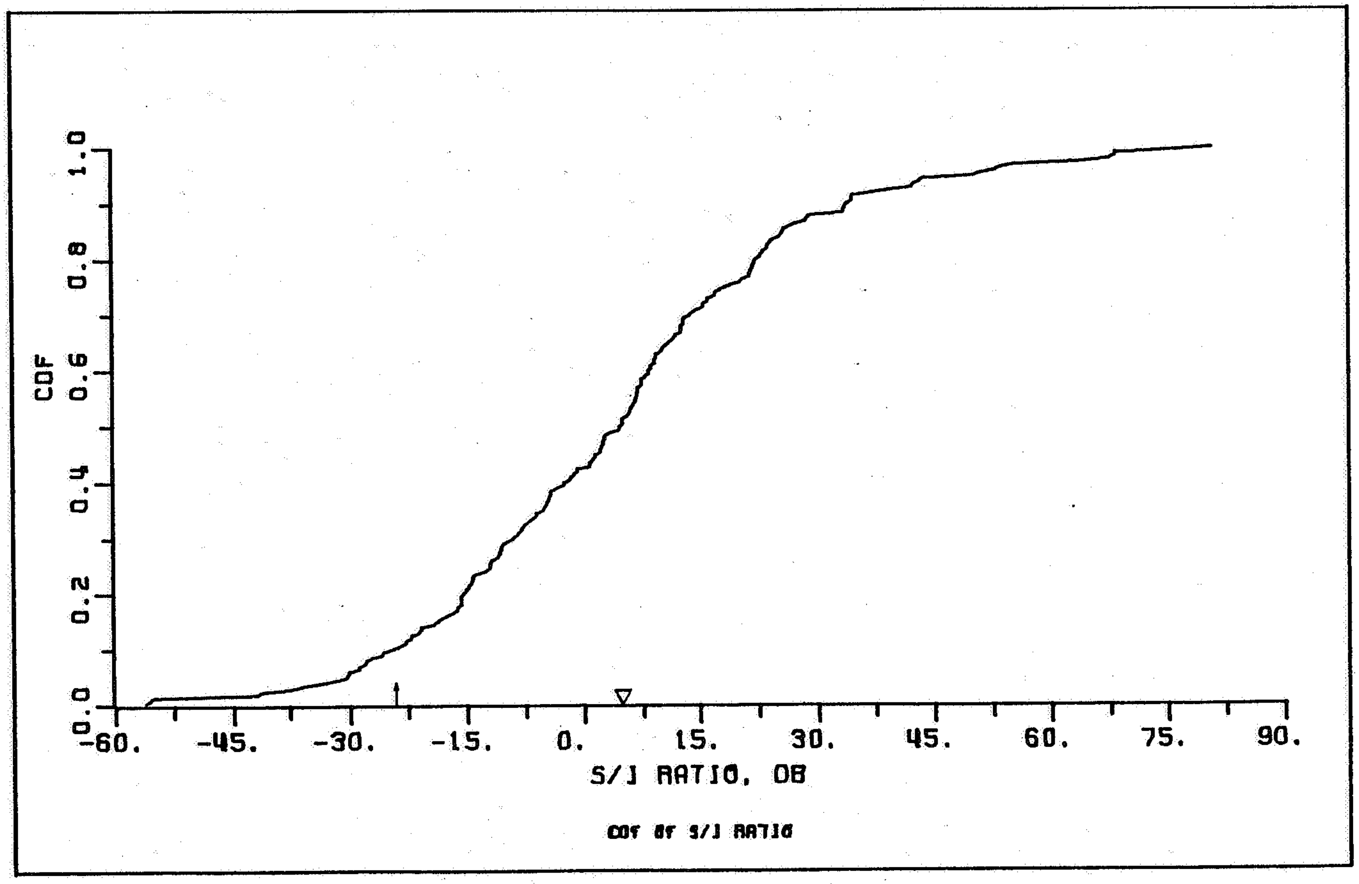

FIGURE C7
CUMULATIVE DISTRIBUTION OF R.F.
SIGNAL-TO-INTERFERENCE RATIO: 1 CHANNEL,
2 INTERFERERS

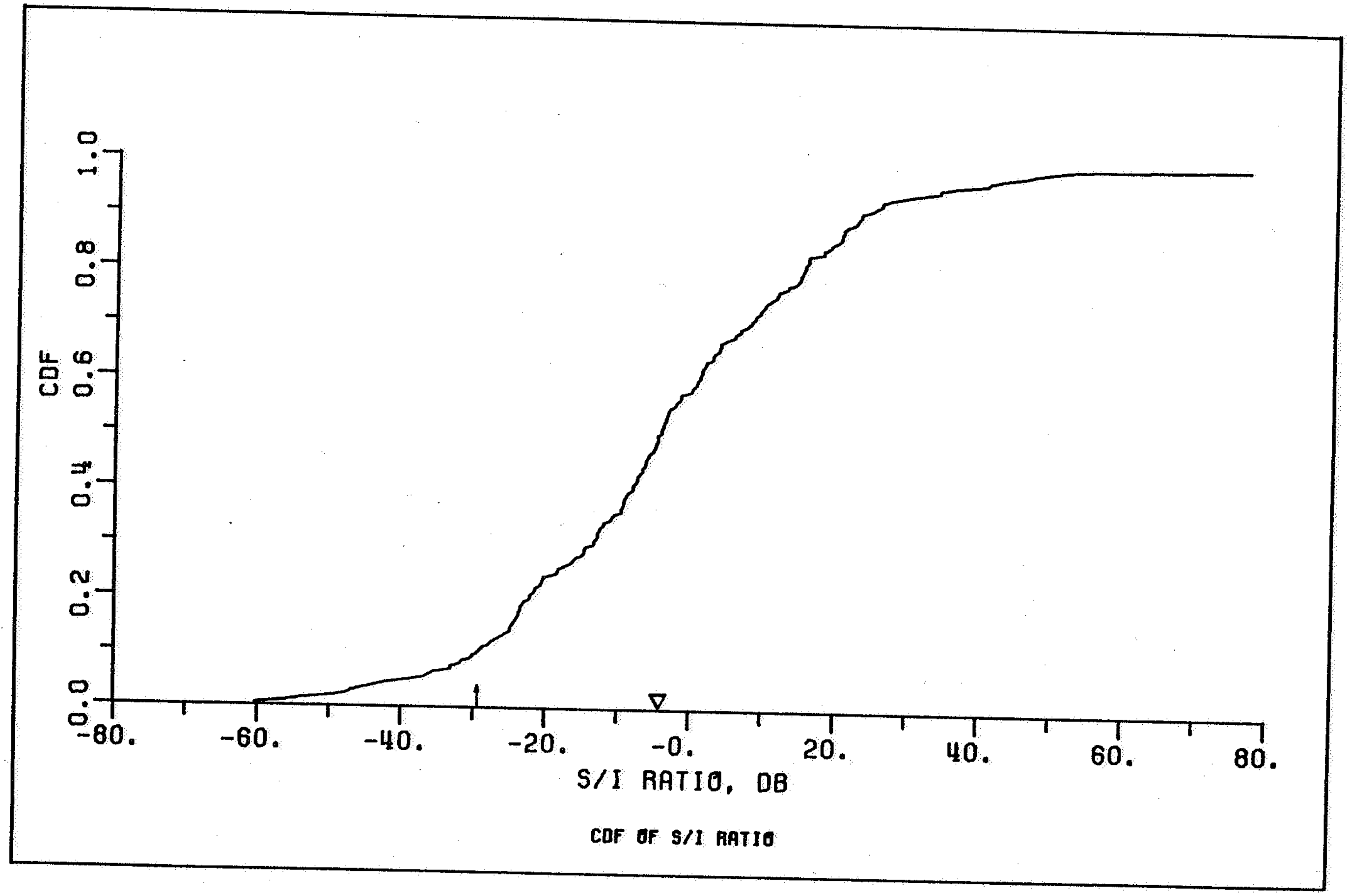

FIGURE C8
CUMULATIVE DISTRIBUTION OF R.F. SIGNAL-TO-INTERFERENCE RATIO: 1 CHANNEL, 4 INTERFERERS

interference ratio of 100 dB has been arbitrarily assigned. In practice, the difference between a 100 dB r.f. signal-to-interference ratio and an infinite r.f. signal-to-interference ratio will not be discernable to the listener. The system will become limited by noise and distortion long before this point. Note also that the program used to plot the CDFs automatically scales the abscissa according to the range of the data points. Thus, the abscissa scales on the various figures are generally different.

One result of the runs made on this model is that a single-channel system with two interferers (Figure C-7) and a 100-channel system with 200 frequency-hopping interferers (Figure C-15) yield roughly the same signal-to-interference statistics (in the absence of splatter), although the FH statistics suggest slightly interior performance. On the basis of such a comparison, some degradation should be expected for the FH system, since the instantaneous channel loadings for FH are not uniform, as with N equally loaded conventional channels.

From Figures C-9 through C-15, it is possible to plot the probability of attaining a 10 dB S/I ratio as a function of the number of active interferers. Such a plot is shown in Figure C-16. Also shown on this figure are the corresponding points for the single-channel simulation results. In both cases, it is possible to interpolate between the data points using piecewise linear segments. The largest residual for either case is about .04. Of course, the continuous line segments are meaningful only for integer values of M (the number of interferers). Extrapolations of these segments beyond the data points would seem to be a conservative procedure, since the probability of achieving a 10 dB C/I ratio must approach zero asymptotically as M increases. The extrapolations, on the other hand, show a zero probability of achieving he desired performance at M = 35 (for one channel) and M = 640 (for 100 channels). The cost of replacing the extrapolations with simulation results was not judged to be justified, since the cost of the simulation rises rapidly for large numbers of interferers if accurate estimates of low-probability events are required.

A simple traffic model (Poisson arrivals, exponential message durations) can now be used to predict the probability of achieving a 10 dB C/I ratio as a function of traffic intensity. Recognizing that the statement "$10 \le C/I$" has no meaning unless there is at least one call in progress or about to be made:

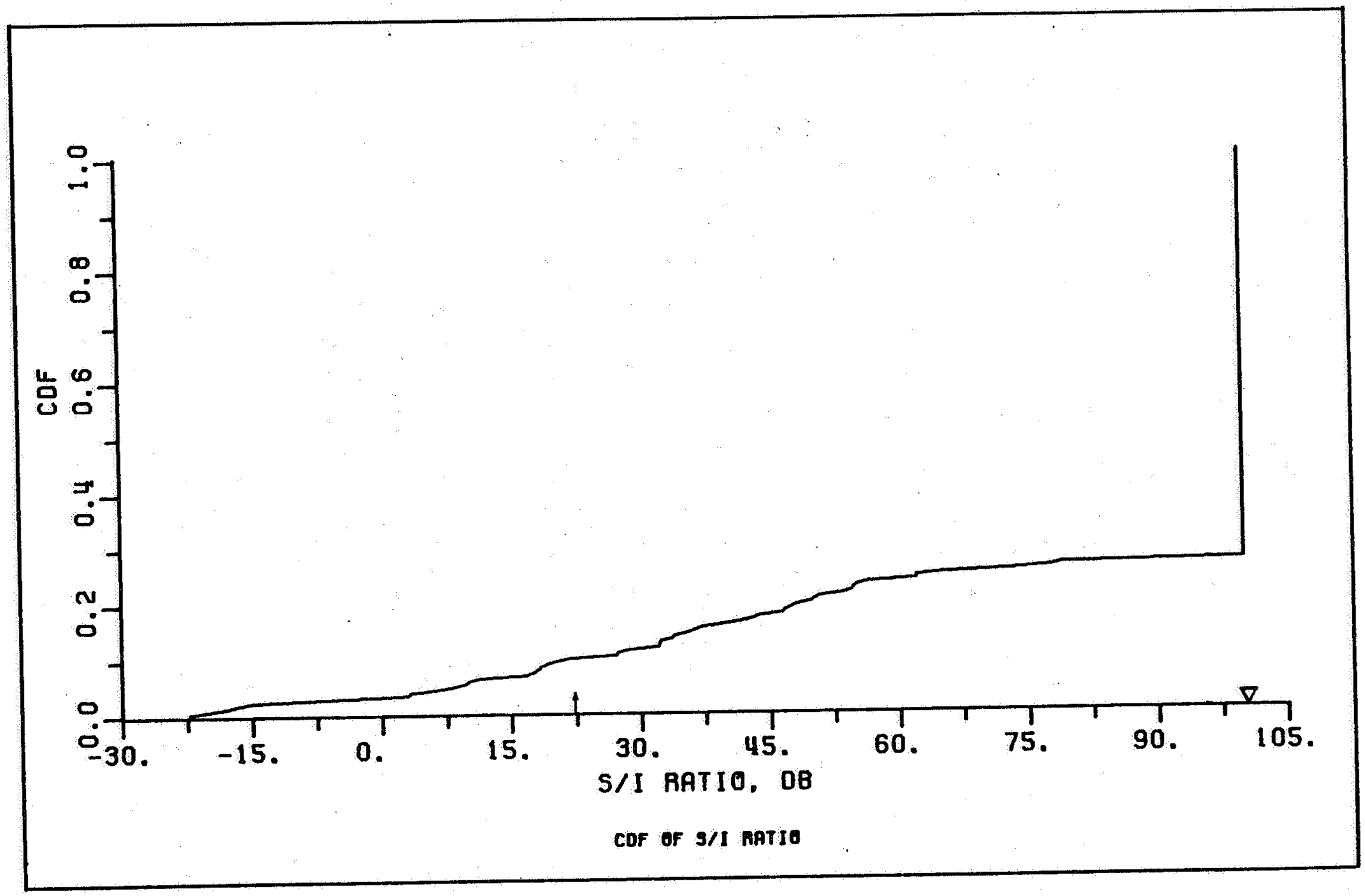

FIGURE C9
CUMULATIVE DISTRIBUTION OF R.F. SIGNAL-TO-INTERFERENCE RATIO: 100 CHANNELS, 5 INTERFERERS

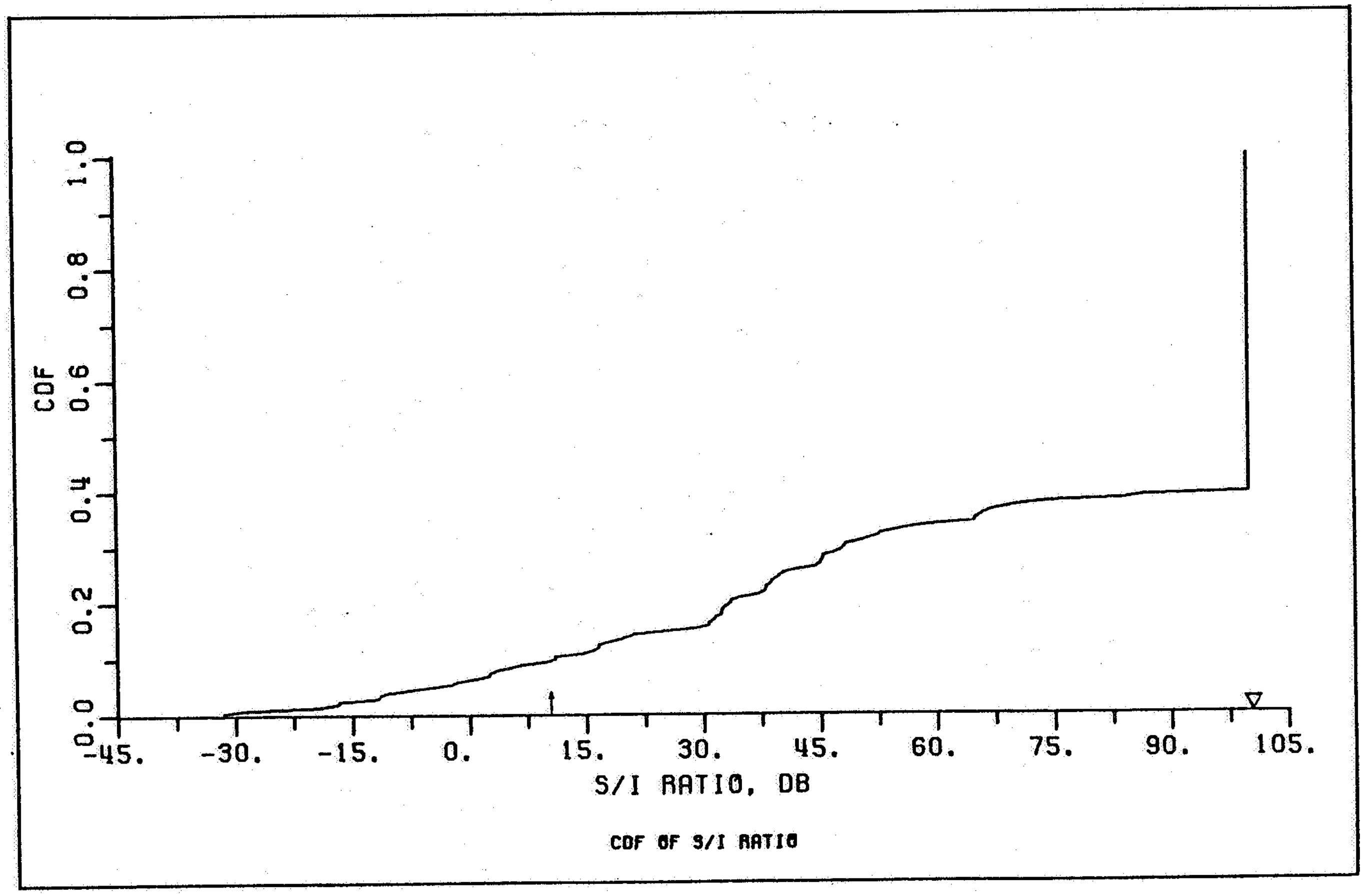

FIGURE C10
CUMULATIVE DISTRIBUTION OF R.F.
SIGNAL-TO-INTERFERENCE RATIO: 100 CHANNELS,
10 INTERFERERS

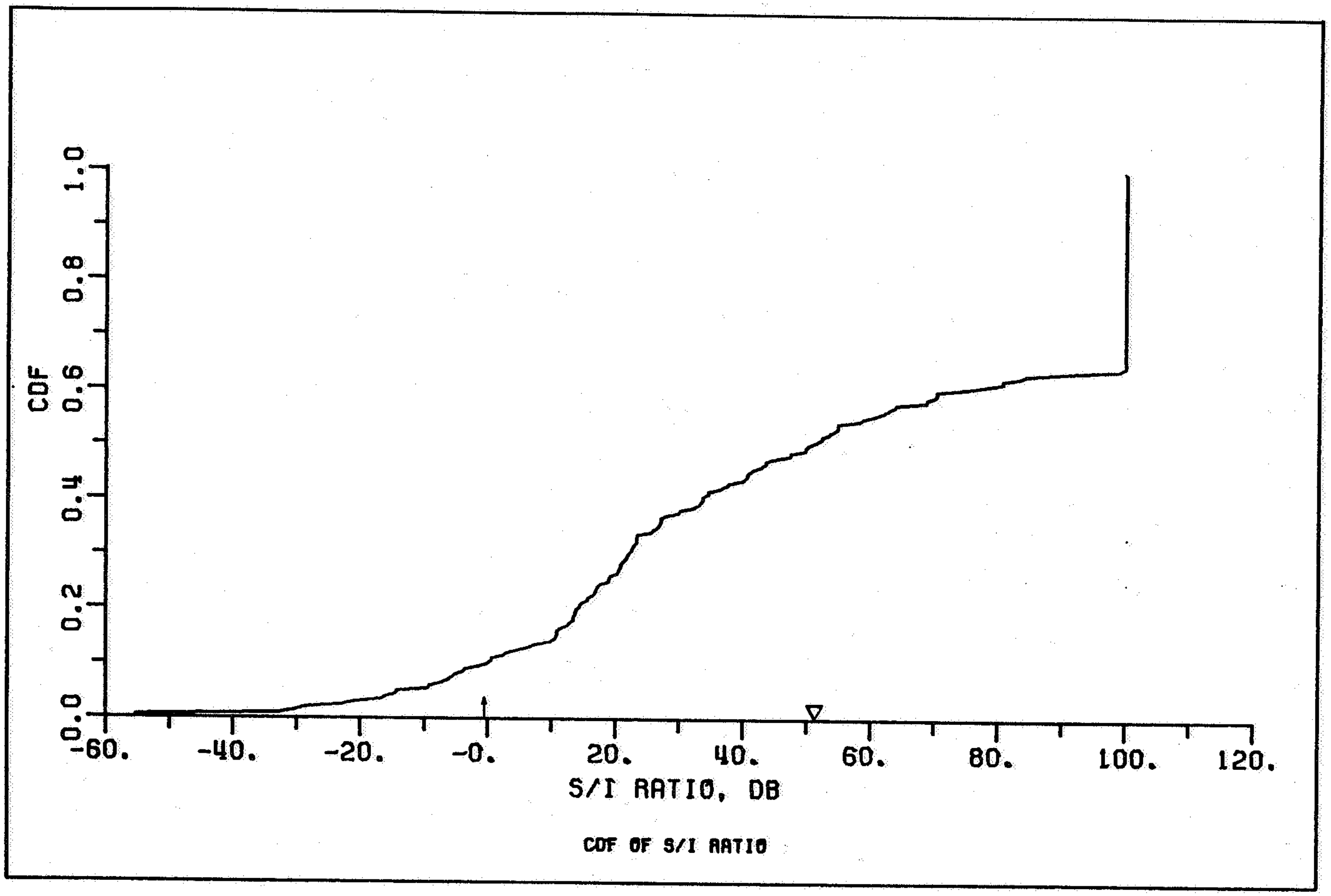

FIGURE C11
CUMULATIVE DISTRIBUTION OF R.F.
SIGNAL-TO-INTERFERENCE RATIO: 100 CHANNELS,
20 INTERFERERS

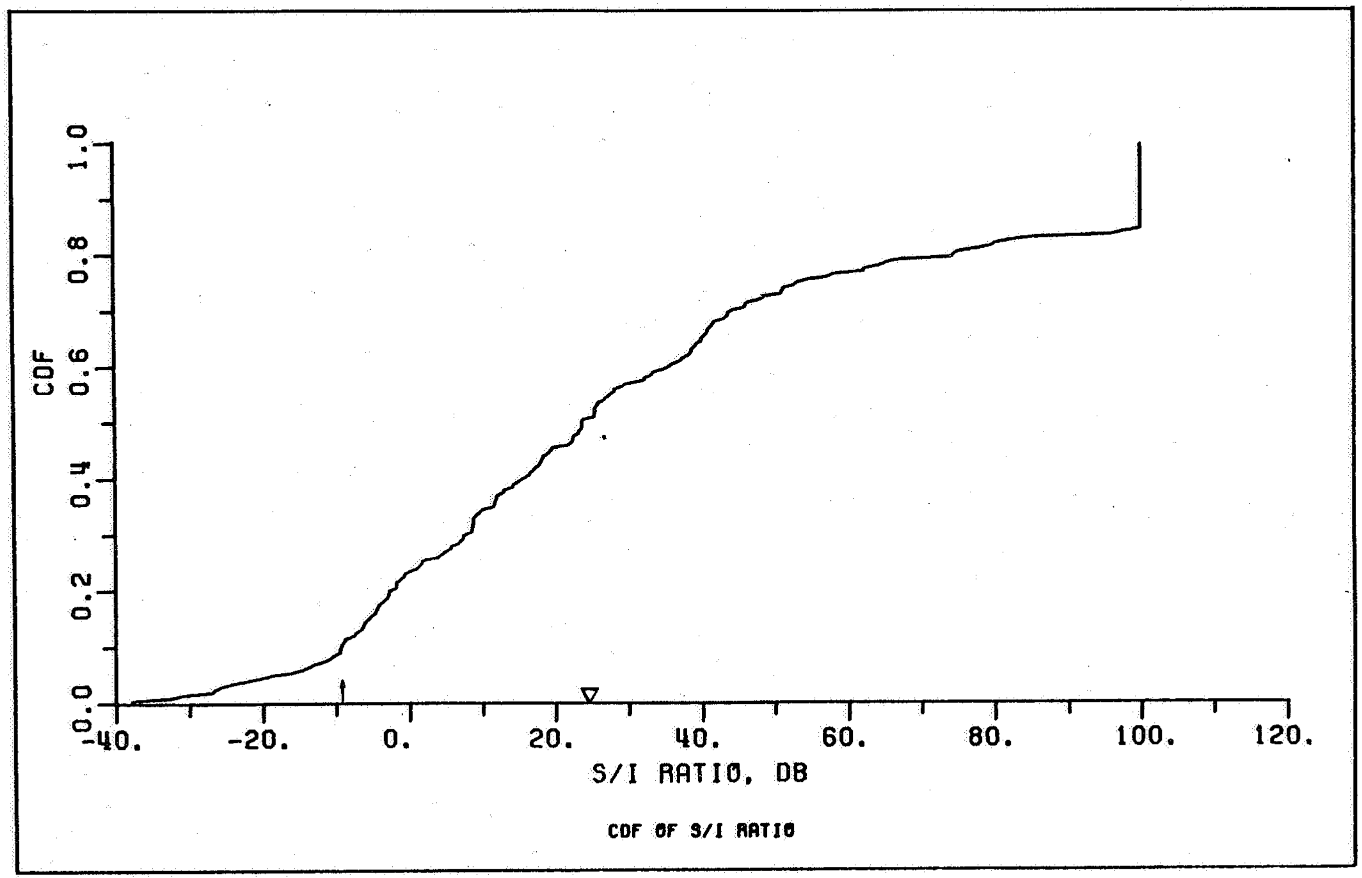

FIGURE C12
CUMULATIVE DISTRIBUTION OF R.F.
SIGNAL-TO-INTERFERENCE RATIO: 100 CHANNELS,
40 INTERFERERS

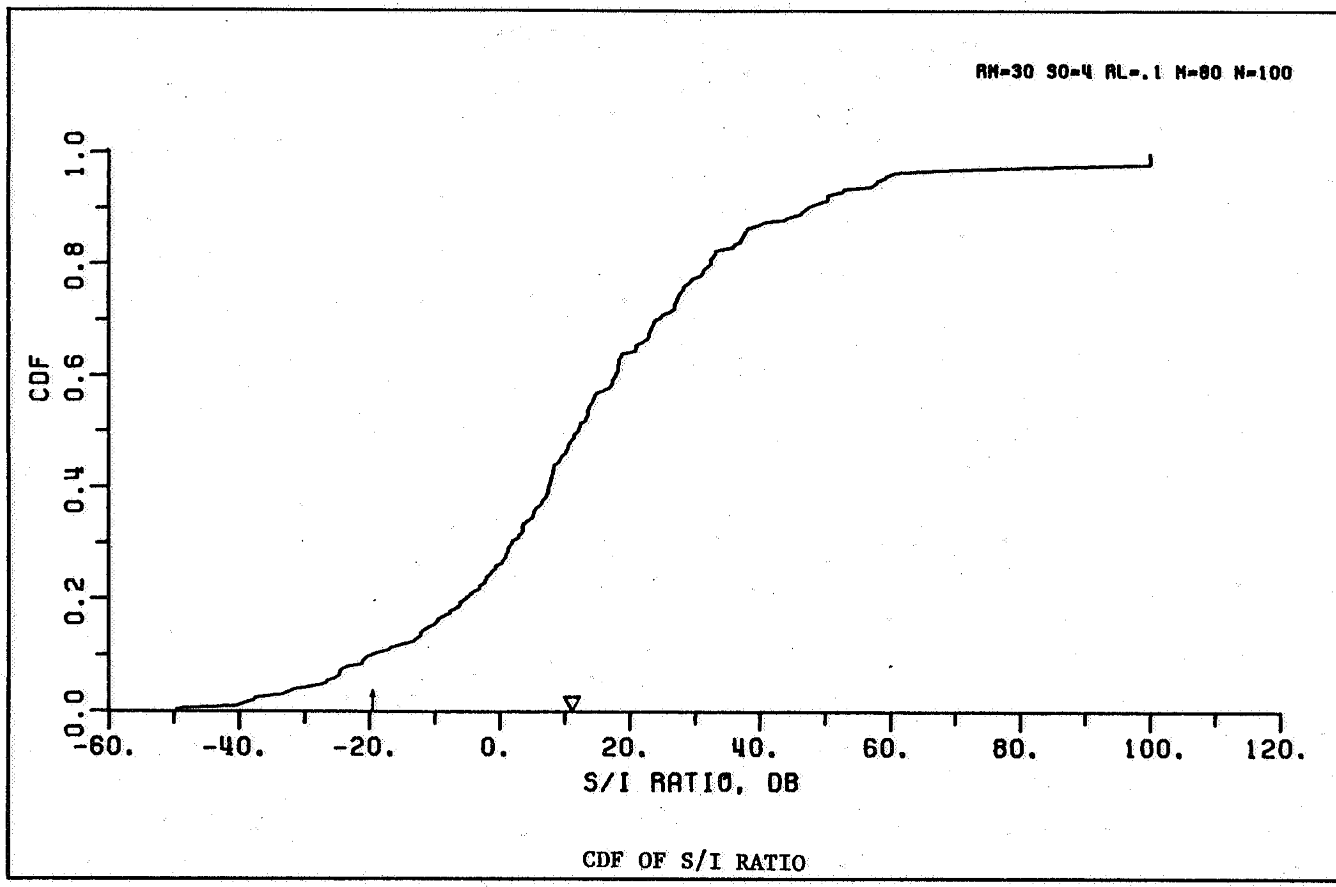

FIGURE C13
CUMULATIVE DISTRIBUTION OF R.F.
SIGNAL-TO-INTERFERENCE RATIO: 100 CHANNELS,
80 INTERFERERS

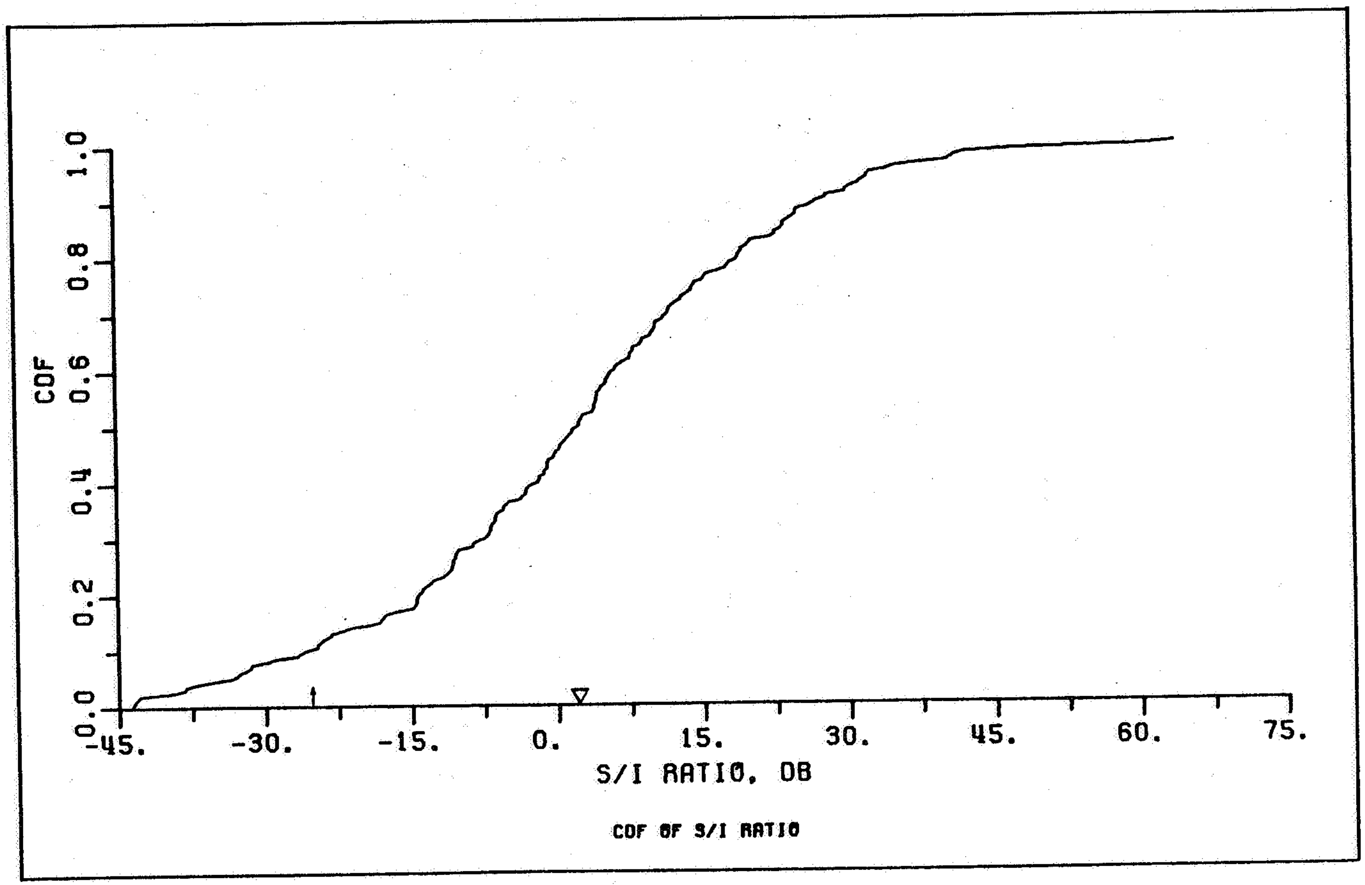

FIGURE C14
CUMULATIVE DISTRIBUTION OF R.F.
SIGNAL-TO-INTERFERENCE RATIO: 100 CHANNELS,
160 INTERFERERS

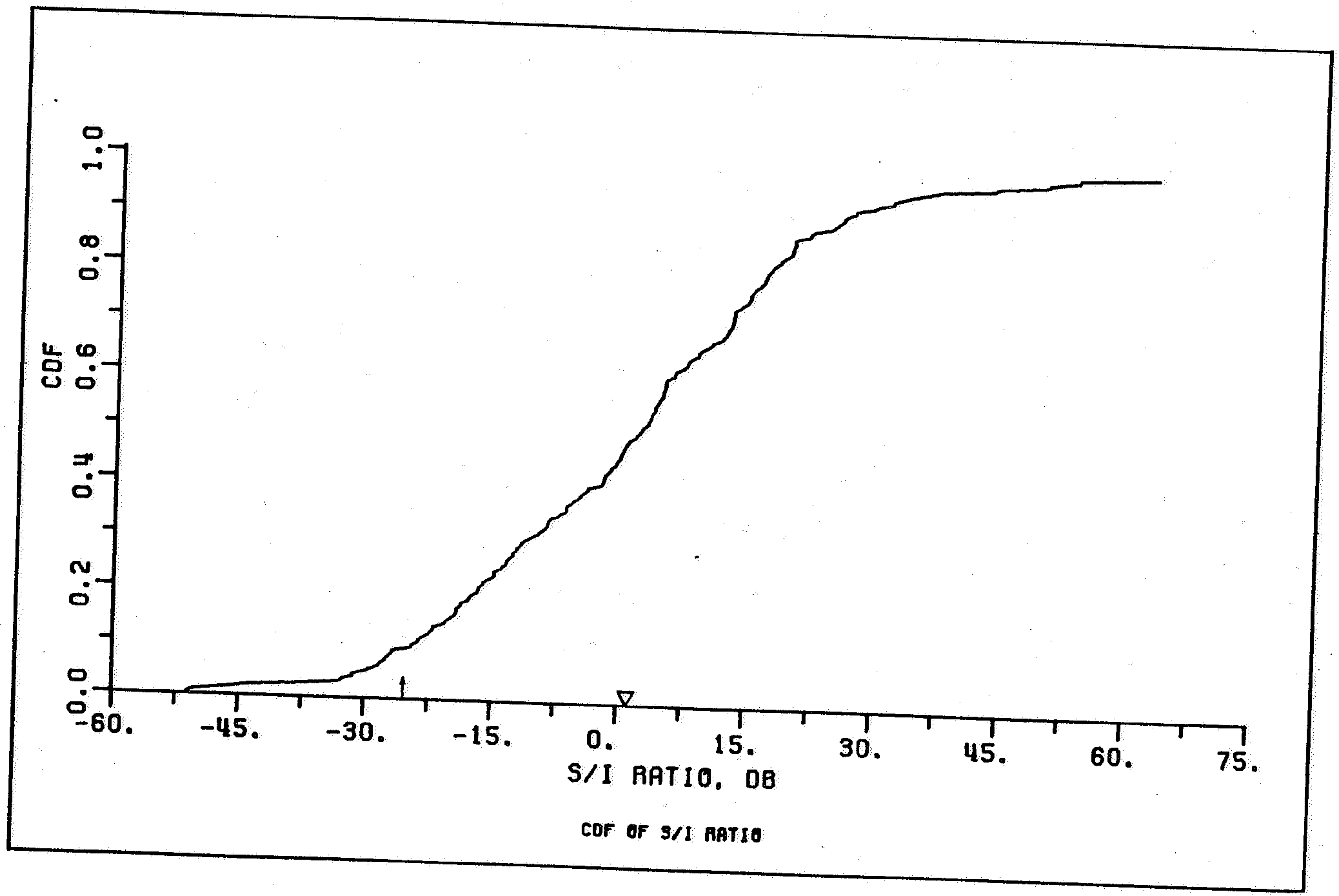

FIGURE C15
CUMULATIVE DISTRIBUTION OF R.F. SIGNAL-TO-INTERFERENCE RATIOS: 100 CHANNELS, 200 INTERFERERS

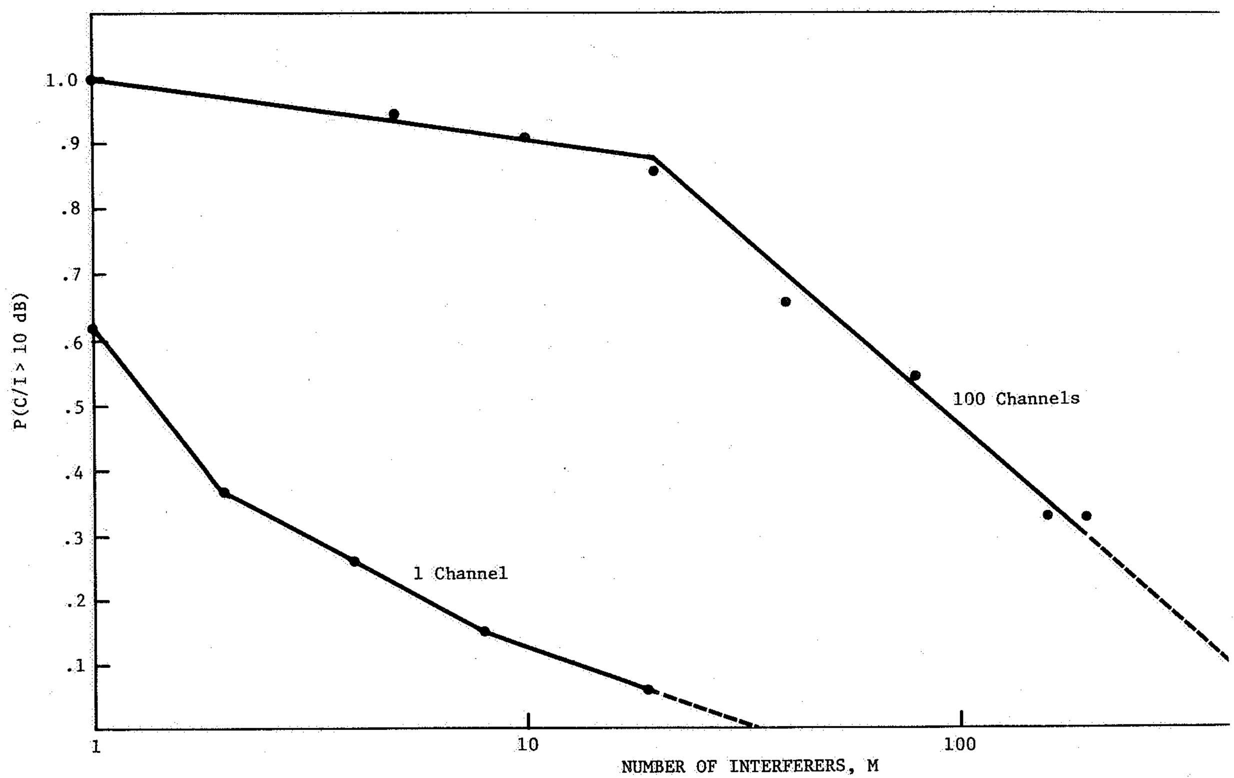

FIGURE C16
PROBABILITY OF ACHIEVING A 10 dB CARRIER-TO-INTERFERENCE RATIO WITH M INTERFERERS

$$P[10 < C/I] = \sum_{M=0}^{\infty} P[10 < C/I|M]\, P[M]$$

$$= \sum_{M=0}^{\infty} P[10 < C/I|M] \left[\frac{\rho^{M} e^{-\rho}}{M!}\right]$$

where ρ = traffic intensity, Erlangs

The results of inserting the simulation-produced estimates of p(10 < C/I) into this traffic model are shown in Figures C-17 and C-18 for single-channel and 100-channel operation, respectively. It can be seen that, for a 90 percent probability of achieving a 10 dB C/I ratio, the maximum traffic intensity is about 0.25 Erlangs for a single channel and 12.5 Erlangs for a 100-channel frequency hopping system having the characteristics described above. Thus, the channel-by-channel version (i.e., nonfrequency hopping) would appear to be about twice as efficient in its use of the spectrum as the FH version, based on the simple models used here. This comparison embodies the implicit assumption that the channels in the conventional system are loaded equally, on the average. Constraints that cause the channels to be loaded unequally (due to regulatory factors, for example) may shift the balance in favor of the FH approach. In addition, it wll be shown in the next section that the spectrum efficiency of the FH system improves significantly when the mean attempted communication distance is reduced.

C.4.2 Intelligibility Model

The intelligibility model incorporates a number of refinements that allow the impact of parameter changes to be assessed. Specifically, it provides for the hopping rate, the amount of "splatter," the mean communication distance, and the link type (mobile-to-base or mobile-to-mobile) to be set for each simulation run. All runs are based on a 100-channel system.

C.4.2.1 Mobile-to-Base Operation

A tabulation of the principal mobile-to-base simulation runs and their parameter settings is shown in Table C-2. The table is self-explanatory except for the column marked "splatter." The figures in this column represent the assumed attenuation on the adjacent and next-adjacent channels, respectively. For example, 60/80 means that an adjacent-channel signal is attenuated by 60 dB relative to a co-channel signal and that a signal two channels removed is attenuated 80 dB relative to a co-channel signal. Instead of trying to group the runs into sets having all but one parameter in common, the runs will simply be presented in chronological order.

TABLE C-2

LIST OF SIMULATION RUNS
(INTELLIGIBILITY MODEL)

FIGURE	NO. OF CHANNELS	NO. OF INTERFERERS	SPLATTER, DB ATTENUATION	MEAN COMM. DIST, KM	HOPPING RATE, /s	NO. OF TRIALS	LINK DIRECTION
C-19	100	5	10/30	8	1	200	MB
C-20	100	5	10/30	8	20	200	MB
C-21	100	20	10/30	8	20	200	MB
C-22	100	40	10/30	8	20	200	MB
C-23	100	20	10/30	4	20	100	MB
C-24	100	40	10/30	4	20	50	MB
C-25	100	40	200/200	4	20	50	MB
C-26	100	80	200/200	4	20	50	MB
C-27	100	120	200/200	4	20	50	MB
C-28	100	80	60/80	8	20	50	MB
C-29	100	80	60/80	4	20	50	MB
C-30	100	100	60/80	2	20	50	MB
C-31	100	130	60/80	2	20	50	MB
C-32	100	160	60/80	2	20	50	MB
C-33	100	20	60/80	16	20	50	MB
C-34	100	40	60/80	16	20	50	MB
C-35	100	60	60/80	16	20	50	MB
C-36	100	60	60/80	8	20	50	MB

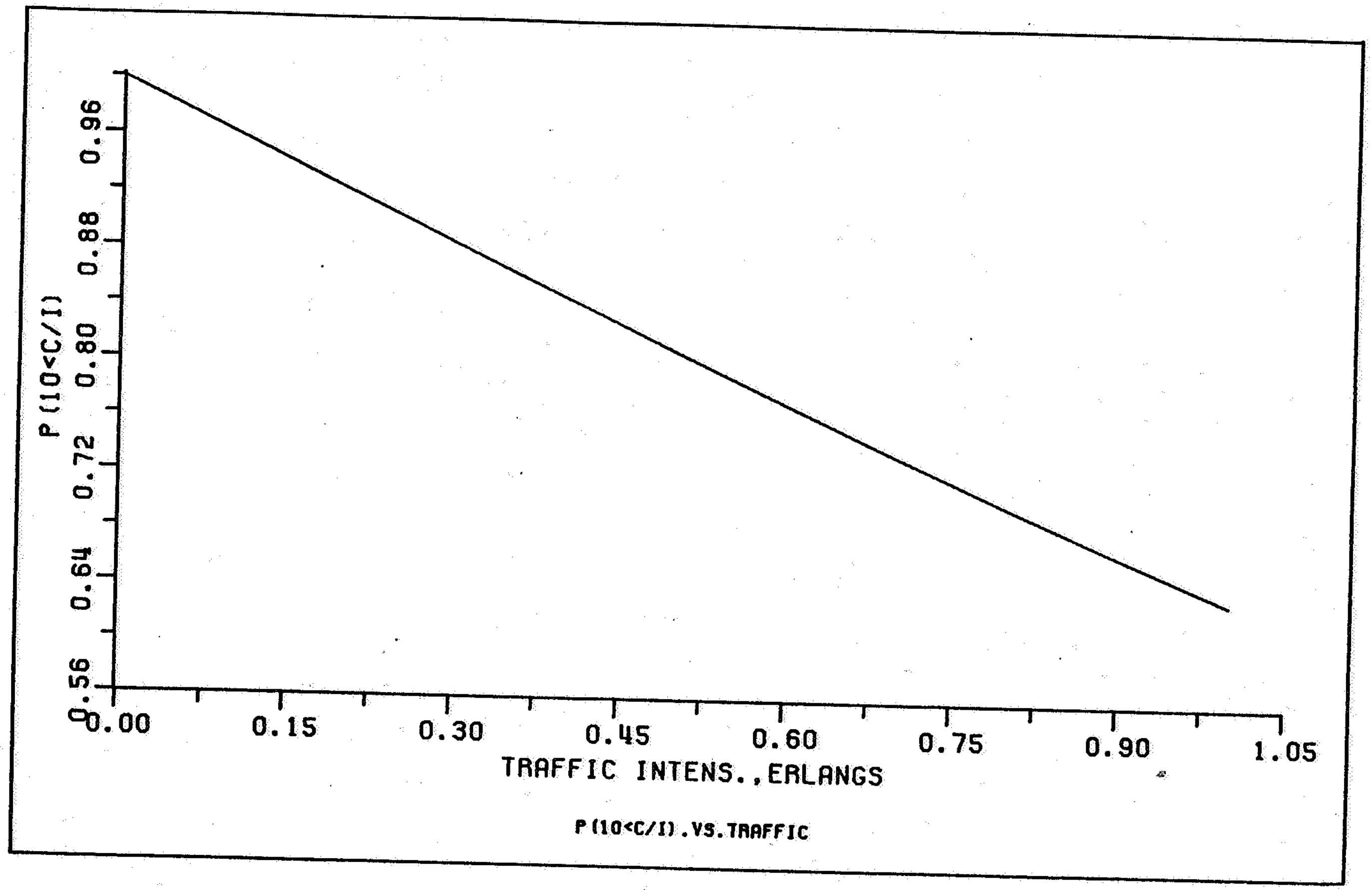

FIGURE C17
PROBABILITY OF ACHIEVING 10 dB C/I
AS A FUNCTION OF TRAFFIC INTENSITY,
1 CHANNEL

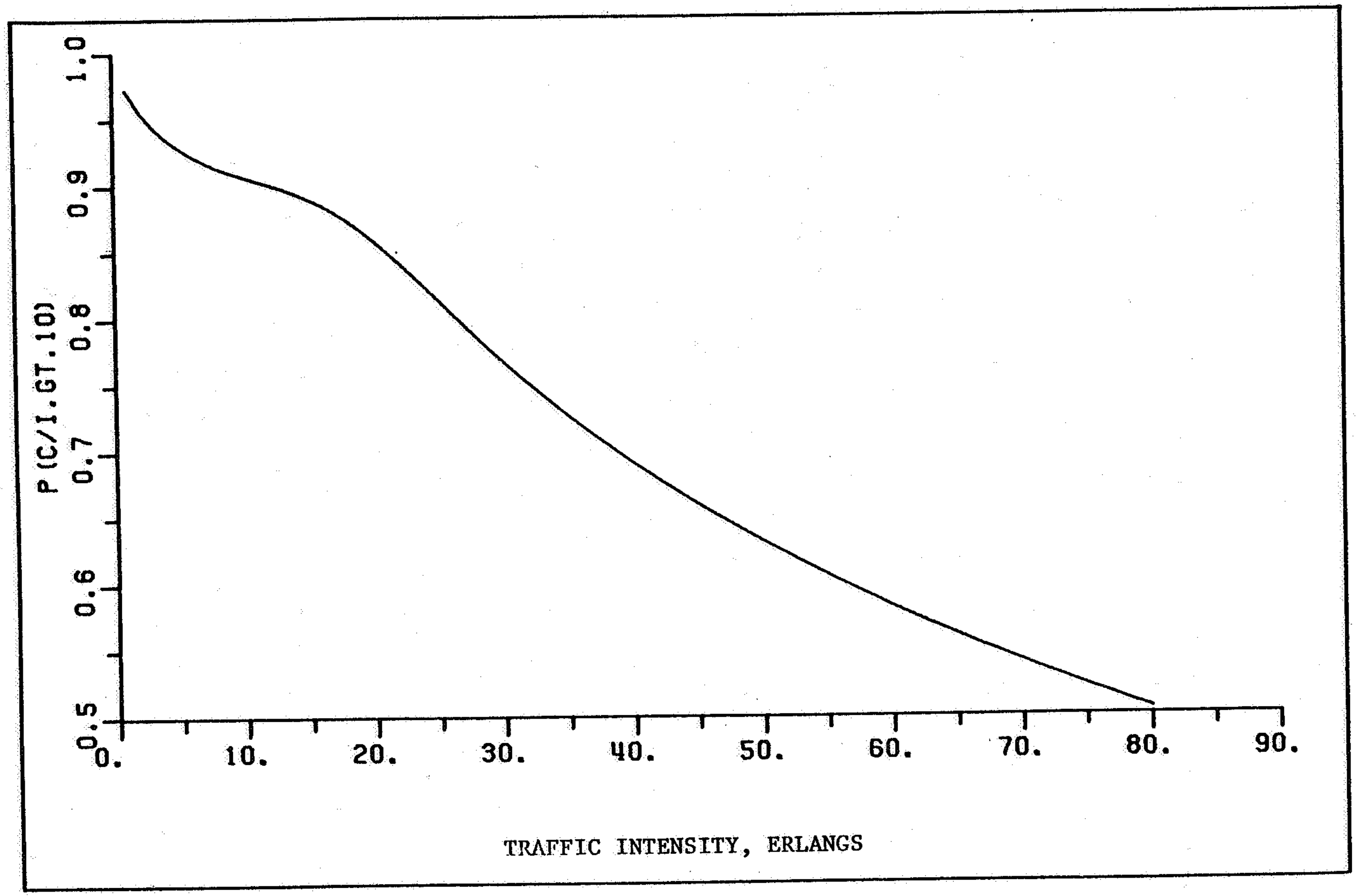

FIGURE C18
PROBABILITY OF ACHIEVING 10 dB C/I
AS A FUNCTION OF TRAFFIC INTENSITY,
100 CHANNELS

Figures C-19 and C-20 show the impact of raising the hopping rate from one per second (Figure C-19; this rate was also used in the signal-to-interference model) and twenty per second (Figure C-20; this rate is used for all of the simulation runs that follow). For both runs, the mean distance at which communication is attempted is 8.0 km (5 miles). The "splatter" attenuation is assumed to be 10 dB and 30 dB for adjacent and next-adjacent channels, respectively. This is representative of what can be achieved using channel spacings of 15 to 20 kHz with the least expensive two-pole monolithic crystal filters (see, for example, reference C-8).

It can be seen from Figures C-19 and C-20 that the impact of increasing the hopping rate from 1/s to 20/s is not great under these conditions. This is primarily because the rate of interruption is usually low with only 5 interferers, even at 20 hops/s. For low rates of interruption, the dependence of articulation on interruption rate is small.

Figure C-21 shows the result of increasing the number of interferers to 20 with all other parameters unchanged. Although the degradation of performance is clearly visible, there is still a 91 percent chance of achieving an articulation score of 0.75 or better.

If the number of interferers is increased to 40 (without changing any other parameters), the CDF of the word articulation score is as shown in Figure C-22. If the criterion for acceptable performance is a 90 percent chance of achieving a 0.75 word articulation score, then this run fails to yield acceptable performance, since the probability of achieving a word articulation score of 0.75 is only 0.79.

In an effort to determine the effect of the assumed 8 km mean attempted communication distance, this parameter was reduced to 4 km (2.5 miles) and the simulation was re-run with 20 and 40 interferers (other parameters, except for the number of trials, were unchanged from the conditions that produced Figures C-21 and C-22). The results are shown in Figures C-23 and C-24, respectively. Although some improvement is evident, the amount of improvement is not great. The probability of attaining an acceptable articulation score (0.75) is still below 90 percent for 40 interferers.

In Figure C-25, the "splatter" parameters have been changed to show the effect of substantially removing splatter. This accomplished by setting adjacent-channel and next-adjacent channel attenuation to 200 dB. Since the maximum possible interference-to-signal ratio for a single interferer (interferer 0.1 km from

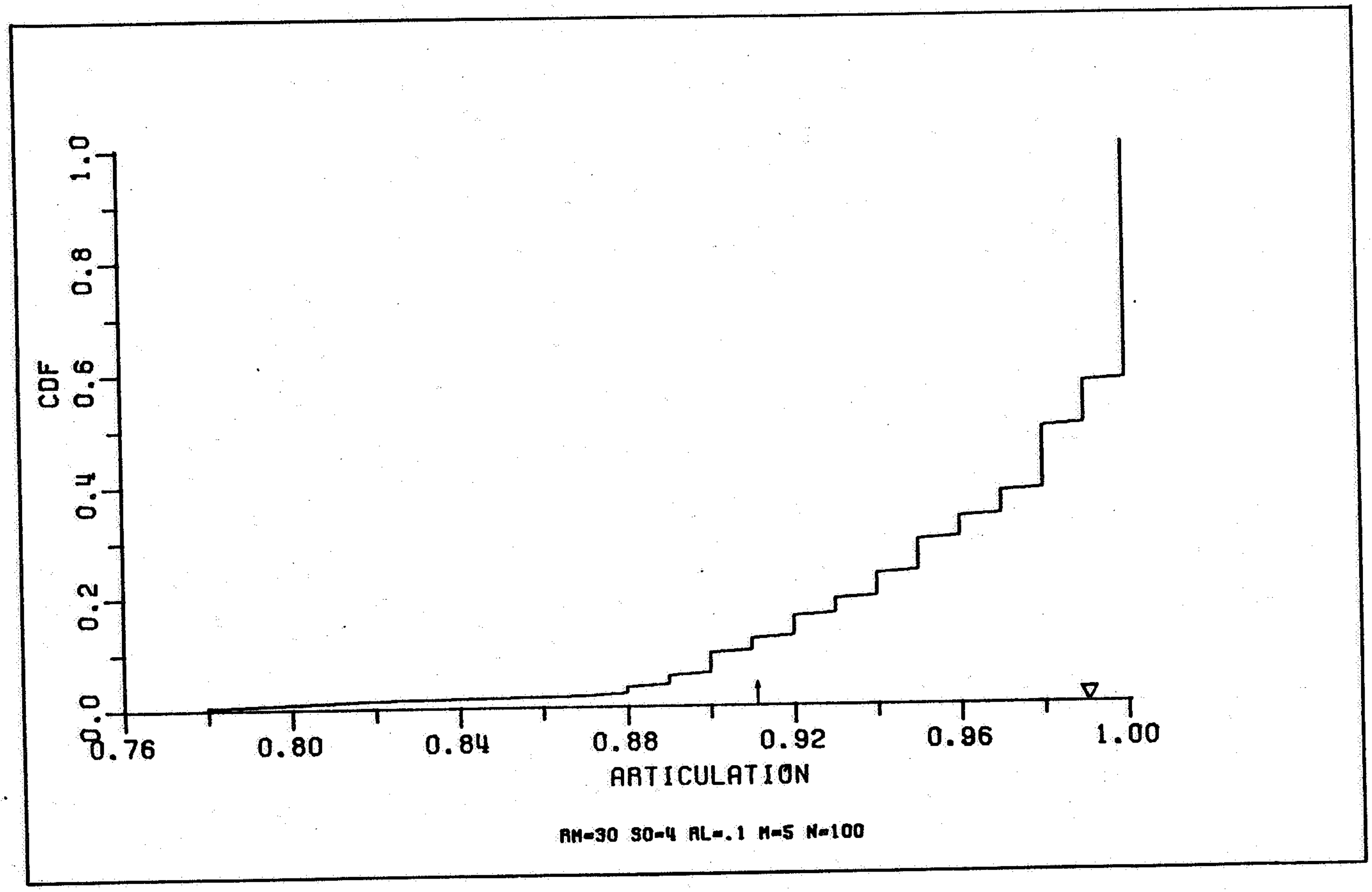

FIGURE C19
CUMULATIVE DISTRIBUTION OF PB WORD ARTICULATION: 5 INTERFERERS, 10/30 SPLATTER, 8 KM MEAN COMM. DIST., 1 HOP/S, 200 TRIALS, MOBILE-TO-BASE

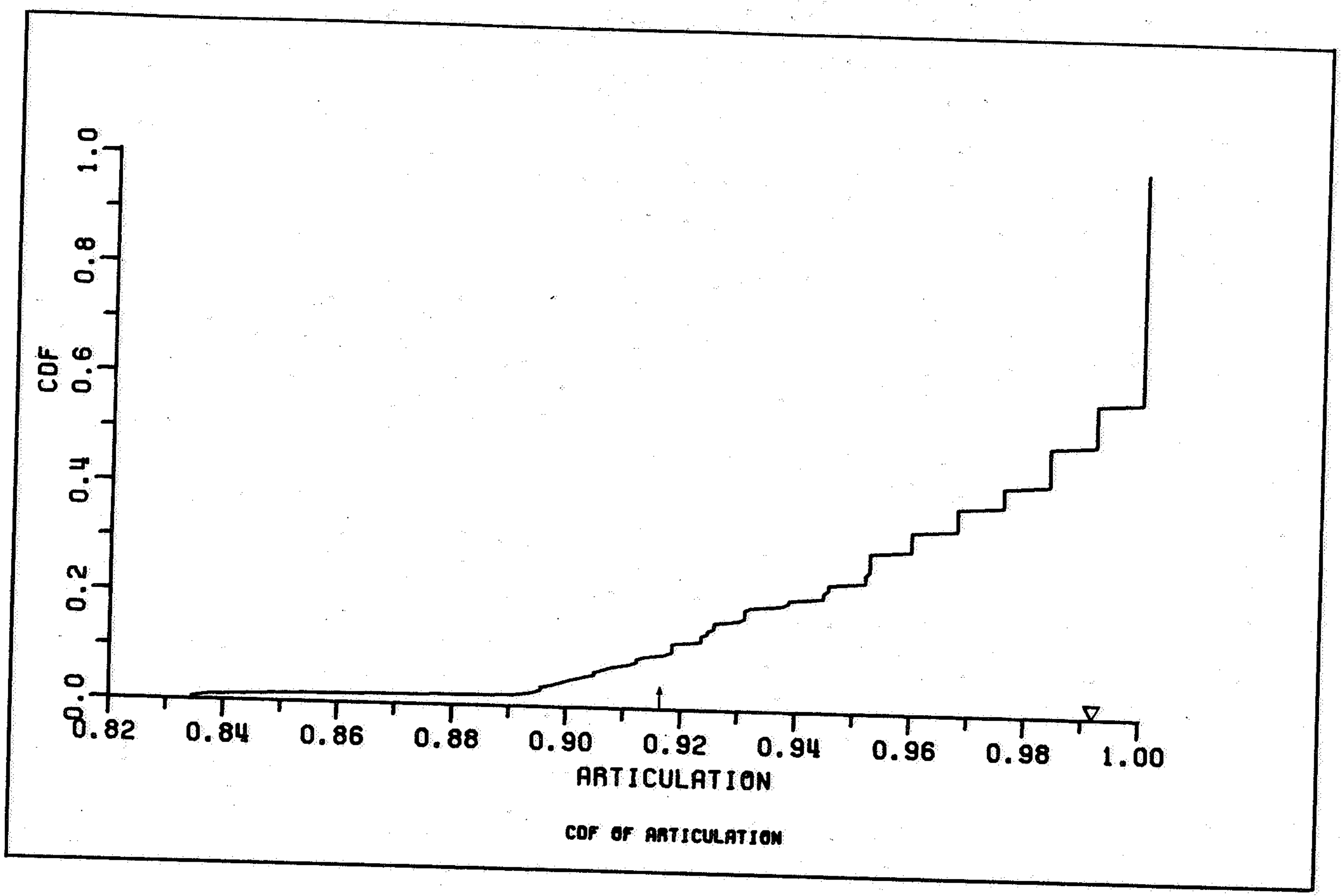

FIGURE C20
CUMULATIVE DISTRIBUTION OF PB WORD ARTICULATION: 5 INTERFERERS, 10/30 SPLATTER, 8 KM MEAN COMM. DIST., 20 HOPS/S, 200 TRIALS, MOBILE-TO-BASE

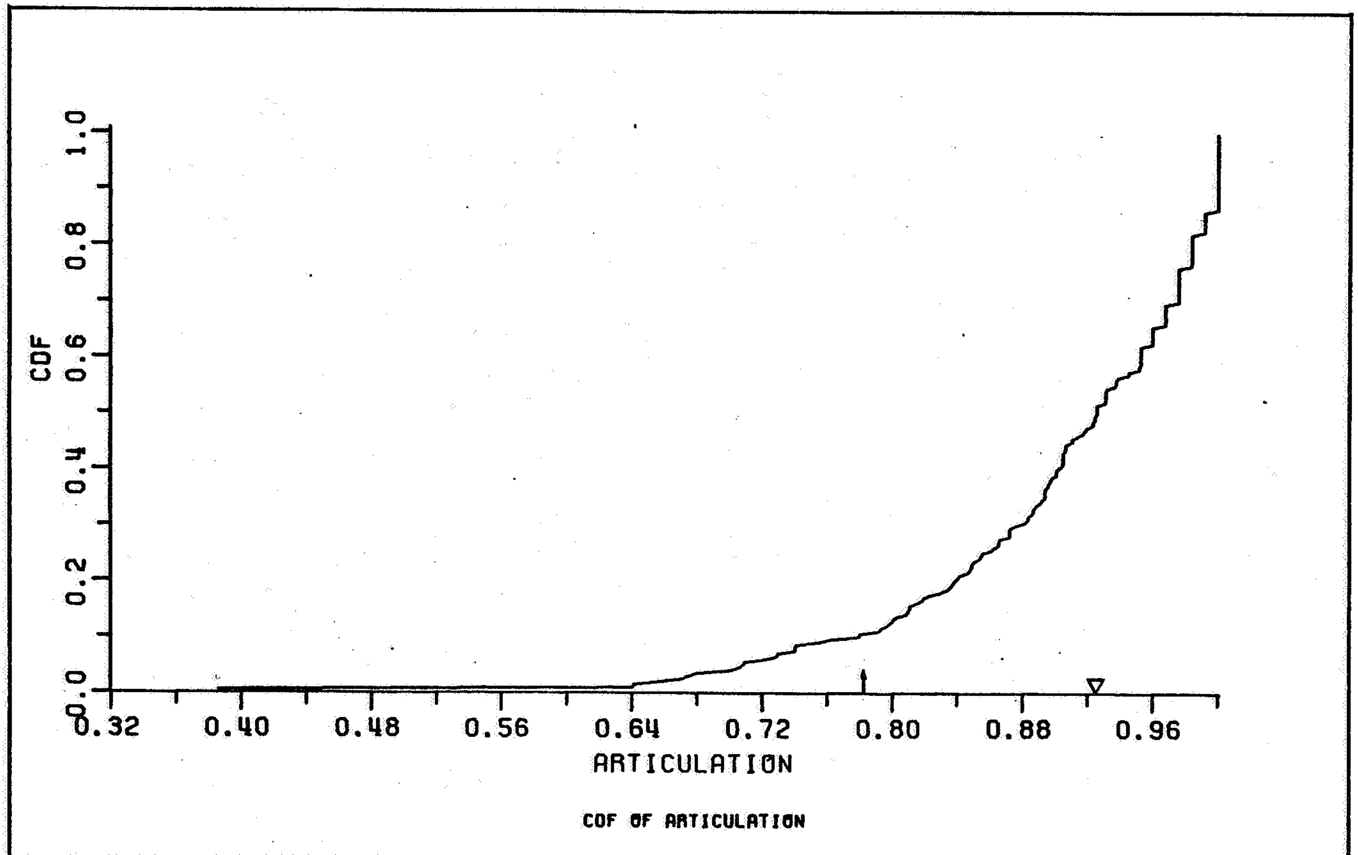

FIGURE C21
CUMULATIVE DISTRIBUTION OF PB WORD ARTICULATION: 20 INTERFERERS, 10/30 SPLATTER, 8 KM MEAN COMM. DIST., 20 HOPS/S, 200 TRIALS, MOBILE-TO-BASE

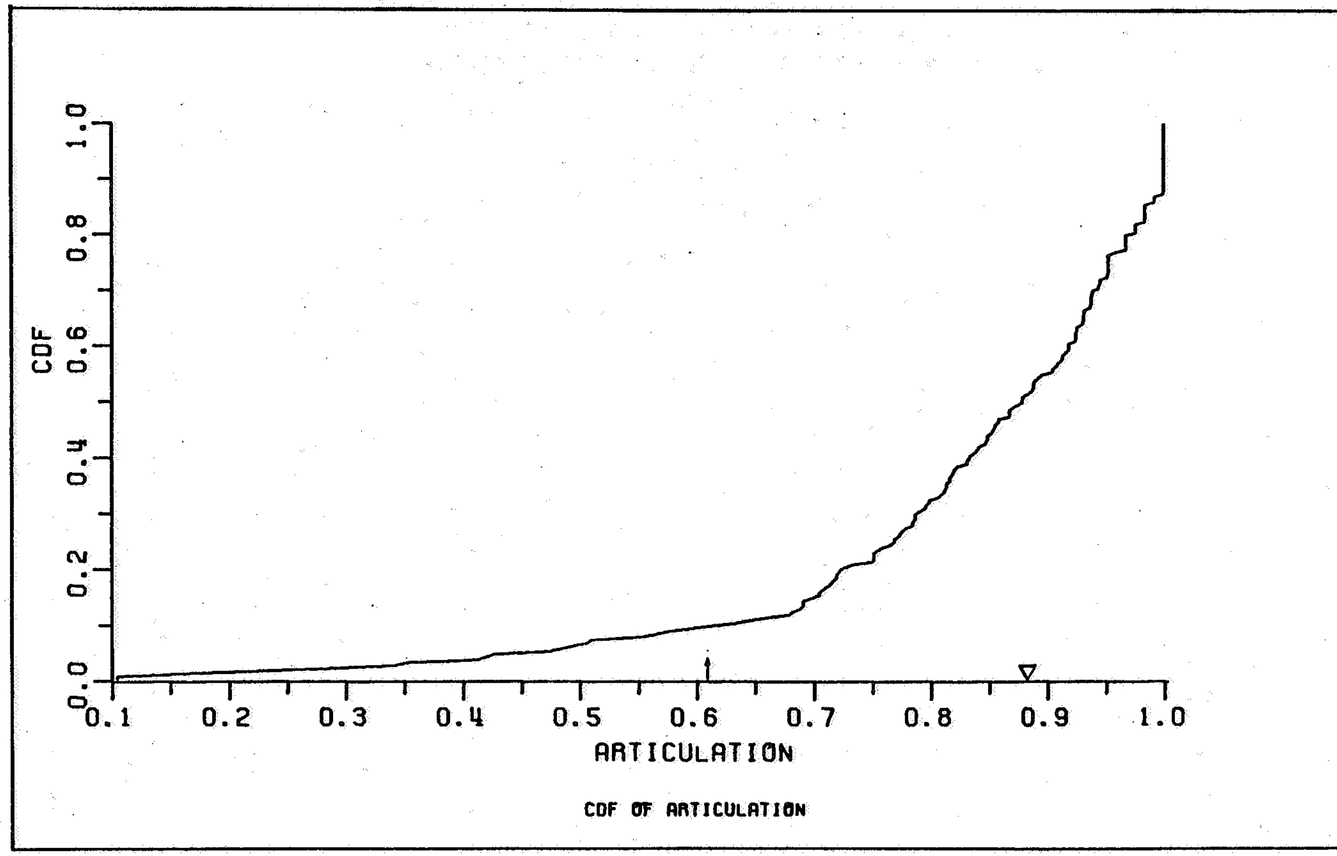

FIGURE C22
CUMULATIVE DISTRIBUTION OF PB WORD ARTICULATION: 40 INTERFERERS, 10/30 SPLATTER, 8 KM MEAN COMM. DIST., 20 HOPS/S, 200 TRIALS, MOBILE-TO-BASE

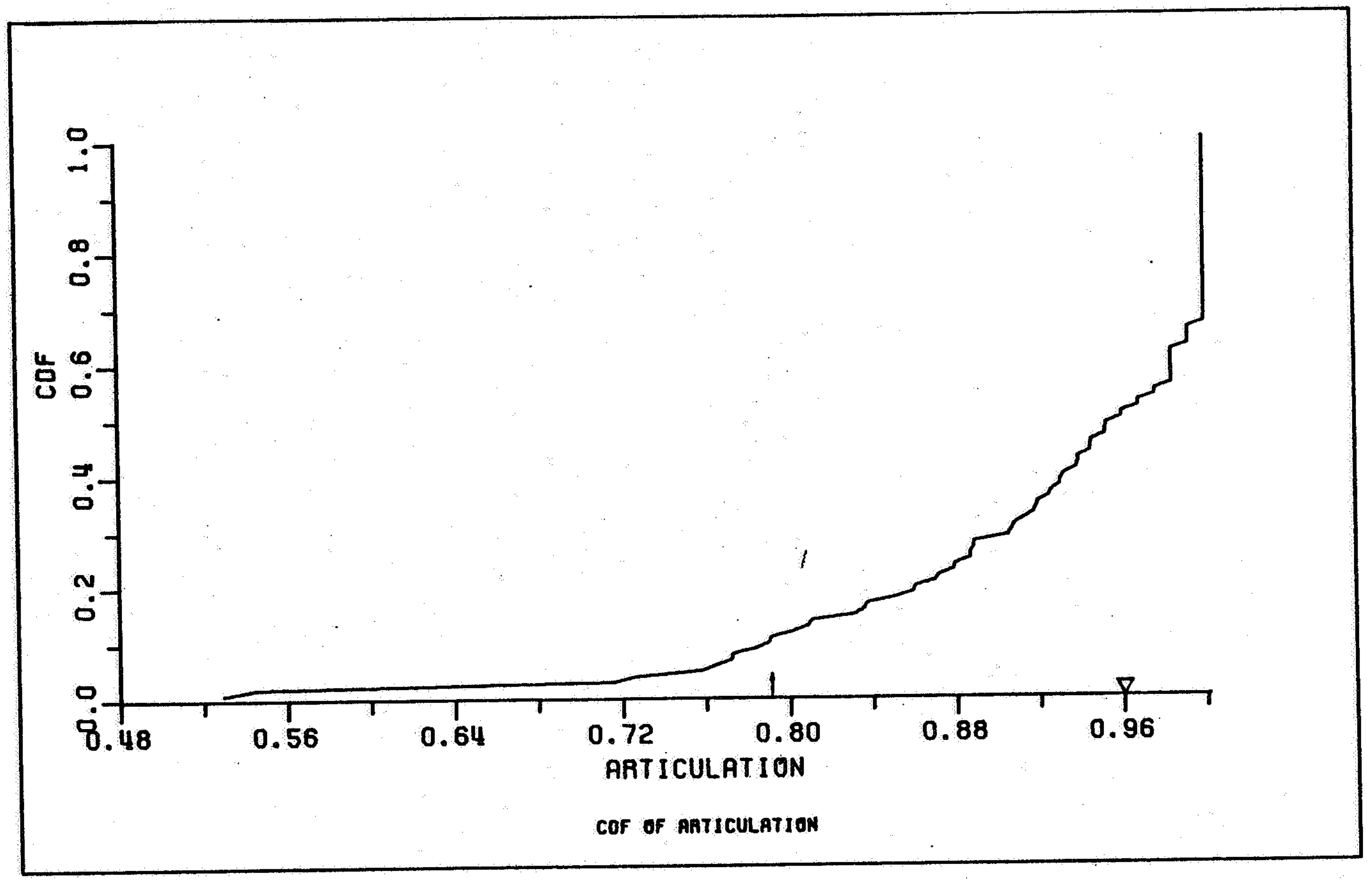

FIGURE C23
CUMULATIVE DISTRIBUTION OF PB WORD ARTICULATION: 20 INTERFERERS, 10/30 SPLATTER, 4 KM MEAN COMM. DIST., 20 HOPS/S, 100 TRIALS, MOBILE-TO-BASE

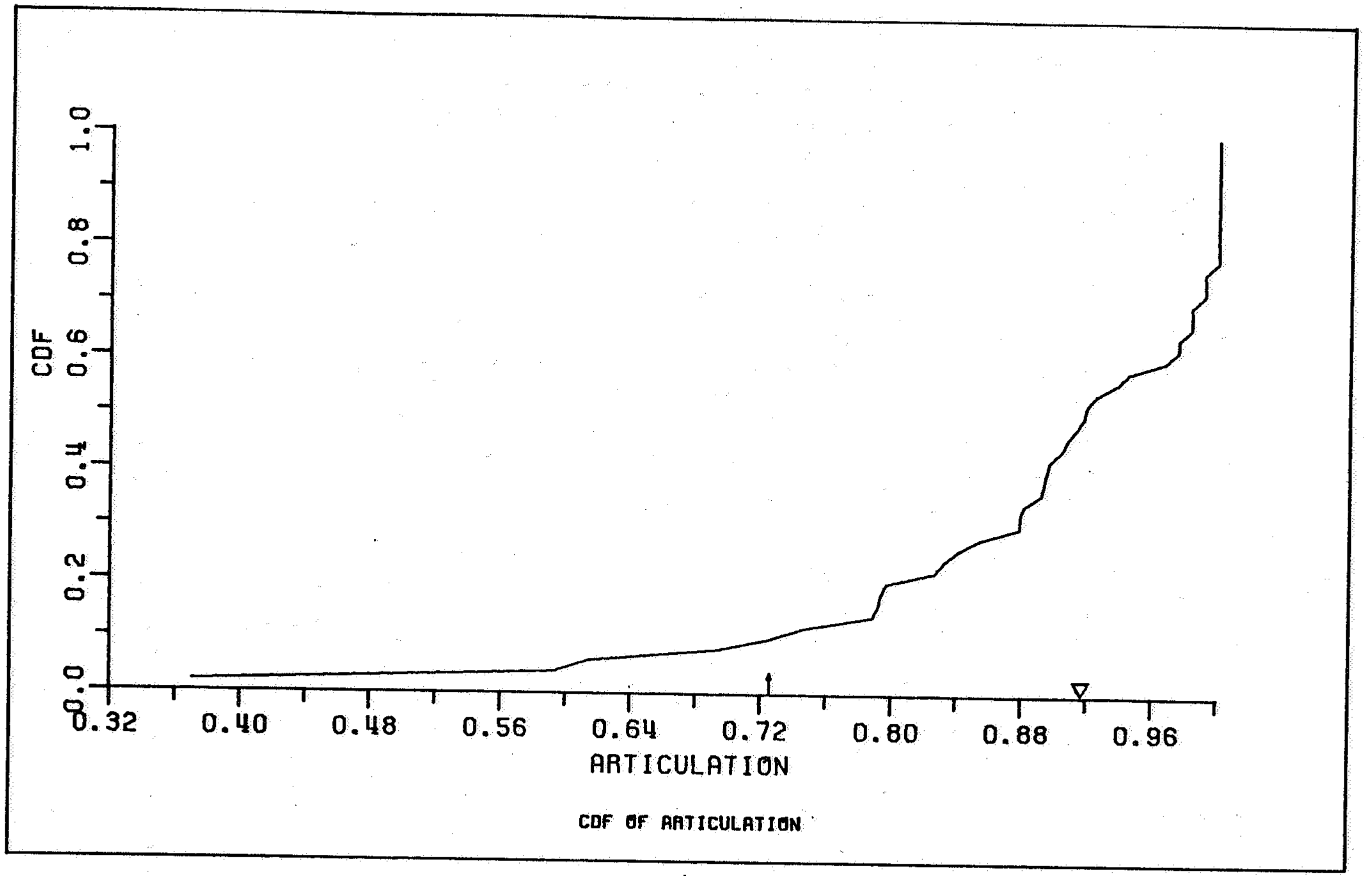

FIGURE C24
CUMULATIVE DISTRIBUTION OF PB WORD ARTICULATION: 40 INTERFERERS, 10/30 SPLATTER, 4 KM MEAN COMM. DIST., 20 HOPS/S, 50 TRIALS, MOBILE-TO-BASE

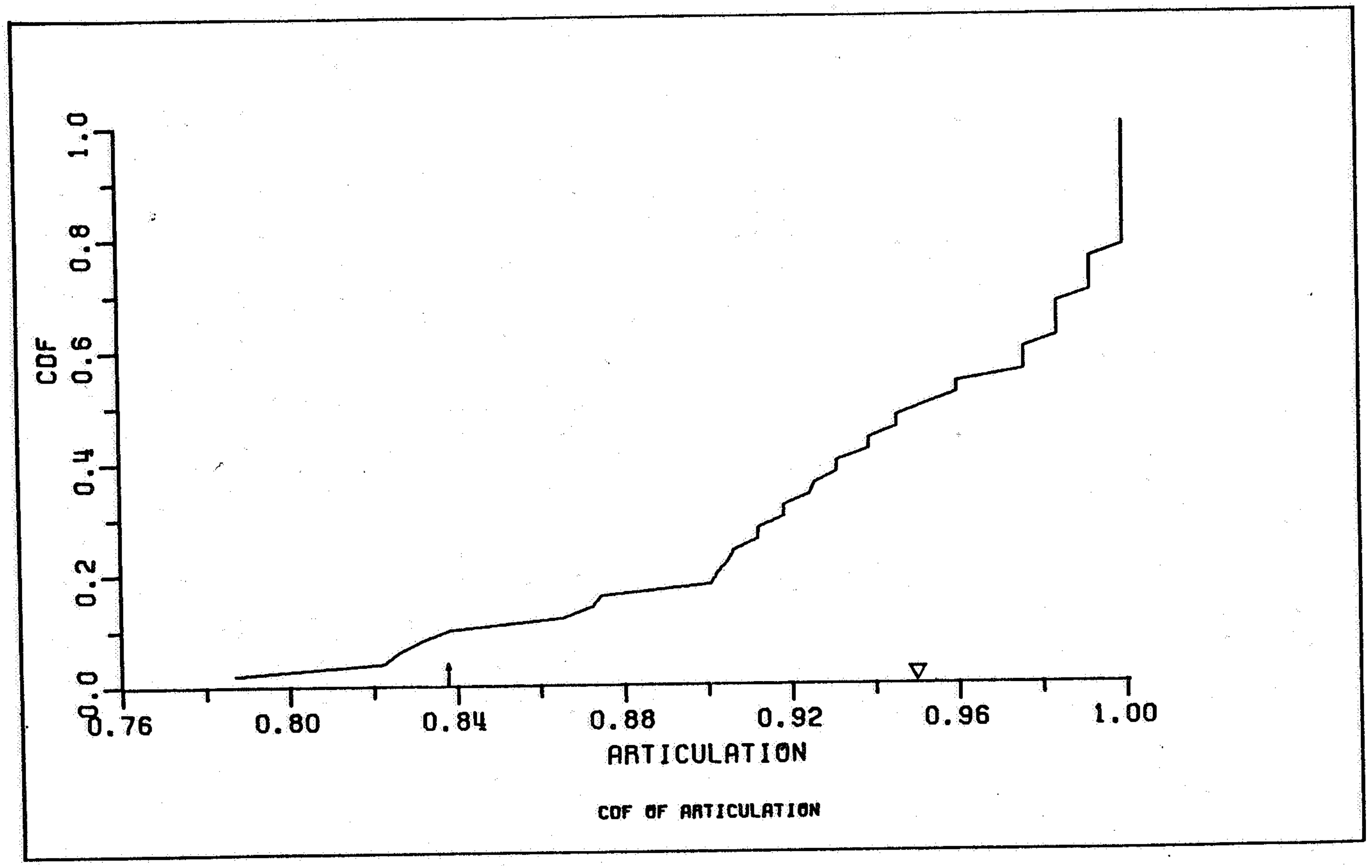

FIGURE C25
CUMULATIVE DISTRIBUTION OF PB WORD ARTICULATION: 40 INTERFERERS, 200/200 SPLATTER, 4 KM MEAN COMM. DIST., 20 HOPS/S, 50 TRIALS, MOBILE-TO-BASE

the receiver and intended signal source 100 km from the receiver) is on the order of 115 dB, there is no way that a single interferer (or any reasonably likely combination of multiple interferers) can significantly contribute to pushing the net signal-to-interference ratio across the 6 dB threshold unless it is already extremely close to that threshold with co-channel interference alone.

The results of this change are significant, indicating that the 10 dB/30 dB attenuation figures used in the initial simulation runs had contributed to a measureable performance degradation. With 200 dB/200 dB splatter, the lower decile of the distribution of articulation scores is about 0.84, compared with 0.73 for the equivalent run with 10/30 splatter. (Both runs were for 40 interferers.)

The impact of increasing the number of interferers from 40 to 80 is shown in Figure C-26. All other parameters are unchanged from the previous run. The performance at this level still appears to be acceptable; the lower decile of the distribution is 0.77. When the number of interferers is increased to 120 (Figure C-27), the lower decile decreases to about 0.63; there is only are 82 percent chance of achieving an articulation score of 0.75. Thus, the maximum number of users for this case (given the performance constraint stated above) is between 80 and 120, or about one per channel.

Although this result is encouraging, it is also unrealistic. The potential cost of developing and producing (1) receivers with 200 dB selectivity and (2) transmitters with 200 dB suppression of sideband noise and other spurious emissions, is likely to dissuade any entrepreneur from undertaking such an effort, particularly for a consumer-oriented application.

In order to see whether similar results could be attained with more realistic "splatter" parameters, the adjacent channel and next-adjacent channel attenuations were changed to 60 dB and 80 dB, respectively. Simultaneously, the mean distance over which communication is attempted was restored to 8.0 km (5 miles) and the number of interferers was set to 80. The results are shown in Figure C-28. It can be seen that the combination of changes has put the performance back into the unacceptable region, since the probability of achieving a 75 percent PB word articulation score is only about 0.76.

The effect of maintaining the 60/80 splatter configuration while reducing the mean attempted communication distance 4 km (2.5 miles) is shown in Figure C-29. This run is comparable to the run displayed in Figure C-26 except for the increased splatter.

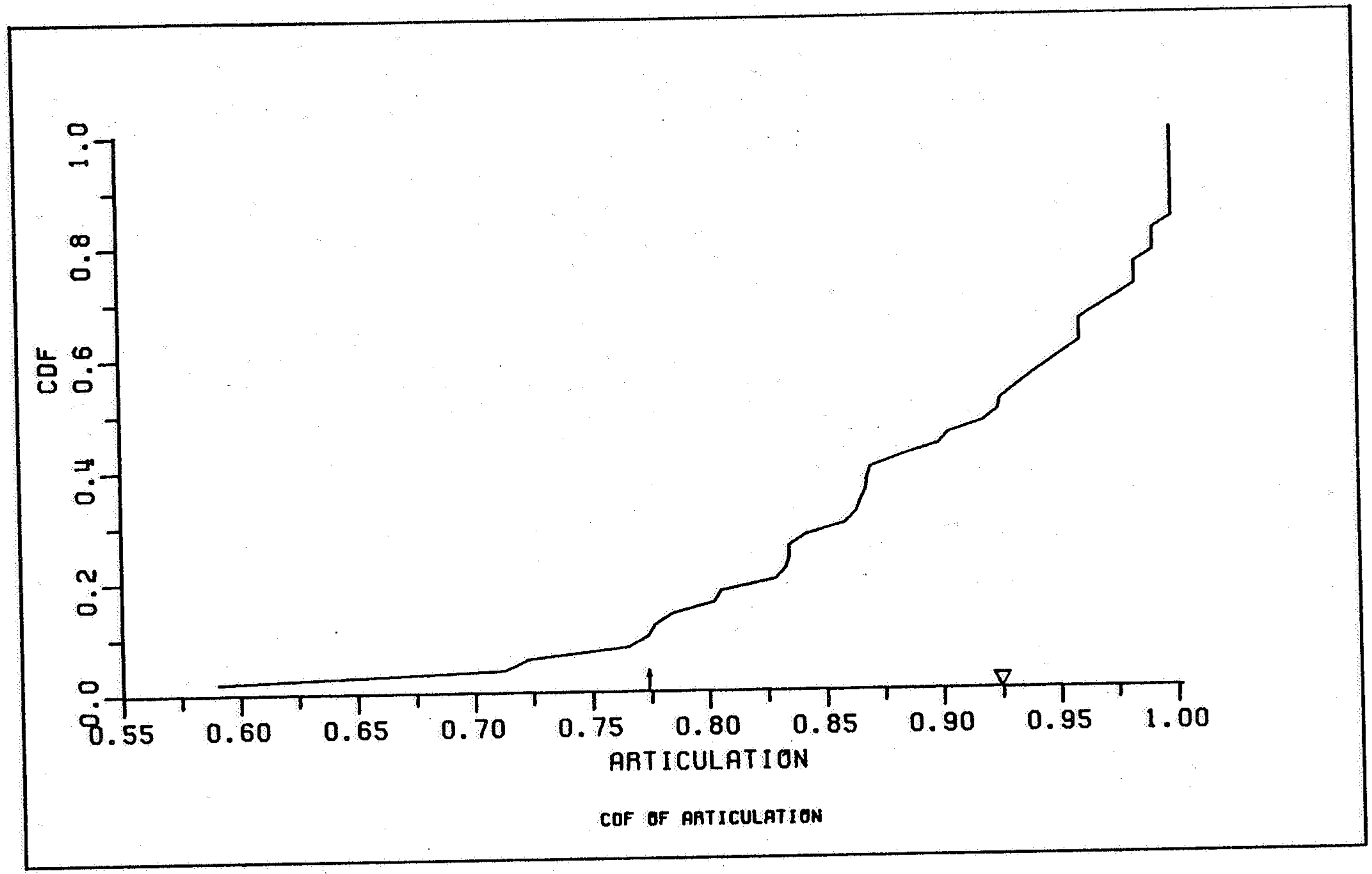

FIGURE C26
CUMULATIVE DISTRIBUTION OF PB WORD ARTICULATION: 80 INTERFERERS, 200/200 SPLATTER, 4 KM MEAN COMM. DIST., 20 HOPS/S, 50 TRIALS, MOBILE-TO-BASE

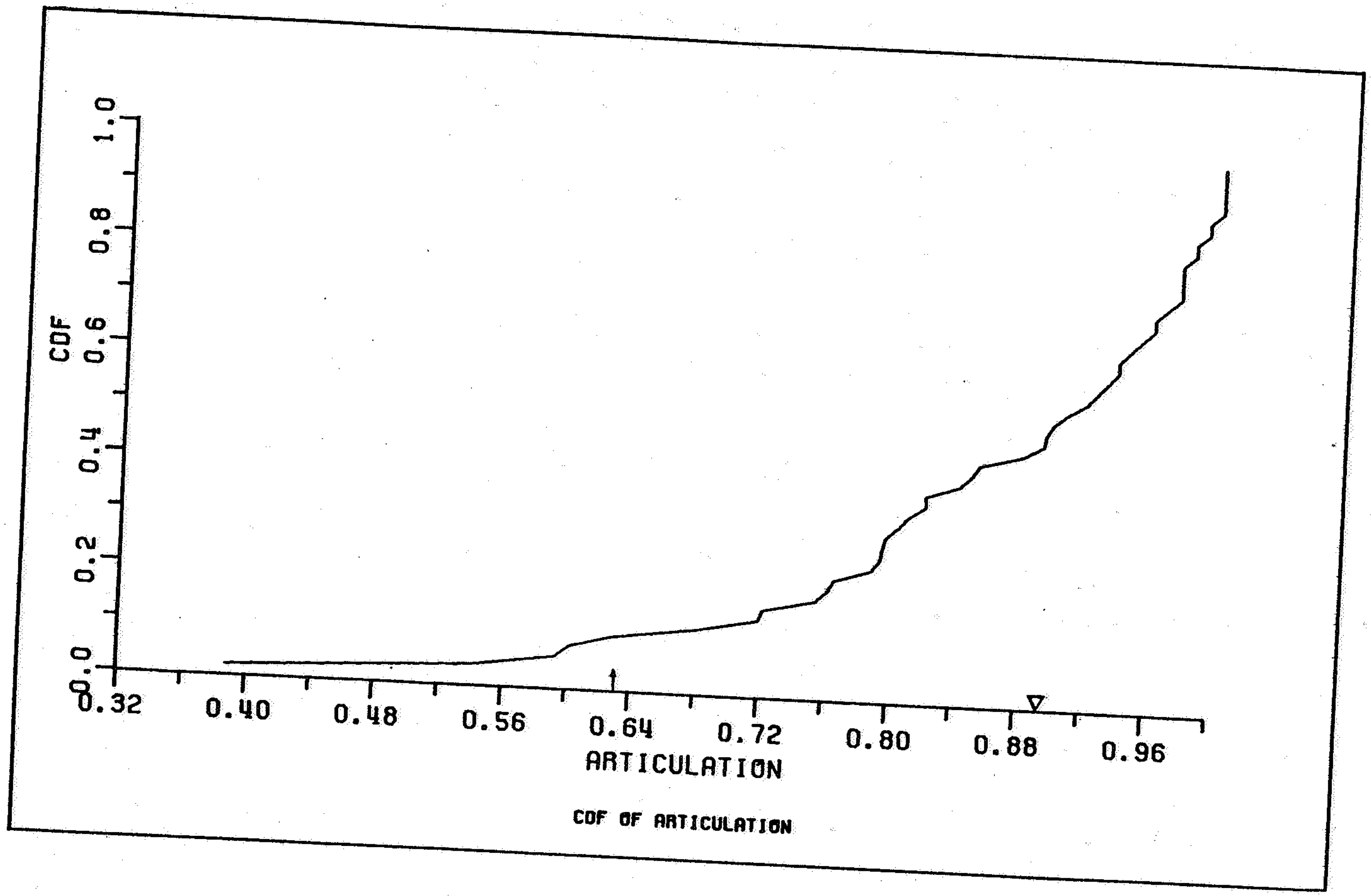

FIGURE C27
CUMULATIVE DISTRIBUTION OF PB WORD ARTICULATION: 120 INTERFERERS, 200/200 SPLATTER, 4 KM MEAN COMM. DIST., 20 HOPS/S, 50 TRIALS MOBILE-TO-BASE

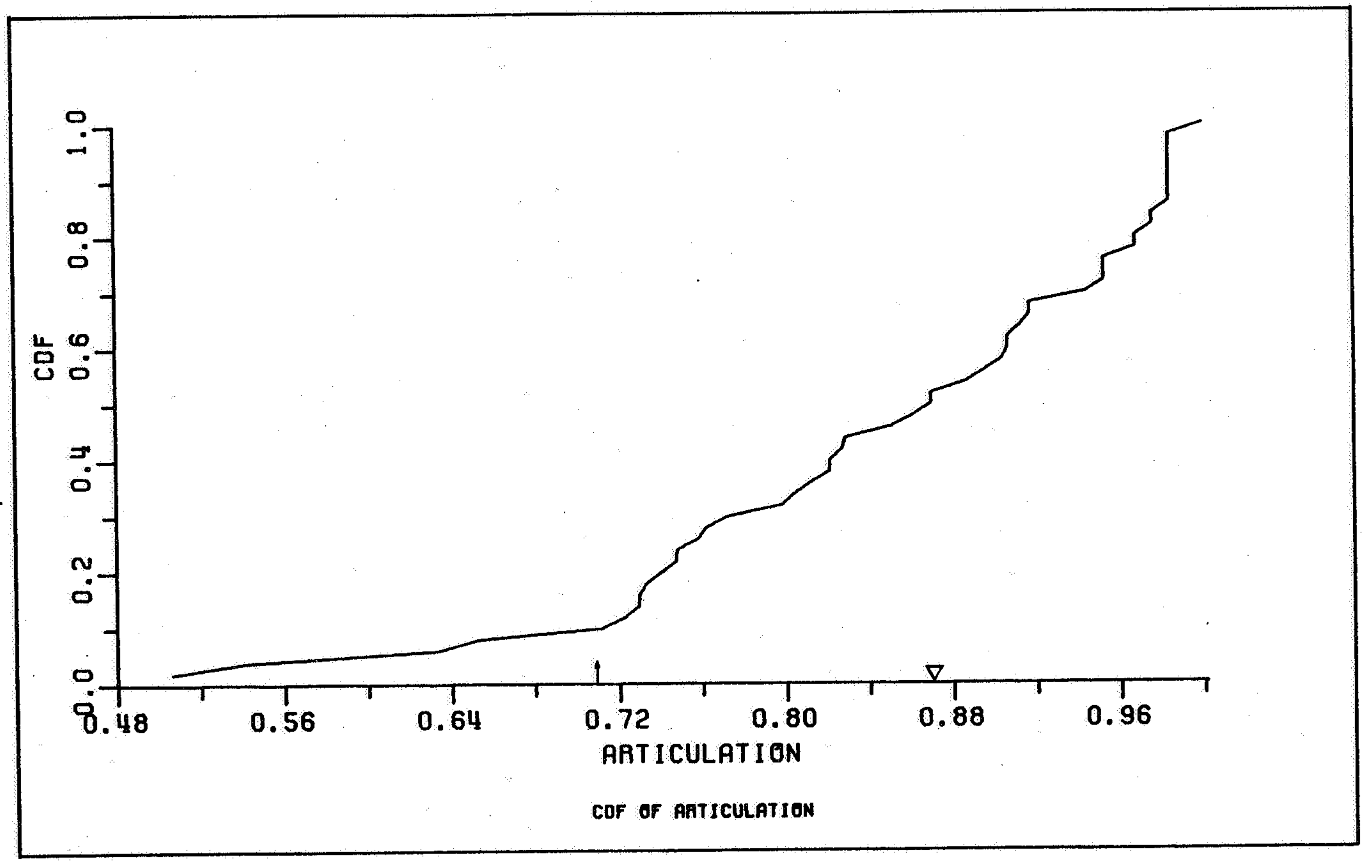

FIGURE C28
CUMULATIVE DISTRIBUTION OF PB WORD ARTICULATION: 80 INTERFERERS, 60/80 SPLATTER, 8 KM MEAN COMM. DIST., 20 HOPS/S, 50 TRIALS, MOBILE-TO-BASE

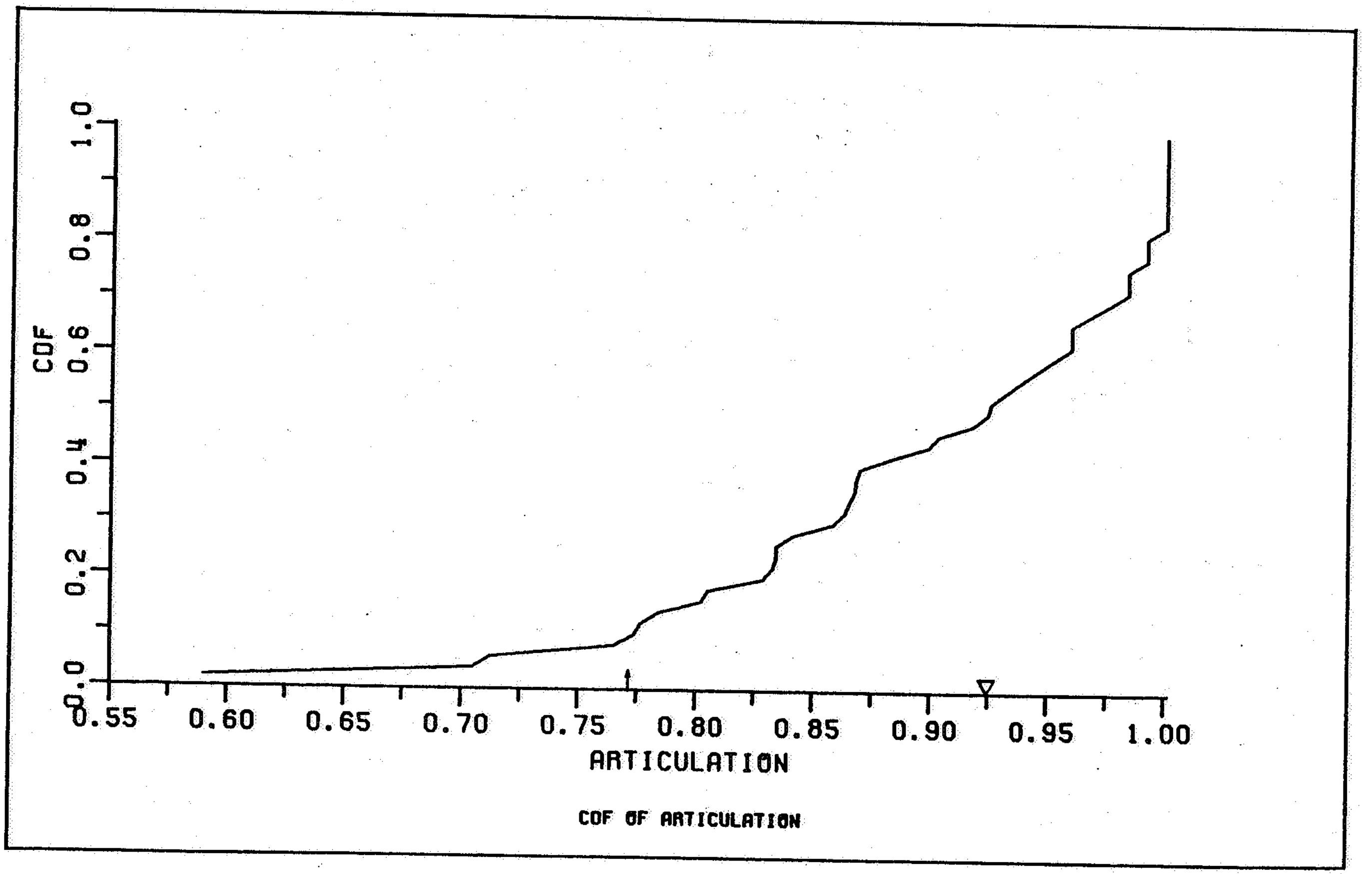

FIGURE C29
CUMULATIVE DISTRIBUTION OF PB WORD ARTICULATION: 80 INTERFERERS, 60/80 SPLATTER, 4 KM MEAN COMM. DIST., 20 HOPS/S, 50 TRIALS, MOBILE-TO-BASE

It can be seen that the performance is now back in the acceptable region; the chance of attaining a 75 percent word articulation score is about 0.93. Thus, the maximum number of users per channel for this case is again close to unity.

Receiver selectivities of 60 dB (adjacent channel) and 80 dB (next-adjacent channel) are attainable using multipole crystal filters in conjunction with 25 KHz channel spacing. However, reducing transmitter sideband splatter to these levels may require careful design efforts or unusually wide channel spacings. The impact of receiver-generated intermodulation products also needs to be investigated.

Additional simulation runs with 60/80 dB splatter are shown in Figures C-30 through C-36 for mean communication distress of 2.0 km through 16.0 km. These results were used to provide a coarse plot of maximum number of simultaneous users versus mean communication distance (see Section C.4.4).

C.4.2.2 Mobile-to-Mobile Operation (Without Repeaters)

In order to simulate direct mobile-to-mobile operation, it is only necessary to generate a new path loss table based on a lower antenna height at the receiver. This assumes that all interferers are mobiles. Antenna heights of 2.0 meters were assumed at both ends of the link. By the Longley-Rice model, this had the net effect of increasing the median path loss by about 5 dB for distance greater than 1.0 km. For shorter distances, free space path loss is used, so the received power is independent of antenna height in this range.

The results are shown in Figures C-37 and C-38 for mean attempted communication distances of 4 km (2.5 miles) and 2 km (1.25 miles), respectively. Both runs are based on 80 interferers and 60/80 dB splatter. At 4 km, the performance is not significantly different from the corresponding mobile-to-base performance (Figure C-29). At 2 km, performance is substantially improved.

C.4.2.3 Base-to-Mobile Operation

The mobile-to-base and base-to-mobile links in land mobile radio systems are normally not equivalent, because the antenna gains and noise levels can differ. However, all of the simulation runs described above are based on the assumptions of interference-limited operation, so neither background noise nor antenna gains are reflected in the results. Also, the (Longley-Rice) propagation model is reciprocal with respect to transmitter and receiver antenna heights, so the uplink and downlink are equivalent in terms of signal-to-interference ratio for a given set of transmitter-to-receiver distances.

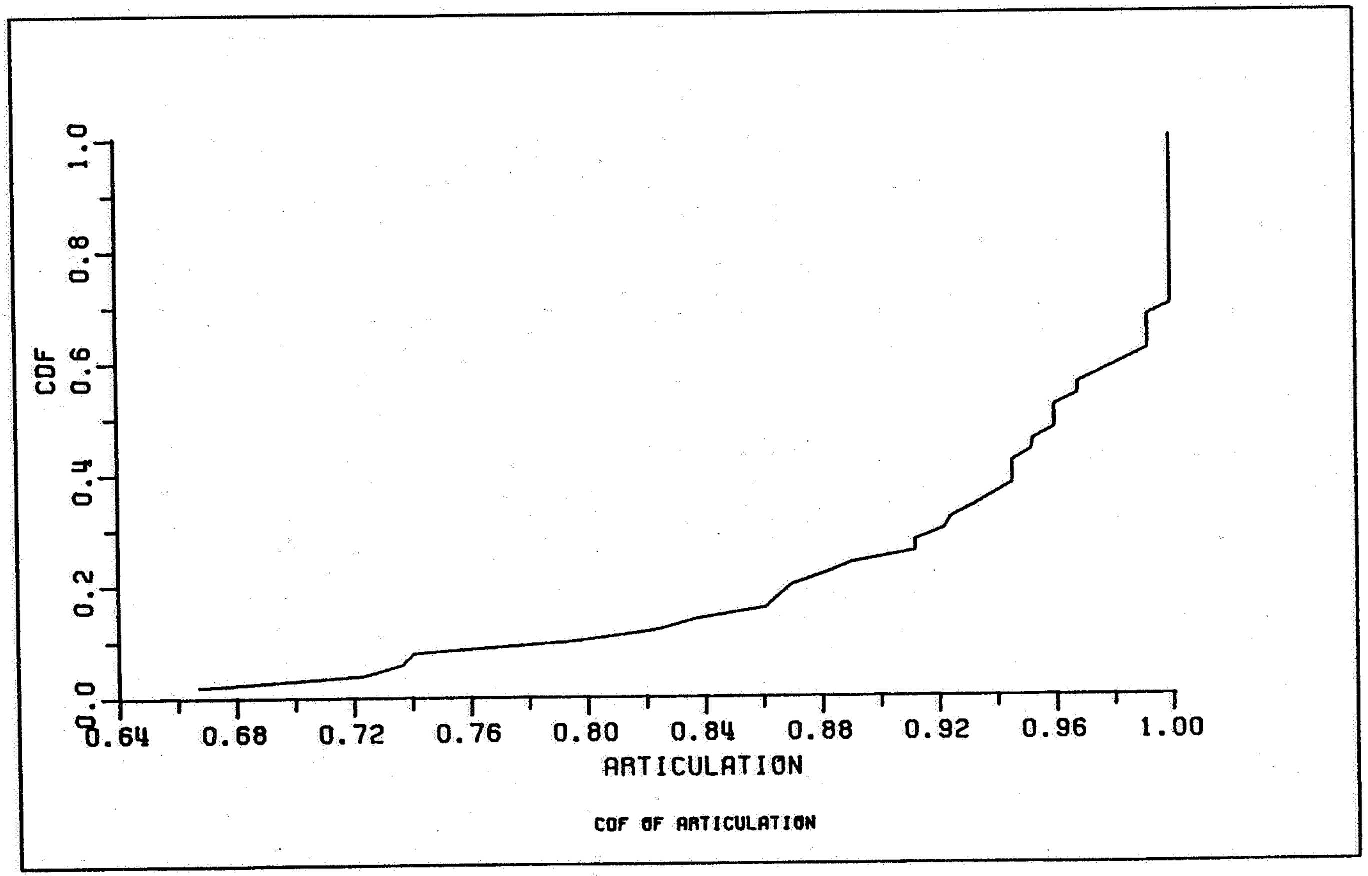

FIGURE C30
CUMULATIVE DISTRIBUTION OF PB WORD ARTICULATION: 100 INTERFERERS, 60/80 SPLATTER, 2 KM MEAN COMM. DIST., 20 HOPS/S, 50 TRIALS, MOBILE-TO-BASE

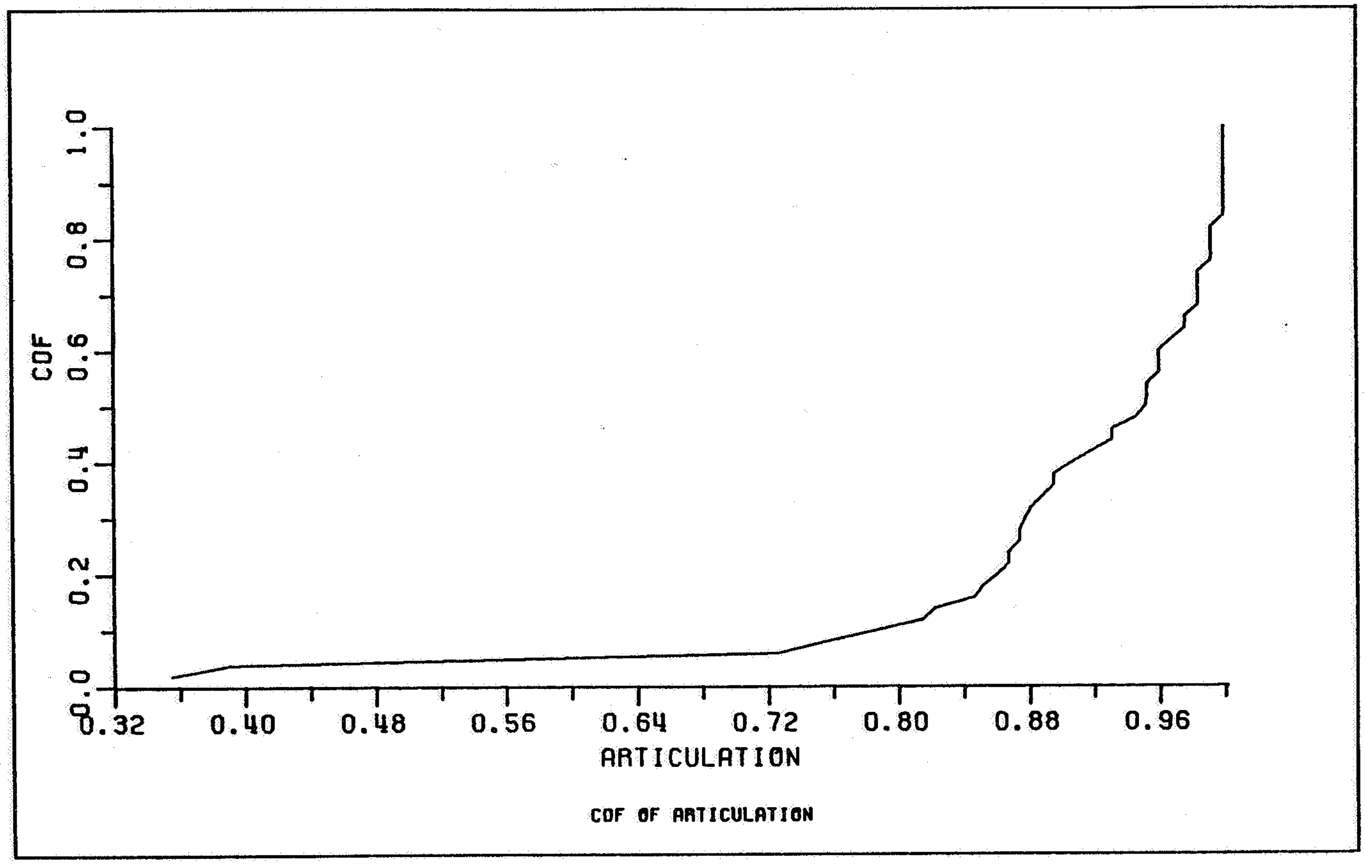

FIGURE C31
CUMULATIVE DISTRIBUTION OF PB WORD ARTICULATION: 130 INTERFERERS, 60/80 SPLATTER, 2 KM MEAN COMM. DIST., 20 HOPS/S, 50 TRIALS, MOBILE-TO-BASE

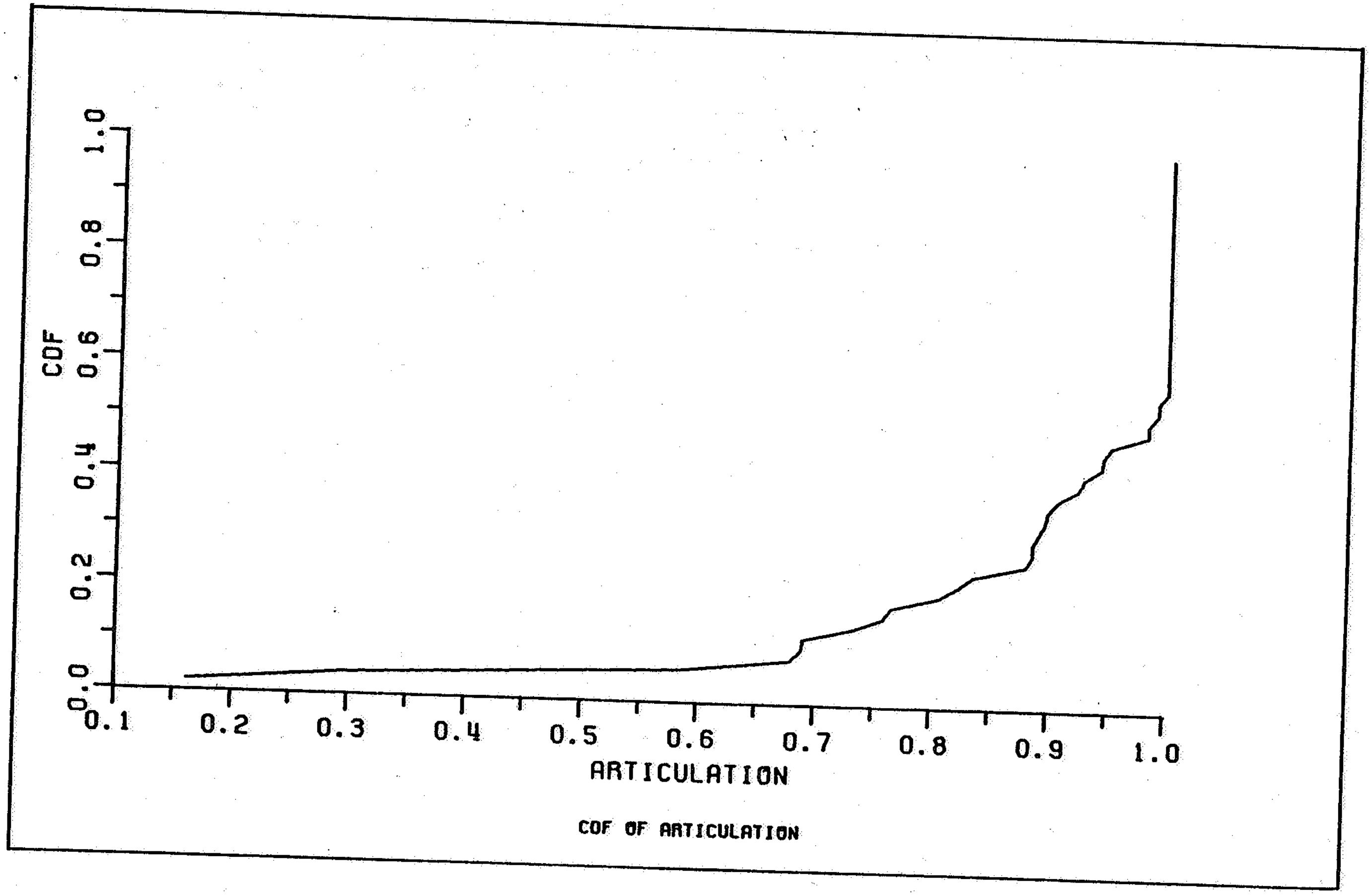

FIGURE C32
CUMULATIVE DISTRIBUTION OF PB WORD ARTICULATION: 160 INTERFERERS, 60/80 SPLATTER, 2 KM MEAN COMM. DIST., 20 HOPS/S, 50 TRIALS, MOBILE-TO-BASE

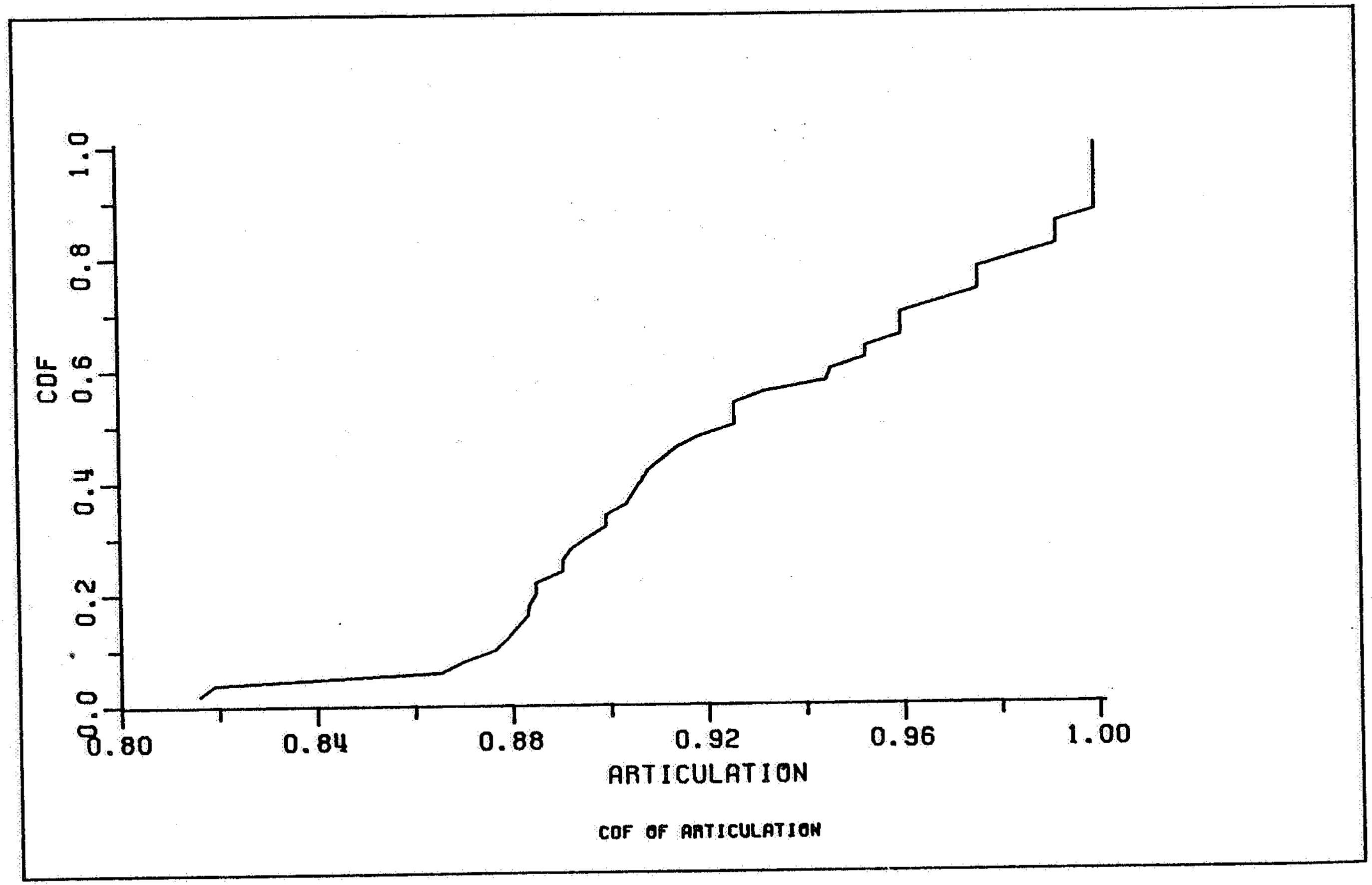

FIGURE C33
CUMULATIVE DISTRIBUTION OF PB WORD ARTICULATION: 20 INTERFERERS, 60/80 SPLATTER, 16 KM MEAN COMM. DIST., 20 HOPS/S, 50 TRIALS, MOBILE-TO-BASE

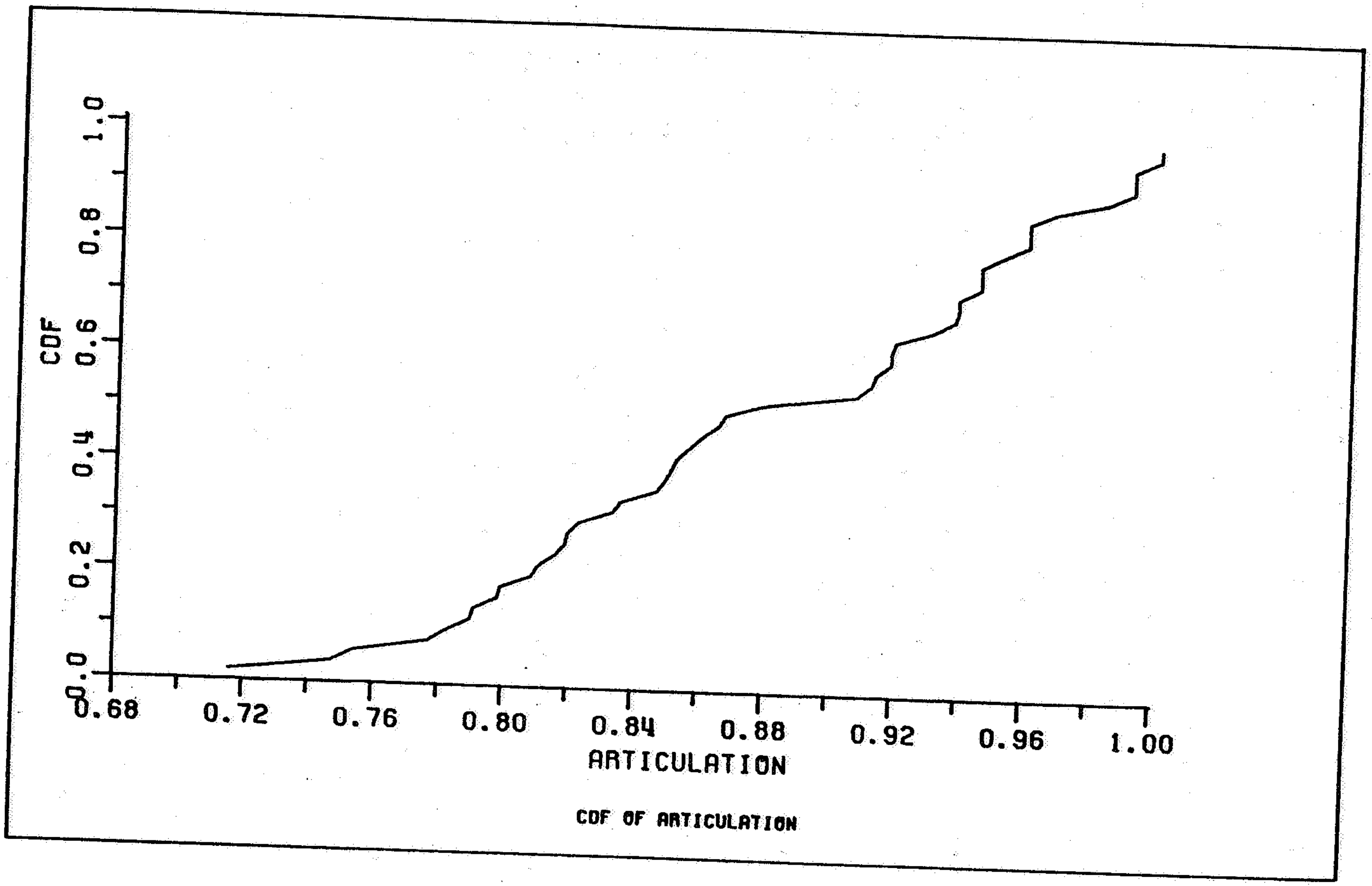

FIGURE C34
CUMULATIVE DISTRIBUTION OF PB WORD ARTICULATION: 40 INTERFERERS, 60/80 SPLATTER, 16 KM MEAN COMM. DIST., 20 HOPS/S, 50 TRIALS, MOBILE-TO-BASE

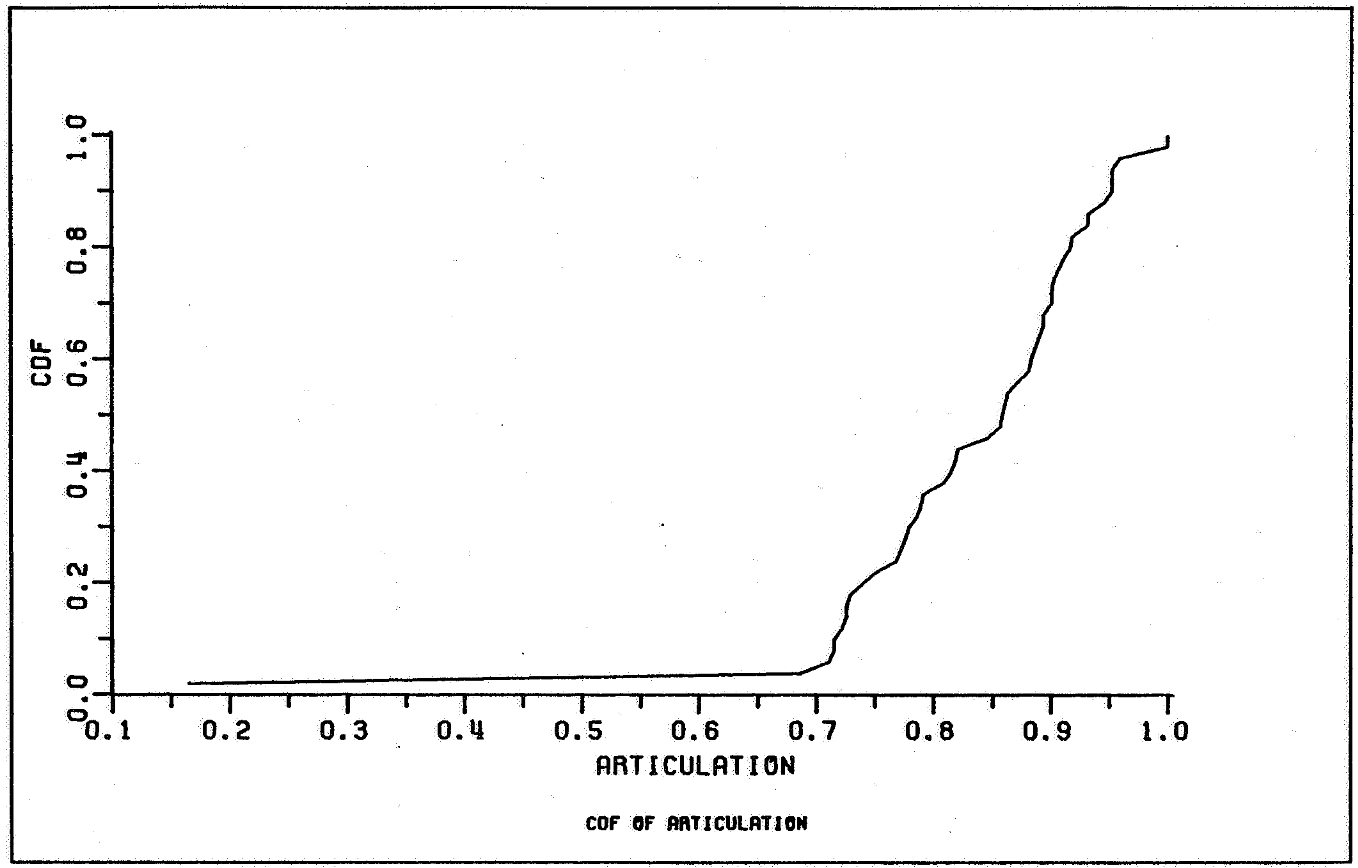

FIGURE C35
CUMULATIVE DISTRIBUTION OF PB WORD ARTICULATION: 60 INTERFERERS, 60/80 SPLATTER, 16 KM MEAN COMM. DIST., 20 HOPS/S, 50 TRIALS, MOBILE-TO-BASE

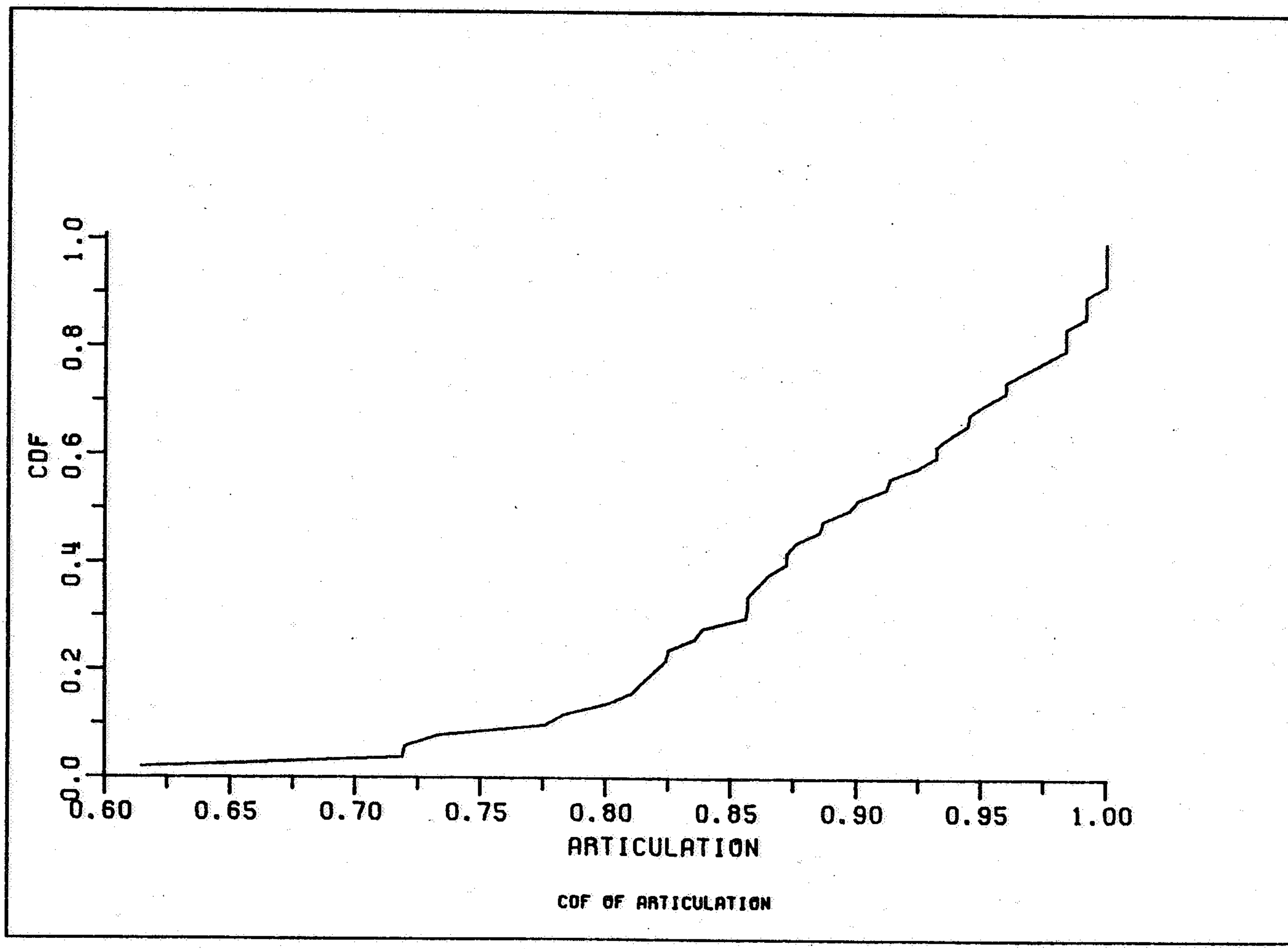

FIGURE C36
CUMULATIVE DISTRIBUTION OF PB WORD ARTICULATION: 60 INTERFERERS, 60/80 SPLATTER, 8 KM MEAN COMM. DIST., 20 HOPS/S, 50 TRIALS, MOBILE-TO-BASE

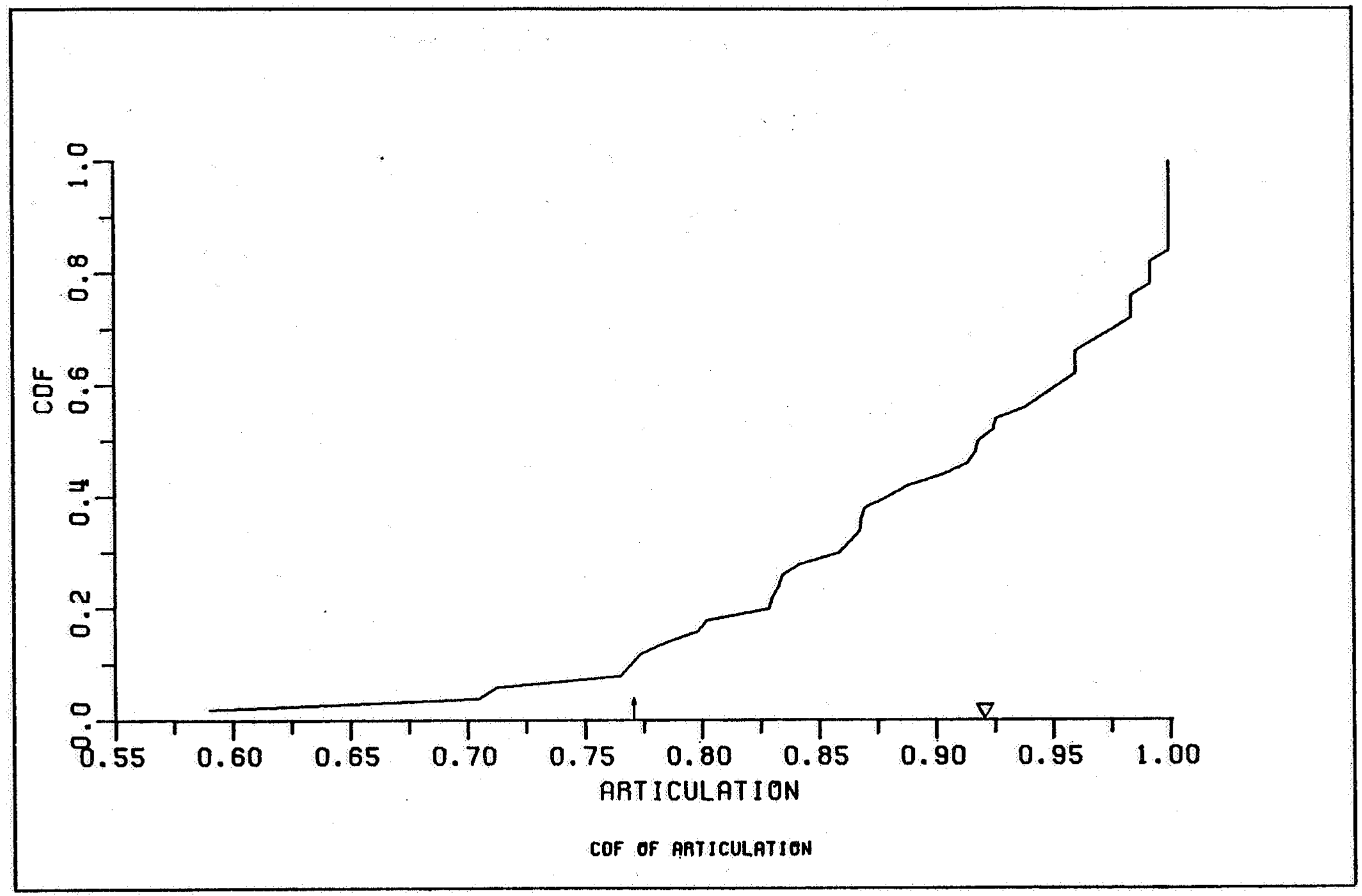

FIGURE C37
CUMULATIVE DISTRIBUTION OF PB WORD ARTICULATION: 80 INTERFERERS, 60/80 SPLATTER, 4 KM MEAN COMM. DIST., 20 HOPS/S, 50 TRIALS, MOBILE-TO-BASE

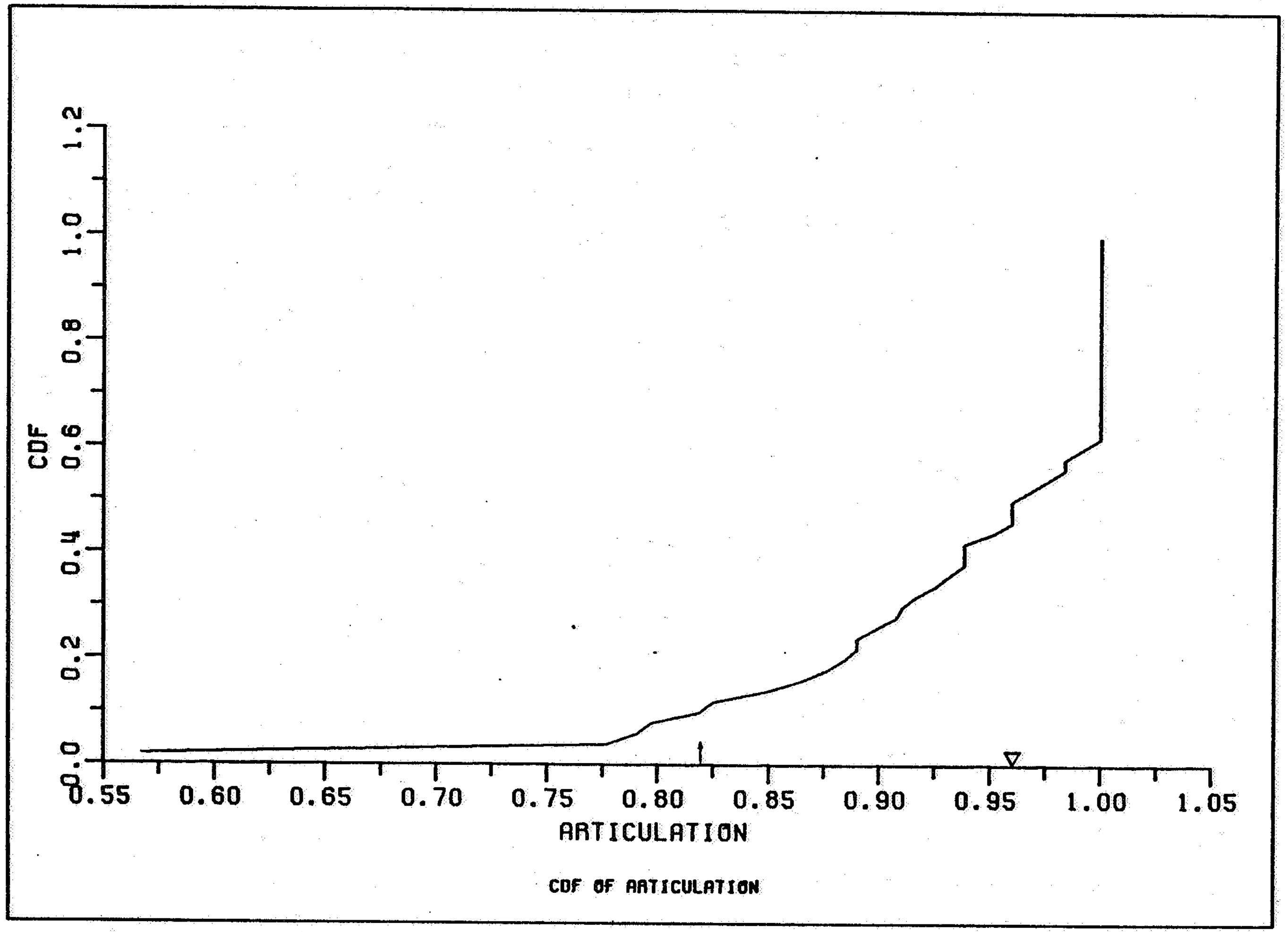

FIGURE C38
CUMULATIVE DISTRIBUTION OF PB WORD ARTICULATION: 80 INTERFERERS, 60/80 SPLATTER, 2 KM, MEAN COMM. DIST., 20 HOPS/S, 50 TRIALS, MOBILE-TO-MOBILE

Thus, the results presented above for mobile-to-base operation are also applicable, in principle, to the base-to-mobile link. In order for this equivalence to hold, however, it is necessary to imagine the mobile receiver at the center of the coverage area, with the base station transmitters distributed around it according to the density functions describe for mobile transmitters in Section C.3.7.

C.4.3 The Impact of Background Noise and Fading

The assumption of interference-limited operation is a useful mechanism that has been applied in other analyses and simulations of FH systems. Nevertheless, real systems are limited in the attached communication range in the absence of interference. Table C-3 illustrates a simplified 915 MHz mobile-to-base link budget for a quiet receiver location. Based on the assumed parameters, the maximum permissible path loss is 155 dB, including a 20 dB fade margin. In order for this loss to be achieved at 90 percent of the mobile locations, a maximum range of about 8.3 km (5.2 miles) is predicted by the Longley-Rice model, based on the parameters shown in Table C-1. This 155 dB path loss will be achieved 50 percent of the time at a range of about 19.3 km (12.1 miles). Noisey receiver locations, lower transmitter powers, line losses, and urban shadowing will all contribute to reduced ranges. Such range limitations obviously apply to conventional narrowband systems as well as frequency-hopping systems.

C.4.4 Summary of Simulation Results

The simulations described here indicate that it may be feasible to provide acceptable communication performance with slow frequency hopping FM. The spectrum efficiency of such an approach, in terms of number of active users per channel over a given geographic area depends on the degree of adjacent and next-adjacent channel interferences as well as the mean communication distance. The slow FH approach may be able to accommodate roughly one active user per channel under favorable conditions. The degree of adjacent-channel and next-adjacent channel "splatter" can have a significant impact on communication performance and spectrum efficiency. The estimated maximum number of users per channel is summarized in Table C-4 for various operating conditions. For 60/80 dB splatter, the estimated maximum number of users per channel is plotted versus the mean communication distance in Figure C-39.

TABLE C-3

SIMPLIFIED MOBILE-TO-BASE LINK BUDGET FOR

QUIET RECEIVER LOCATION

Receiver Sensitivity:		-146 dBW	(.35 NV)
Receiver Antenna Gain:		9 dB	
Transmit Antenna Gain:		5 dB	
Transmit Power:	15 dBW		
Maximum Nonfading Path Loss:		175 dB	
Fade Margin	20 dB		
Maximum Path Loss:	155 dB		
	8.3 km, 90%		
	19.3 km, 50%		

TABLE 4

ESTIMATED MAXIMUM NUMBER OF USERS PER CHANNEL

FOR INTERFERENCE-LIMITED OPERATION

Splatter, dB Attenuation	Mean Communication Distance, km	Type of Link	Estimated Maximum Number of Users Per Channel
10/30	8.0	MB	0.3
10/30	4.0	MB	0.3
200/200	4.0	MB	1.0
60/80	16.0	MB	0.5
60/80	8.0	MB	0.6
60/80	4.0	MB	0.9
60/80	2.0	MB	1.4
60/80	4.0	MM	0.9
60/80	2.0	MM	1.4

C.5 REFERENCES

1. A. G. Longley and P. L. Rice, "Prediction of Tropospheric Radio Transmission Loss Over Irregular Terrain, A Computer Method-1968," ESA Technical Report ERL-79-ITS67, AD-676874, July 1968.

2. Anita G. Longley, "Location Variability of Transmission Loss -- Land Mobile and Broadcast Systems," OT Report 76-87, May 1976.

3. Anita G. Longley, "Radio Propagation in Urban Areas," 2nd IEEE Vehicular Technology Conference, Denver, March 22-24, 1978.

4. George A. Miller and J.C.R. Licklider, "The Intelligibility of Interrupted Speech," J. Acoustical Soc. of America, Vol. 22, No. 2 March 1950.

5. J. P. Egan, "Articulation Testing Methods," Laryngoscope 58, 1948.

6. Carl E. Williams, et al., "Intelligibility Test Methods and Procedures for the Evaluation of Speech Communication Systems," Report No. ESD-TR-66-677, by Bolt, Beranek, and Newman for Decision Sciences Laboratories, Electronic Systems Division, U.S. Air Force, December 1966.

7. Don J. Torrieri, "Simultaneous Mutual Interference and Jamming in a Frequency Hopping Network," DARCOM Countermeasures/Counter-Countermeasures Office, DARCOM CM/CCM 79-6, October 1979.

8. Publicaton R29-3-15, Motorola, Inc., Communications Systems Division, Component Products, Franklin Park, Illinois.

Department Approval: R. Braff

R. Braff

Author Approval: Walter C Scales

W. Scales