AD-A050 844 IIT RESEARCH INST ANNAPOLIS MD F/G 17/2

PERFORMANCE OF VOICE COMMUNICATIONS SYSTEMS IN THE PRESENCE OF --ETC(U)

DEC 77 L FARBER, J CORMACK F19628-78-C-0006

UNCLASSIFIED ESD-TR-77-005 NL

1 OF 1

AD A050 844

END

DATE FILMED

4-78

DDC

ESD-TR-77-005

(12)

PERFORMANCE OF VOICE COMMUNICATIONS SYSTEMS IN THE PRESENCE OF SPREAD SPECTRUM INTERFERENCE

IIT Research Institute
Under Contract to
DEPARTMENT OF DEFENSE
Electromagnetic Compatibility Analysis Center
Annapolis, Maryland 21402

December 1977

FINAL REPORT

Approved for public release; distribution unlimited.

Prepared for

U.S. Naval Electronic Systems Command
Washington, DC 20360

ESD-TR-77-005

This report was prepared by the IIT Research Institute under Contract F-19628-78-C-0006 with the Electronic Systems Division of the Air Force Systems Command for operation of the DoD Electromagnetic Compatibility Analysis Center, Annapolis, Maryland.

This report has been reviewed and is approved for publication.

Leonard Farber

LEONARD FARBER
Project Engineer, IITRI

David S. Anderson

DAVID S. ANDERSON
Assistant Director
Contractor Operations

Approved by:

Thomas A. Anderson

THOMAS A. ANDERSON
Colonel, USAF
Director

H. E. Segelhorst

H. E. SEGELHORST
Commander, USN
Navy Deputy Director

UNCLASSIFIED

SECURITY CLASSIFICATION OF THIS PAGE (When Data Entered)

(19) **REPORT DOCUMENTATION PAGE**

READ INSTRUCTIONS BEFORE COMPLETING FORM

(18) 1. REPORT NUMBER: ESD-TR-77-005

2. GOVT ACCESSION NO.

3. RECIPIENT'S CATALOG NUMBER

(6) 4. TITLE (and Subtitle): PERFORMANCE OF VOICE COMMUNICATIONS SYSTEMS IN THE PRESENCE OF SPREAD SPECTRUM INTERFERENCE.

(9) 5. TYPE OF REPORT & PERIOD COVERED: FINAL rept.

6. PERFORMING ORG. REPORT NUMBER

(10) 7. AUTHOR(s): Leonard /Farber, J. /Cormack

(15) 8. CONTRACT OR GRANT NUMBER(s): F19628-78-C-0006

9. PERFORMING ORGANIZATION NAME AND ADDRESS:
DoD Electromagnetic Compatibility Analysis Center
North Severn
Annapolis, MD 21402

10. PROGRAM ELEMENT, PROJECT, TASK AREA & WORK UNIT NUMBERS: (16) 649E

11. CONTROLLING OFFICE NAME AND ADDRESS:
U.S. Naval Electronic Systems Command
Washington, DC 20360

(11) 12. REPORT DATE: Dec[redacted]77 (12)

13. NUMBER OF PAGES: 78 73p.

14. MONITORING AGENCY NAME & ADDRESS (if different from Controlling Office)

15. SECURITY CLASS. (of this report): UNCLASSIFIED

15a. DECLASSIFICATION/DOWNGRADING SCHEDULE

16. DISTRIBUTION STATEMENT (of this Report)

Approved for public release; distribution unlimited.

17. DISTRIBUTION STATEMENT (of the abstract entered in Block 20, if different from Report)

UNLIMITED

18. SUPPLEMENTARY NOTES

19. KEY WORDS (Continue on reverse side if necessary and identify by block number)

SPREAD-SPECTRUM
DIRECT-SEQUENCE
FREQUENCY-HOPPING
VOICE COMMUNICATIONS
COMPUTER SIMULATION

20. ABSTRACT (Continue on reverse side if necessary and identify by block number)

A computer simulation model developed for determining the manner in which undesired spread spectrum signals affect the performance of conventional narrowband AM and FM voice communication systems is described. Direct-sequence, frequency-hopping and hybrid, frequency-hopping/direct-sequence, spread spectrum interference signals were modeled. Binary phase-shift keying and minimum-shift keying were included for the carrier modulation of the direct-sequence signals and the direct-sequence portion of the

175 300

20. Abstract (Continued).

hybrid signals. In addition to the spread spectrum interference, pulsed interference and white Gaussian noise were modeled. Using articulation index as the measure of performance, comparisons are made between system performance for direct-sequence interference and white Gaussian noise. System performance for frequency-hopping or frequency-hopping/direct-sequence interference is compared with that for pulsed interference.

PREFACE

The Electromagnetic Compatibility Analysis Center (ECAC) is a Department of Defense facility, established to provide advice and assistance on electromagnetic compatibility matters to the Secretary of Defense, the Joint Chiefs of Staff, the military departments and other DoD components. The center, located at North Severn, Annapolis, Maryland 21402, is under executive control of the Assistant Secretary of Defense for Communication, Command, Control, and Intelligence and the Chairman, Joint Chiefs of Staff, or their designees, who jointly provide policy guidance, assign projects, and establish priorities. ECAC functions under the direction of the Secretary of the Air Force and the management and technical direction of the Center are provided by military and civil service personnel. The technical operations function is provided through an Air Force sponsored contract with the IIT Research Institute (IITRI).

This report was prepared as part of AF Project 649E under Contract F-19628-78-C-0006 by the staff of the IIT Research Institute at the Department of Defense Electromagnetic Compatibility Analysis Center.

To the extent possible, all abbreviations and symbols used in this report are taken from American Standard Y10.19 (1967) "Units Used in Electrical Science and Electrical Engineering" issued by the USA Standards Institute.

Users of this report are invited to submit comments that would be useful in revising or adding to this material to the Director, ECAC, North Severn, Annapolis, Maryland 21402, Attention ACY.

EXECUTIVE SUMMARY

The Naval Electronic Systems Command (NAVELEX) tasked the Electromagnetic Compatibility Analysis Center (ECAC) to determine criteria for acceptable performance of conventional voice communication systems operating in the presence of a spread spectrum system and to submit a document containing the results to the International Radio Consultative Committee (CCIR) for publication. A document was prepared and submitted in March 1977 to CCIR United States Study Group (USSG) 1A for publication. The results in the submitted document were obtained using an analysis based on assumptions as to the manner in which undesired spread spectrum signals would affect the performance of conventional voice communications systems. Subsequently, a model was developed for determining the manner in which three types of undesired spread spectrum signals affect the performance of conventional narrowband AM and FM voice communication systems, and the model is described in this report.

The model developed utilizes computer simulation. Direct-sequence, frequency-hopping and hybrid, frequency-hopping/direct sequence, spread spectrum interference signals were modeled. Binary phase-shift keying and minimum-shift keying were included for the carrier modulation of the direct-sequence signals and the direct-sequence portion of the hybrid signals. In addition to the spread spectrum interference, pulsed interference and white Gaussian noise were modeled. Articulation index is used as the measure of system performance.

For an AM and FM voice communication system, comparisons are made between system performance for direct-sequence interference and white Gaussian noise. System performance for frequency-hopping or frequency-hopping/direct-sequence interference is compared with that for pulsed interference.

EXECUTIVE SUMMARY (Continued)

The results indicate the following, for conventional narrowband AM and FM voice communications systems at typical operating performance levels and a given amount of undesired power produced within the IF filter:

1. A direct-sequence interfering signal would result in approximately the same system performance as white Gaussian noise of the same bandwidth.

2. A frequency-hopping or short-dwell-time, frequency-hopping/direct-sequence interfering signal having only one of its frequencies hop within the IF filter would result in approximately the same system performance as a periodic pulsed signal in which the dwell time of the hopper is equivalent to the pulse width of the periodic signal and the hopper's mean frequency occurrence rate is equivalent to the PRF of the pulsed signal.

TABLE OF CONTENTS

TABLE OF CONTENTS (Continued)

LIST OF ILLUSTRATIONS

TABLE OF CONTENTS (Continued)

LIST OF ILLUSTRATIONS (Continued)

LIST OF TABLES

LIST OF APPENDIXES

SECTION 1

INTRODUCTION

BACKGROUND

A class of communication systems termed spread spectrum has been evolving in recent years. The ability of a communication system to resist interference is directly related to the emitted signal bandwidth for a given information bandwidth. One characteristic of a spread spectrum system is that the emitted signal bandwidth is much greater than the bandwidth of the information being transmitted. Thus, one of the primary reasons for the current interest in spread spectrum systems is that they should be less susceptible to interference than conventional systems. Communication systems that employ a variety of spread spectrum modulation techniques are being increasingly developed and, in the near future, this should lead to an environment in which a relatively small number of spread spectrum systems are operating along with a much larger number of conventional systems.

The Naval Electronic Systems Command (NAVELEX) tasked the Electromagnetic Compatibility Analysis Center (ECAC) to determine criteria for acceptable performance of conventional voice communication systems operating in the presence of a spread spectrum system and to submit a document containing the results to the International Radio Consultative Committee (CCIR) for publication. A document[1] was prepared and submitted in March 1977 to CCIR United States Study Group (USSG) 1A for publication. The results in the submitted document were obtained using an analysis based on assumptions as to the manner in which undesired spread spectrum signals would affect the performance of conventional

[1]Farber, L., *Some Criteria For Frequency Sharing Between Voice Communication and Spread Spectrum Systems,* ECAC-CR-77-020, ECAC, Annapolis, MD, March 1977.

voice communication systems. Subsequently, a model was developed for determining the manner in which three types of undesired spread spectrum signals affect the performance of conventional narrowband AM and FM voice communication systems, and the model is described in this report.

OBJECTIVE

The objective of the effort reported herein was to develop a model for determining the manner in which direct-sequence, frequency-hopping and hybrid, frequency-hopping/direct-sequence spread spectrum interference signals affect the performance of conventional narrowband AM and FM voice communication systems.

APPROACH

A computer simulation model was developed for determining the performance of conventional narrowband AM and FM voice communication systems operating in the presence of spread spectrum interference signals. Direct-sequence, frequency-hopping and frequency-hopping/direct-sequence spread spectrum interference signals were modeled. Binary phase-shift keying and minimum-shift keying were included for the carrier modulation of the direct-sequence signals and the direct-sequence portion of the hybrid signals. In addition to the spread spectrum interference, pulsed interference and white Gaussian noise were modeled.

The receiver stages modeled include the intermediate frequency filter, detector and baseband filter. Composite signals comprised of a desired signal, an interfering signal and white Gaussian noise were processed through the receiver models in the time domain and the output signal was then transformed to the frequency domain where system performance was evaluated. Articulation index and the desired-to-undesired power ratio at the receiver output were the performance measures computed. Comparisons were made between system performance for direct-sequence interference and white Gaussian noise. System performance for frequency-hopping or frequency-hopping/direct-sequence interference was compared with that for pulsed interference.

SECTION 2

OVERALL SIMULATION MODEL

GENERAL

The Spread Spectrum Interference Model (SSIM) was developed to determine the manner in which spread spectrum (SS) interfering signals affect the performance of conventional narrowband AM and FM voice communication systems. The model is based on the Receiver Waveform Simulation (RWS) program[2] developed by ECAC. Direct-sequence (DS), frequency-hopping (FH) and frequency-hopping/direct sequence (FH/DS) SS signals were considered. Brief descriptions of the DS, FH and FH/DS techniques are given here with detailed descriptions of these and additional SS modulation techniques available in Dixon.[3]

The DS method of generating the wideband signal consists of appropriately combining the digital baseband data with a high-speed (relative to the data rate of the baseband signal) binary code sequence and then using the resulting sequence to modulate a carrier. The signaling speed of the code sequence is termed the chip rate while the term bit rate is reserved for the speed of the baseband data. Theoretically, any suitable form of amplitude or angle modulation may be used, with some form of phase-shift keying commonly employed. Binary phase-shift keying (PSK) and minimum-shift keying (MSK) were included in the SSIM.

The FH technique consists of generating the broad spectrum by hopping the carrier frequency among a number of frequencies over a wide band at some rate, with the specific order of frequency

---

[2]Meyers, R., *A Receiver Waveform Simulation Model,* ESD-TR-71-099, ECAC, Annapolis, MD, August 1971.

[3]Dixon, R. C., *Spread Spectrum Systems,* John Wiley and Sons, New York, 1976.

selection controlled by a code sequence. The system requirements determine the number of frequencies used and the hopping rate. The FH/DS technique consists of hopping the carrier frequency of a DS signal.

The SSIM is a simulation model and requires that the input signals (desired and interfering) and noise be sampled in the time domain. A composite signal is formed at the receiver input by combining the input signals, and the composite signal is then modified in accordance with the transfer characteristics of the various receiver stages. The model output is a sampled waveform representing the output of the receiver being modeled. Equivalent lowpass modeling techniques, often used in analytic solutions and representations of system characteristics, are used extensively throughout the model. The following subsection describes the functional structure of a receiver and examines the implications of modeling significant receiver functions (Reference 2).

RECEIVER FUNCTIONAL STRUCTURE

The basic receiver structure involved in the processing of desired and undesired signals is shown in Figure 1. The receiver signal-processing elements include the radio frequency (RF) amplifier, mixer, intermediate frequency (IF) amplifier, detector, baseband amplifier, and feedback circuitry. Receiver inputs, in general, consist of a desired signal, interfering signal(s) and noise. The receiver baseband output is a waveform which contains the desired and undesired signals.

The receiver processing of the input signals is simulated in the SSIM. Rather than a detailed circuit analysis of each receiver element, the model is a representation of the significant functions of the receiver signal processing illustrated in Figure 2. The implications of modeling significant receiver functions are examined next.

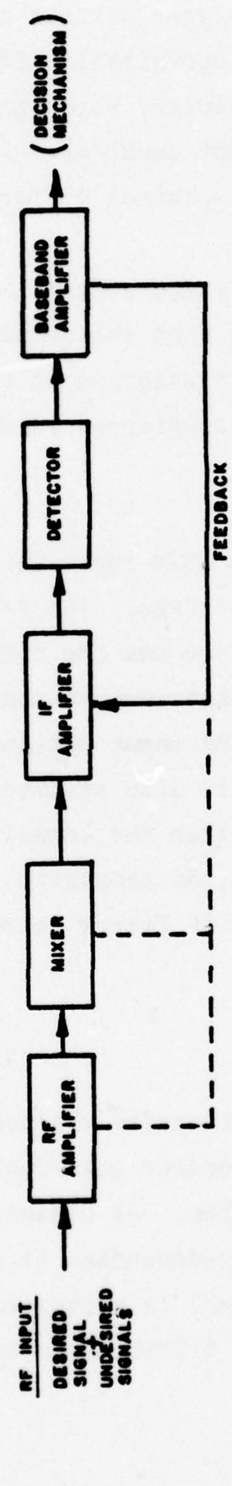


Figure 1. Basic receiver structure.

RF Amplifier and Mixer

RF amplifier and mixer stages possess nonlinear characteristics that are responsible for the undesirable effects of cross modulation, intermodulation, spurious response, saturation and desensitization. These nonlinear effects are not considered in what is generally termed cochannel and adjacent-channel performance degradation analysis.

RF amplifier and mixer stage effects are not considered in the SSIM. The assumption is made that the composite waveform at the IF input is simply a frequency translation of the receiver input waveform. The receiver stages actually considered in the model are shown in Figure 2.

The input signal to the SSIM represents the composite signal at the input to the IF amplifier stage. The relative amplitudes of the desired signal, the interference and the noise are specified program variables. Any processing losses due to the RF filter and mixer stages can be applied to the RF input SIR and SNR ratios to convert them to IF input ratios. It is also assumed that the bandwidth of the RF amplifier is much greater than the bandwidth of the IF amplifier. If this is not the case, then the composite selectivity of the RF and IF filters can be used as the IF filter selectivity characteristic in the model.

IF Amplifier

The IF amplifier linear transfer characteristics can be considered in terms of a frequency-independent gain function and a normalized frequency-dependent filter function. As discussed in the feedback circuitry subsection, only the frequency-dependent filter function is required in the model. The composite signal is processed through the IF filter in the time domain using digital filtering techniques.

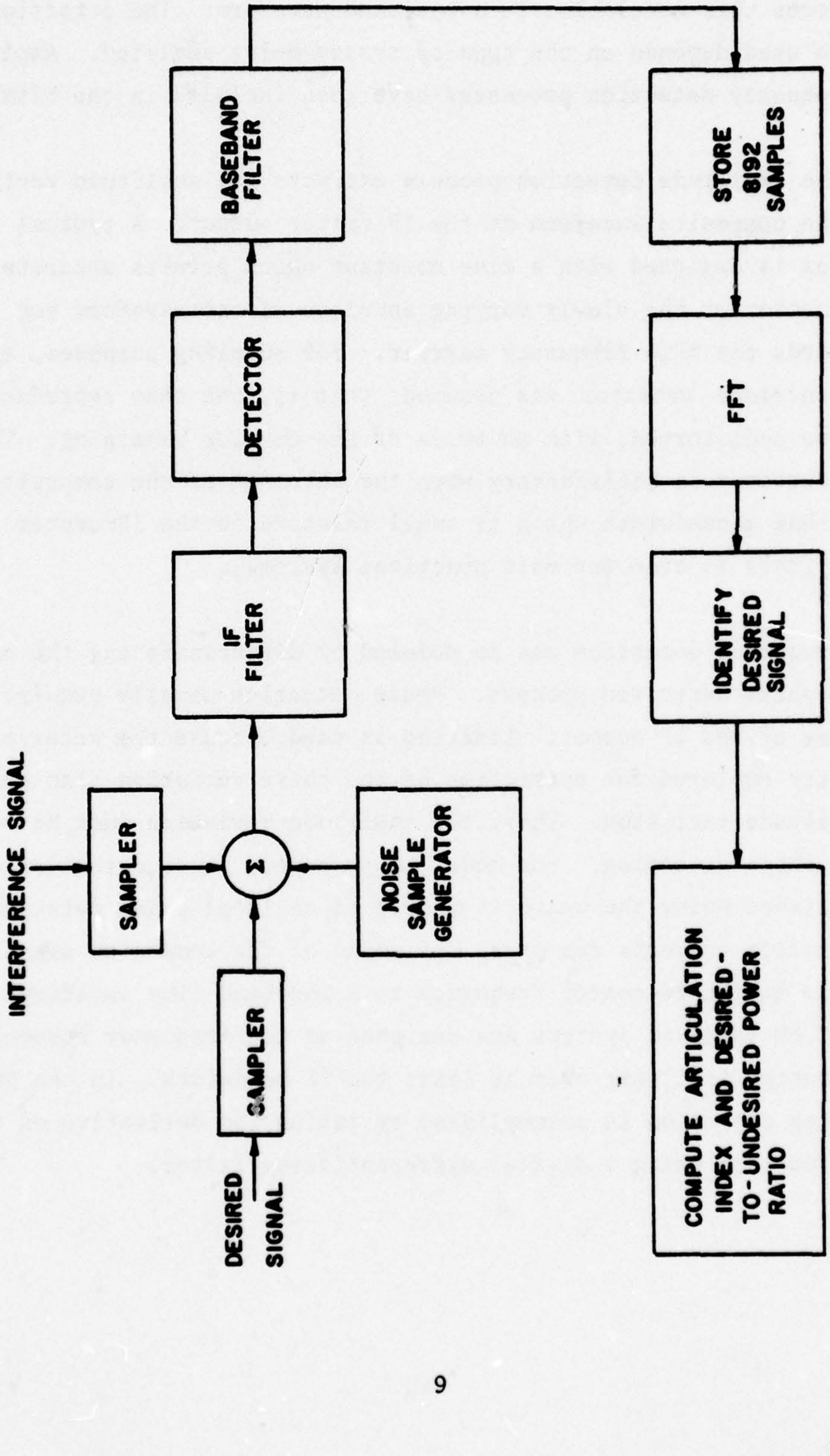


Figure 2. Overall simulation model.

Detector

The detector stage separates the modulation from the carrier and transforms this modulation to a baseband waveform. The detection process used depends on the type of system being analyzed. Amplitude and frequency detection processes have been included in the SSIM.

The amplitude detection process extracts the amplitude variations from the composite waveform at the IF filter output. A typical detector is designed with a time constant which permits accurate reproduction of the slowly varying envelope of the waveform but disregards the high frequency carrier. For modeling purposes, an ideal envelope detector was assumed; that is, one that reproduces the envelope undistorted, with no trace of the carrier remaining. This ideal detector is satisfactory when the envelope of the composite signal has a bandwidth which is small relative to the IF center frequency (this is true for most practical systems).

Frequency detection can be modeled by differentiating the output from a phase detection process. Phase detection usually requires limiting of the IF output. Limiting is used because the receiver circuitry employed for extraction of the phase variation also responds to amplitude variation. Thus, the amplitude variation must be removed before phase detection. For modeling purposes, phase detection was accomplished using the characteristics of an ideal phase detector. This process converts the phase variation of the composite signal relative to the IF center frequency to a baseband time waveform. Typical FM receiver systems are designed so the frequency response of the detector is linear over at least the IF bandwidth. In the SSIM, frequency detection is accomplished by taking the derivative of the phase function using a digital differentiating filter.

Baseband Amplifier

Baseband filtering is employed in receivers to eliminate undesired signals in frequency ranges outside the desired baseband signal bandwidth. These undesired signals result from interference and/or noise processed through the receiver with the desired signal, and from distortion introduced by the receiver stages themselves. Only the normalized frequency-dependent filter characteristics of the baseband amplifier are required in the SSIM. The rationale for this is the same as that explained for the IF amplifier modeling.

Bandpass filters are normally used for baseband filtering. In addition, FM systems may employ a de-emphasis filter at baseband; however, no provision was made for including a de-emphasis filter in the SSIM. The model uses a bandpass baseband filter.

Feedback Circuitry

Feedback circuitry can be grouped into two categories. The first category is designed to maintain a constant signal level. These circuits are generally classified as automatic gain control (AGC) or automatic volume control (AVC). The second basic type is employed in synchronous phase or frequency detection circuitry termed automatic frequency control (AFC). Simulation of either of these categories of feedback is actually not required in most compatibility analysis problems. These feedback effects were not simulated in the SSIM.

For the common AGC case, it can be shown that the AGC has no appreciable degradation effect unless the performance of the communication system has been substantially reduced. This does not hold for certain receivers which involve digital signal processing wherein the automatic control of the signal level is an integral part of the signal processing.

The modeling of the feedback loop in frequency detection systems is not required when phase lock is being maintained. Since phase lock is maintained when system performance is high, the model can be used to calculate performance at low levels of degradation in FM systems.

Decision Mechanism

The baseband output signal contains the desired and undesired signals. The undesired signal results from interference and/or noise processed through the receiver with the desired signal, and from distortion introduced by the receiver stages. The output waveform of the model represents the receiver system output waveform.

Reference 2 describes techniques for evaluating the performance of receiver systems. These descriptions are included herein as Appendix A. Articulation Index (AI) is a comprehensive measure of performance for voice communication systems. AI and the desired-to-undesired power ratio are calculated at the output of the SSIM. Each of these methods requires that the output desired signal be identified and separated from the total audio output. This is accomplished in the model by comparing the receiver output when only an input desired signal is used, with the output for the case of desired signal plus interference and noise at the input.

INPUT SIGNALS

In the SSIM, signals of various types are generated and combined to form the composite signal at the IF filter input. The absolute power levels used in generating the signals are irrelevant since only power ratios are specified as input parameters. All signals in the computer program are generated with an arbitrarily selected power level of 0.5 before application of any amplitude modulation or any power level adjustment to obtain an appropriate power ratio. Any unit of power may be associated with the 0.5 value as long as the same unit of power is associated with all signals. All signals are generated in the time domain, and the complex envelopes of the signals are presented next.

Desired Signals

AM Signal. The desired AM signal is generated as:

$$s_i(t) = 1 + m_a \cos(2\pi f_d t) \tag{1}$$

where

m_a = the equivalent amplitude modulation index for a tone modulating signal (30% of the corresponding value for a voice modulating signal; Kravitz and Lemke[4])

f_d = the frequency of the modulating signal.

FM Signal. The desired FM signal is generated as:

$$s_i(t) = \exp\left[j \frac{f_p}{f_d} \sin(2\pi f_d t)\right] \tag{2}$$

where

f_p = the equivalent peak frequency deviation for a tone modulating signal (60% of the corresponding value for a voice modulating signal; see Reference 4).

Undesired Signals

Direct-Sequence Binary PSK (DS/PSK) Signal. During a given chip interval (the reciprocal of R_c, the chip rate), the DS/PSK signal is expressed as:

---

[4]Kravitz, F., Lemke, M., *Communication/Electronics Receiver Performance Degradation Handbook* (Second Edition), ESD-TR-75-013, ECAC, Annapolis, MD, August 1975.

$$i(t) = K \exp\,(2\pi\Delta ft + \theta) \qquad (3)$$

where θ takes on the value 0 or π. At the beginning of each chip interval, the value of θ is randomly selected. In Equation 3, Δf represents the off-tuning between the interfering transmitter and the victim receiver, i.e., $\Delta f = f_{c_i} - f_{c_s}$ where f_{c_i} and f_{c_s} are, respectively, the interfering and desired signal carrier frequencies. Also in Equation 3, K is a multiplying constant which serves to adjust the level of the interference such that the specified signal-to-interference ratio at the IF filter input is obtained. The factor K is given by:

$$K = \sqrt{\frac{S}{I}\,10^{-(SIR/10)}} \qquad (4)$$

where

S = the desired signal mean power

I = *the interfering signal mean power before* adjustment

SIR = the specified mean signal-to-mean interference power ratio at the IF filter input, in dB.

For the AM desired signal, S may be obtained from:

$$S = \frac{1}{2}\left(1 + \frac{m_a^2}{2}\right) \qquad (5)$$

whereas for the FM desired signal, S = 0.5, and for the DS/PSK interfering signal, I = 0.5.

Direct-Sequence MSK (DS/MSK) Signal. As derived by Kivett,[5] the complex envelope for a DS/MSK signal can be expressed as:

$$i(t) = K L_Q \exp \left\{ \left[0.5 \; (L_I/L_Q) \; R_c + 2\pi\Delta f \right] t \right\} \qquad (6)$$

where

R_c = the chip rate

L_I and L_Q = keying sequences offset from each other in time by $1/R_c$ and such that each takes on the value +1 or -1 and each is keyed at the rate of $R_c/2$

Δf = the off-tuning

K = the interfering signal level adjustment factor.

In the model, Equation 6 is used to generate the DS/MSK signal with L_I and L_Q randomly selected at times $(2m + 1)/R_c$ and $2m/R_c$, respectively, where m is a nonnegative integer. In computing K, the value of I for the DS/MSK signal is 0.5.

Frequency-Hopping Signal. During a given dwell time (the amount of time the FH signal remains on each frequency before hopping), the FH signal is generated as:

$$i(t) = K \exp \left[2\pi (f_n + \Delta f) t \right] \qquad (7)$$

where

f_n = one of n_f possible frequencies

Δf = the off-tuning

K = the interfering signal level adjustment factor.

---

[5]Kivett, J. A., *Wideband Command and Control Modem (Waveform and Modem Conceptual Design Study)*, RADC-TR-73-12, December 1972, Appendix A.

The number of frequencies, n_f, the separation between adjacent frequencies, f_s, and the dwell time, τ_f, are input parameters to the simulation program. It is assumed that all pairs of adjacent frequencies have the same frequency separation and that the FH signal is always on one of the frequencies. Let $f_1, f_2, \ldots, f_{n_f}$ denote the frequencies to be hopped. If n_f is odd, the model selects frequency $f_{(n_f + 1)/2}$ to be on-tune with the carrier frequency of the desired signal, whereas if n_f is even, the frequency $f_{\frac{n_f}{2} + 1}$ is selected. The frequencies may be shifted from these values by appropriately selecting the value of Δf. The hopped frequencies are randomly selected with the restriction that during the time for each problem run (i.e., each articulation index computation), the number of times a frequency occurs will differ from the number of times any other frequency occurs by at most one. Also, all frequencies being hopped do not have to be processed. An input parameter, f_{max}, may be used to process only those frequencies, $f_n + \Delta f$, such that $|f_n + \Delta f| \leq f_{max}$. When a frequency greater than f_{max} occurs, i(t) is set equal to zero for the dwell time. In computing K for the FH signal, I = 0.5.

Hybrid FH/DS Signal. The normalized (i.e., K = 1) complex envelope of a hybrid FH/DS signal is generated as the product of the normalized complex envelopes of an FH signal and a DS signal. The complex envelope of an FH/DS signal is then given by $i(t) = K\, i_{FH}(t) \cdot i_{DS}(t)$, where K is computed using Equation 4 with I = 0.5, $i_{FH}(t)$ represents the right side of Equation 7 with K = 1, and $i_{DS}(t)$ represents, respectively, the right side of Equation 3 or Equation 6 with K = 1 for an FH/(DS/PSK) or an FH/(DS/MSK) signal. The FH and DS portions of the hybrid signal are generated as previously described.

Pulsed Signal. During each pulse repetition interval, the pulsed interference signal is generated as:

$$i(t) = \begin{cases} 0 & -\frac{1}{2\ PRF} \leq t \leq -\frac{\tau}{2} \\ \hat{K} \exp(2\pi\Delta ft) & -\frac{\tau}{2} \leq t \leq \frac{\tau}{2} \\ 0 & \frac{\tau}{2} \leq t \leq \frac{1}{2\ PRF} \end{cases} \tag{8}$$

where

τ = the pulse width

PRF = the pulse repetition frequency

$\hat{K}$ = the interfering signal level adjustment factor.

In the case of pulsed interference, the SIR specified at the IF filter input is the signal-to-peak interference power ratio, and the adjustment factor is given by:

$$\hat{K} = \sqrt{\frac{S}{\hat{I}}\ 10^{-SIR/10}} \tag{9}$$

where

$\hat{I}$ = the peak power of the interfering signal before adjustment.

The value of $\hat{I}$ used in the SSIM is 0.5.

White Gaussian Noise. Time samples of white Gaussian noise are generated by separately generating samples of a phase function uniformly distributed over the interval from zero to 2π and samples of a Rayleigh-distributed envelope function. The complex envelope of the noise voltage is given by:

$$n(t) = a(t)e^{j\theta(t)} \tag{10}$$

A time sample of $\theta(t)$, $\theta(kT)$, is computed as:

$$\theta(kT) = 2\pi r_u \tag{11}$$

where

r_u = a random number from a uniform distribution of random numbers from zero to one.

The Rayleigh amplitude probability distribution (APD) is used in the technique[6] to generate an envelope sample. For white Gaussian noise of average power σ^2, the noise voltage envelope has the Rayleigh APD function:

$$P_A(a) = e^{-a^2/(2\sigma^2)} \tag{12}$$

Using Equation 12, the desired *envelope* sample is obtained as:

$$a(kT) = \sqrt{-2\sigma^2 \ln \left\{ P_A \left[a(kT)\right] \right\}} \tag{13}$$

Since $P_A(a)$ is an APD, $0 \leq P_A(a) \leq 1$. Values of $P_A\left[a(kT)\right]$ are obtained from a uniform distribution of random numbers from zero to one and Equation 13 is used to compute the value of the envelope sample. The complex envelope noise sample is then generated from $a(kT)$ and $\theta(kT)$ as:

$$n(kT) = a(kT)e^{j\theta(kT)} \tag{14}$$

In using the model, the decibel value of signal-to-noise ratio, SNR, is specified at the output of the IF filter. The noise power,

---

[6] Hull, T. E., and Dobell, A. R., "Random Number Generators," *SIAM Review,* Vol. 4, No. 3, July 1962.

σ^2, is distributed in frequency over $\pm 1/(2T)$ when the time samples are generated with sampling interval T. If S represents the desired signal mean power used in the simulation program and BW_N denotes the noise bandwidth of the IF filter, then the SNR at the IF filter output is given by:

$$10^{SNR/10} = \frac{S}{\sigma^2 T \; BW_N} \tag{15}$$

Using Equation 15, the appropriate value of σ^2 for the specified SNR is then calculated as:

$$\sigma^2 = \frac{S}{T \; BW_N \; 10^{SNR/10}} \tag{16}$$

The value of BW_N is computed within the simulation program and the sampling frequency, 1/T, is an input parameter.

RECEIVER MODELS

From the block diagram of the overall simulation model shown in Figure 2, it is seen that the receiver model includes the IF filter, detector, and baseband filter. The input signal to the IF filter is generated as a series of discrete time samples, and the samples are passed one at a time through the filters and detector. For each receiver stage, a new output sample is generated for each new input sample and this new output sample then becomes the new input sample for next stage in the receiver model.

This process continues with single samples being processed through the IF filter, detector and baseband filter. After an appropriate amount of time has elapsed to allow for the time delay through the receiver and for the transient settling time, the next 8192 time samples at the output of the baseband filter are stored in a register. The amount of time allowed to elapse is at least five times the composite midband group delay of the IF and baseband filters. The stored time signal is transformed to the frequency domain where the desired signal is extracted from the composite signal and the performance of the receiver is evaluated.

IF Filter

The IF filter is modeled as an nth order digital lowpass Butterworth filter. The digital transfer function of the filter is obtained by applying the standard z-transformation to the transfer function of the corresponding analog filter. Implementation of the digital filter is then accomplished in the parallel form. Appendix B includes the details of obtaining the digital transfer function of the filter and the associated input-output relationship. The amplitude versus frequency and group delay versus frequency characteristics for Butterworth filters have been published and are available.[7]

Using the input-output relationship for the filter as given in Appendix B, the IF filter model determines the output signal time sample, $y(kT)$, from the present input signal time sample, $x(kT)$, and appropriate past input and output signal time samples. The composite input signal time sample is the sum of the desired signal, noise, and interfering signal time samples, i.e., $x(kT) = s(kT) + n(kT) + i(kT)$. The output signal time sample is then applied as the input to the detector.

Detector

Detection is accomplished differently for the AM and FM systems. The signal at the detector input is viewed as a joint AM-PM signal

[7] *A Handbook on Electrical Filters,* White Electromagnetics, Inc., Rockville, MD, 1963.

with the present signal sample being $y(kT) = a(kT) \exp[j\theta(kT)]$ where $a(kT)$ includes the amplitude modulation and $\theta(kT)$ includes the angle modulation. Let the real and imaginary components of $y(kT)$ be, respectively, $y_1(kT) = a(kT) \cos \theta(kT)$ and $y_2(kT) = a(kT) \sin \theta(kT)$. The AM detector output signal sample, $w(kT)$, is obtained from:

$$w(kT) = \sqrt{y_1^2(kT) + y_2^2(kT)} = a(kT) \tag{17}$$

The FM detector is modeled as shown in Figure 3. This model of an ideal phase detector followed by a differentiator is equivalent to a limiter-discriminator combination. The present output sample from the ideal phase detector is given by:

$$\begin{aligned} \theta(kT) = \theta[(k-1)T] &+ \frac{1}{2\pi} \arctan\left[\frac{y_2(kT)}{y_1(kT)}\right] \\ &- \frac{1}{2\pi} \arctan\left\{\frac{y_2[(k-1)T]}{y_1[(k-1)T]}\right\} \end{aligned} \tag{18}$$

Thus, the phase variation with respect to the carrier phase is accumulated from time sample to time sample. Since the inverse tangent function is used to compute instantaneous phase, this computation is independent of the modulus of $y(kT)$, and therefore effectively incorporates hard-limiting. Furthermore, since the output is only the phase variation with respect to the fundamental carrier phase, this computation also effectively incorporates ideal zonal filtering following the hard-limiting. This model includes a reasonable representation of a hard limiter, providing that the input signal to the limiter has a relatively narrow bandwidth compared with the carrier frequency of the desired signal, and this is typically the case.

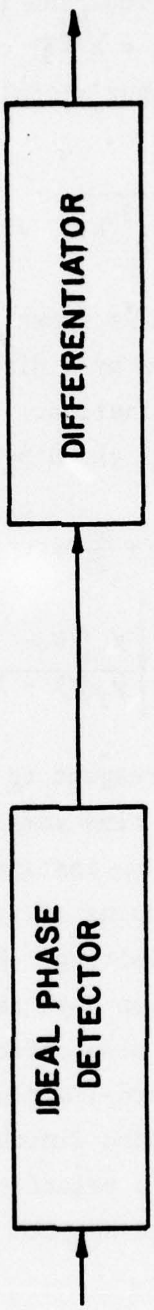


Figure 3. The FM detector (limiter-discriminator) model.

The phase function at the output of the ideal phase detector is then processed through a differentiator to obtain the discriminator action, i.e., an output proportional to instantaneous frequency. The differentiator used in the model is a 21-tap nonrecursive digital differentiator.[8]

This differentiator is within 1% of true derivative action for frequencies up to 87% of half the sampling frequency. The frequency characteristic of this digital differentiator is depicted in Figure 4. The abcissa values are fractions of half the sampling frequency, i.e., 1.0 corresponds to half the sampling frequency. The present output time sample from the differentiator, w(kT), is obtained from:

$$w(kT) = \frac{1}{T}\sum_{j=0}^{9} c_j \Big\{ y\big[(k - j)T\big] - y\big[(k - 20 + j)T\big] \Big\} \tag{19}$$

where T is the sampling interval, y(kT) is the present input sample, y[(k - j)T] and y[(k - 20 + j)T] are the jth previous and (20 - j)th previous input samples and the c_j are the coefficients of the digital differentiator. The coefficients, c_j, are given in TABLE 1. The output signal time sample from the detector, w(kT), is applied as the input to the baseband filter.

Baseband Filter

The baseband filter is modeled as a digital bandpass Butterworth filter. Applying a lowpass-to-bandpass transformation to the transfer function of an nth order, normalized lowpass filter, the transfer

[8]Kuo, F. F. and Kaiser, J. F., Eds., *System Analysis by Digital Computer,* John Wiley and Sons, New York, 1966, Chapter 7.

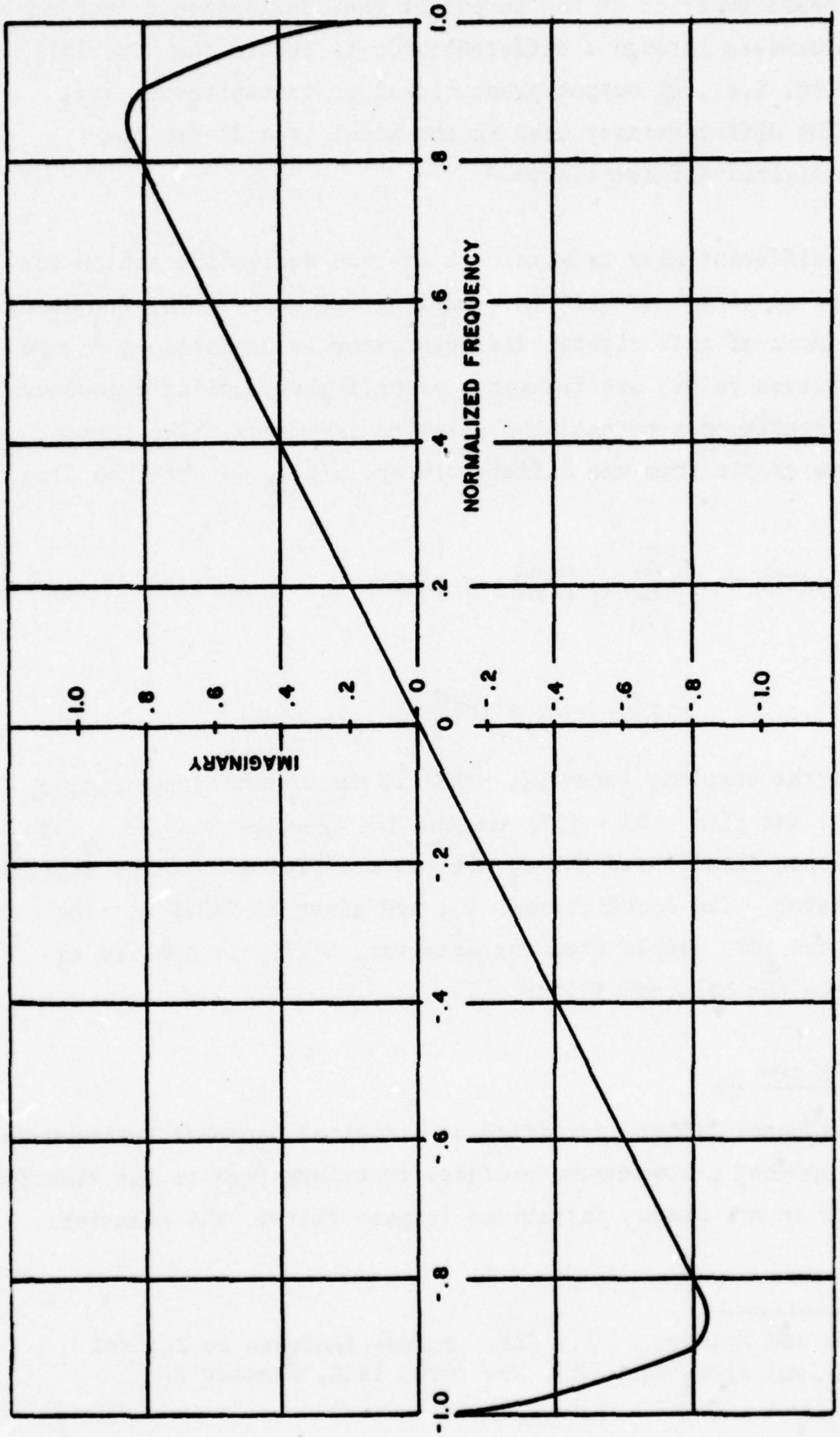


Figure 4. Characteristic of the differentiator.

TABLE 1

DIGITAL DIFFERENTIATOR COEFFICIENTS

c_0 = -0.009885	c_1 = 0.025113
c_2 = -0.038097	c_3 = 0.059197
c_4 = -0.088958	c_5 = 0.013062
c_6 = -0.191	c_7 = 0.28729
c_8 = -0.468363	c_9 = 0.983896

function of the analog bandpass filter is obtained first. The standard z-transformation is then applied to the transfer function of the analog bandpass filter to obtain the digital transfer function of the filter. Implementation of the digital filter is then accomplished in the parallel form. Appendix B includes the details of obtaining the digital transfer function of the filter and the associated input-output relationship. Using the input-output relationship for the filter as given in Appendix B, the baseband filter model determines the output signal time sample, v(kT), from the present input signal time sample, w(kT), and appropriate past input and output time samples.

Output Desired and Undesired Signal Identification

Desired signal identification is accomplished in the frequency domain by correlating the composite signal baseband filter output, V(f), with the desired signal baseband filter output, $S_o(f)$ (see Reference 2). The frequency function, $S_o(f)$, is obtained by processing only the desired signal through the receiver and then taking the discrete Fourier transform of the output time signal using a fast Fourier transform (FFT) algorithm. The FFT algorithm used is described in Reference 2.

The frequency function, $V(f)$, is obtained in the same way when the composite signal is processed through the receiver. In each case, time samples are processed through the receiver and, after an amount of time equal to at least five times the composite midband group delay has elapsed, the next 8192 time samples at the output of the baseband filter are stored in a register. This stored time signal is transformed to the frequency domain.

The correlation process is simplified since the desired output signal is a tone at frequency f_d. Figure 5 shows the projection of the $V(f_d)$ phasor on the $S_o(f_d)$ phasor. The length, L, of the projection may be positive or negative and is given by:

$$L = \frac{\text{Re}\left[V(f_d)\ S_o^*(f_d)\right]}{|S_o(f_d)|} \tag{20}$$

where

$V(f_d)$ = the complex frequency coefficient of the composite output signal at frequency f_d.

$S_o(f_d)$ = the previously stored complex frequency coefficient representing the desired tone

$*$ = the complex conjugate

$|\,|$ = the modulus of

Re = the real part of.

When undesired components are present at the baseband filter output, the desired tone frequency coefficient will usually be different from $S_o(f_d)$. The desired signal portion of the composite output frequency function will be zero except at the desired tone frequency, where it has the modulus:

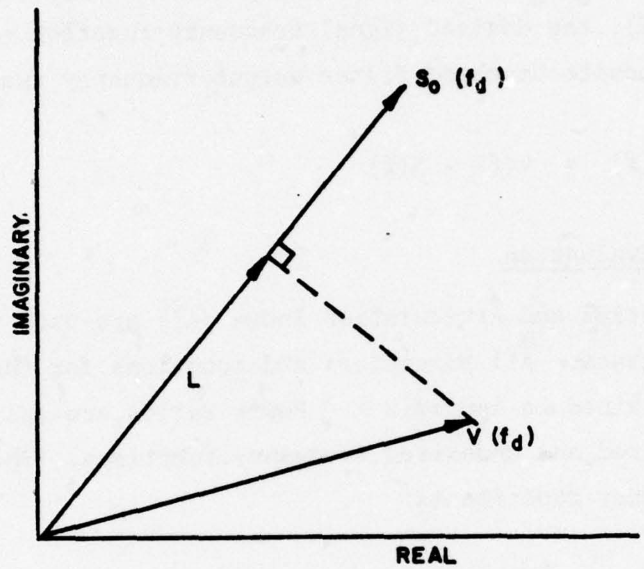


Figure 5. Phasor diagram used in determining the desired output tone amplitude.

$$|S(f_d)| = \begin{cases} |S_o(f_d)| & \text{if } L \geq |S_o(f_d)| \\ L & \text{if } |S_o(f_d)| > L > 0 \\ 0 & \text{if } L < 0 \end{cases} \tag{21}$$

The phase angle of the $S(f_d)$ frequency component is the same as the phase angle of $S_o(f_d)$. To obtain the undesired signal frequency function, $U(f)$, the desired signal frequency function is subtracted from the composite baseband filter output frequency function, i.e.:

$$U(f) = V(f) - S(f) \tag{22}$$

Performance Evaluation

Power ratios and Articulation Index (AI) are used to evaluate model performance. All parameters and equations for the AI calculation are contained in Appendix A. Power ratios are easily obtained from the desired and undesired frequency functions. The desired signal frequency function is:

$$S(f) = \begin{cases} S(f_d) & \text{for } f = f_d \\ S^*(f_d) & \text{for } f = -f_d \\ 0 & \text{elsewhere} \end{cases} \tag{23}$$

The desired signal power is:

$$P_s = 2\,|S(f_d)|^2 \tag{24}$$

The undesired frequency function, $U(f)$, can have frequency components at multiples of f_Δ and the interference power is:

$$P_u = |U(0)|^2 + 2\sum_{i=1}^{N/2} |U(if_\Delta)|^2 . \tag{25}$$

where

f_Δ = the fundamental frequency as determined by the sampling rate ($f_\Delta = \frac{1}{NT}$)

N = the number of sample points.

The output power ratio in decibels is:

$$(P_s/P_u)_{dB} = 10 \log_{10} (P_s/P_u) \tag{26}$$

SECTION 3

SIMULATION RESULTS

AM AND FM SYSTEMS SIMULATED

The model developed was used to compare the performance of an AM and an FM voice communication system operating in the presence of direct-sequence interference with the performance of the same system operating in the presence of white Gaussian noise. Similarly, the simulation program was used to compare the performance of each of these systems operating in the presence of frequency-hopping or frequency-hopping/direct-sequence interference with the performance of the same system operating in the presence of pulsed interference. Representative AM and FM voice systems were selected for analysis from Reference 4. The significant characteristics of these systems are given in TABLE 2.

TABLE 2

AM AND FM SYSTEM CHARACTERISTICS

Modulation Type	Receiver IF 3 dB Bandwidth (kHz)	Modulation Index	Peak Deviation (kHz)
A3	8	1.0[a]	—
F3	16	1.67	±5

[a]This number corresponds to a root-mean-square (rms) modulation index of 0.3 as given in Reference 4.

For each receiver, the IF filter was modeled as a 5-pole lowpass Butterworth filter and the baseband filter was modeled as a 12-pole bandpass Butterworth filter (i.e., the filter obtained when the lowpass-to-bandpass transformation is applied to a 6-pole lowpass filter). The baseband filter had lower and upper 3 dB cutoff frequencies of 300 Hz and 3500 Hz, respectively.

DIRECT-SEQUENCE INTERFERENCE CASES

The performance of each of the two receivers was obtained for a DS/PSK and a DS/MSK undesired signal and for white Gaussian noise. Each of the SS signals was on-tune and had a chip rate of 80 kbps. The sampling frequency used was 409.6 kHz. The results were plotted in terms of articulation index versus S/I or S/N ratio as appropriate, where S is the mean desired signal power and I and N are, respectively, the mean interference power and mean noise power within the IF filter. For each receiver, the results for each SS signal are plotted along with the corresponding results for white Gaussian noise. The results for the AM system with DS/PSK and DS/MSK undesired signals are shown in Figures 6 and 7, respectively. The results for the FM system with DS/PSK and DS/MSK undesired signals are depicted in Figures 8 and 9, respectively. These results indicate that, for conventional narrow-band AM and FM voice communication systems, approximately the same performance is obtained for a given amount of undesired power within the IF filter resulting from a DS signal or white Gaussian noise of the same bandwidth.

FREQUENCY-HOPPING INTERFERENCE CASES

The performance of each of the two receivers was obtained for two FH signals with different dwell times. Each FH signal hopped among 20 equally spaced frequencies with adjacent frequencies being 60 kHz apart. The frequencies of the FH signal were such that one frequency was on-tune with the receiver tuned frequency, ten frequencies were below, and the remaining nine frequencies were above the receiver tuned frequency. Thus only one of the hopped frequencies occurred within the IF bandwidth of each receiver. The FH signals were pseudorandomly hopped with dwell times of 200 μs in one case and 50 μs in the other case. With 20 frequencies and a 200-μs dwell time, the mean rate at which a given frequency occurs is 250 occurrences per second (ops). With the 50-μs dwell time, the mean frequency occurrence rate (FOR) is 1000 ops. Using a sampling frequency of 409.6 kHz, receiver

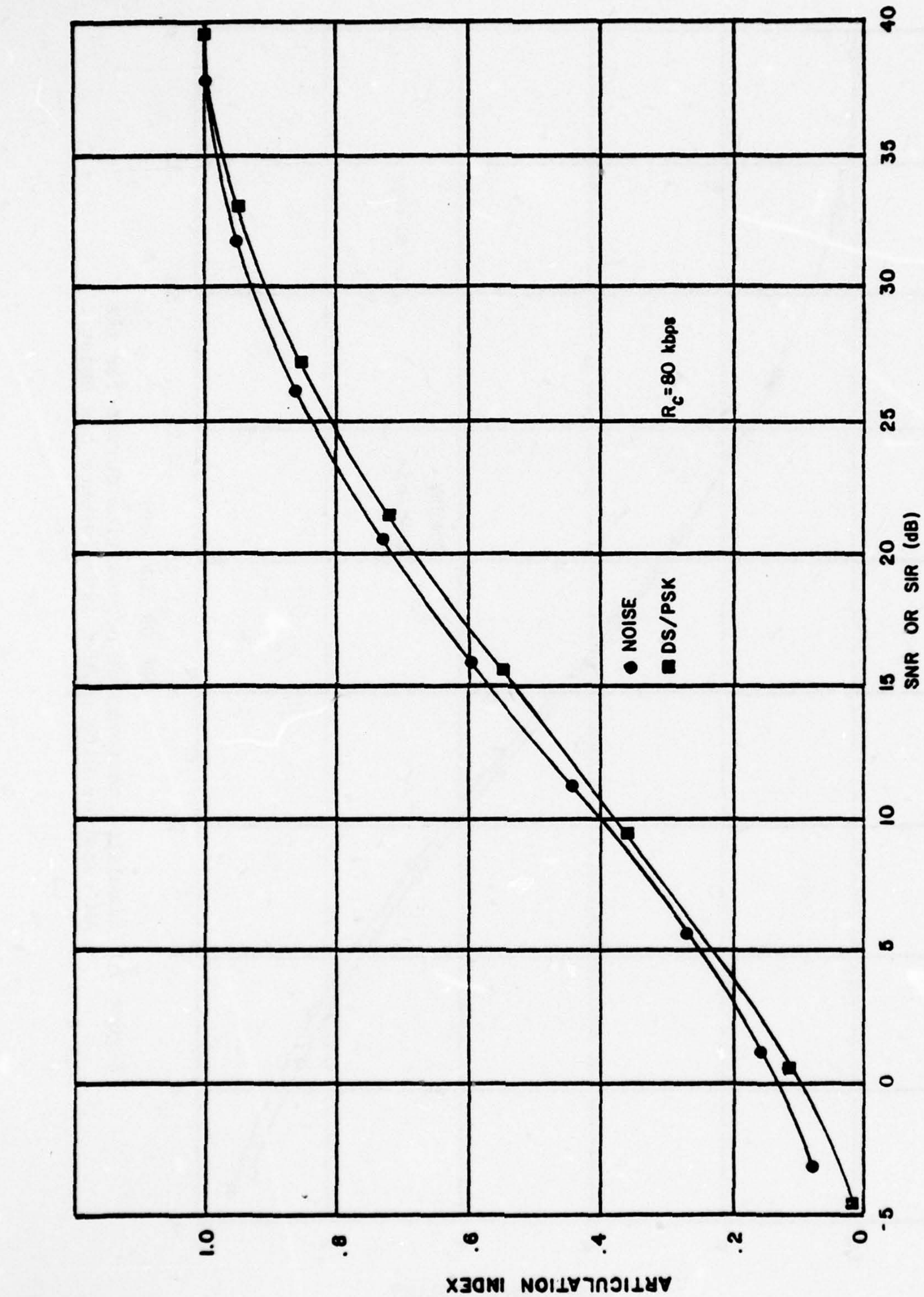


Figure 6. Simulated performance degradation curves for the AM receiver with DS/PSK interference and noise.

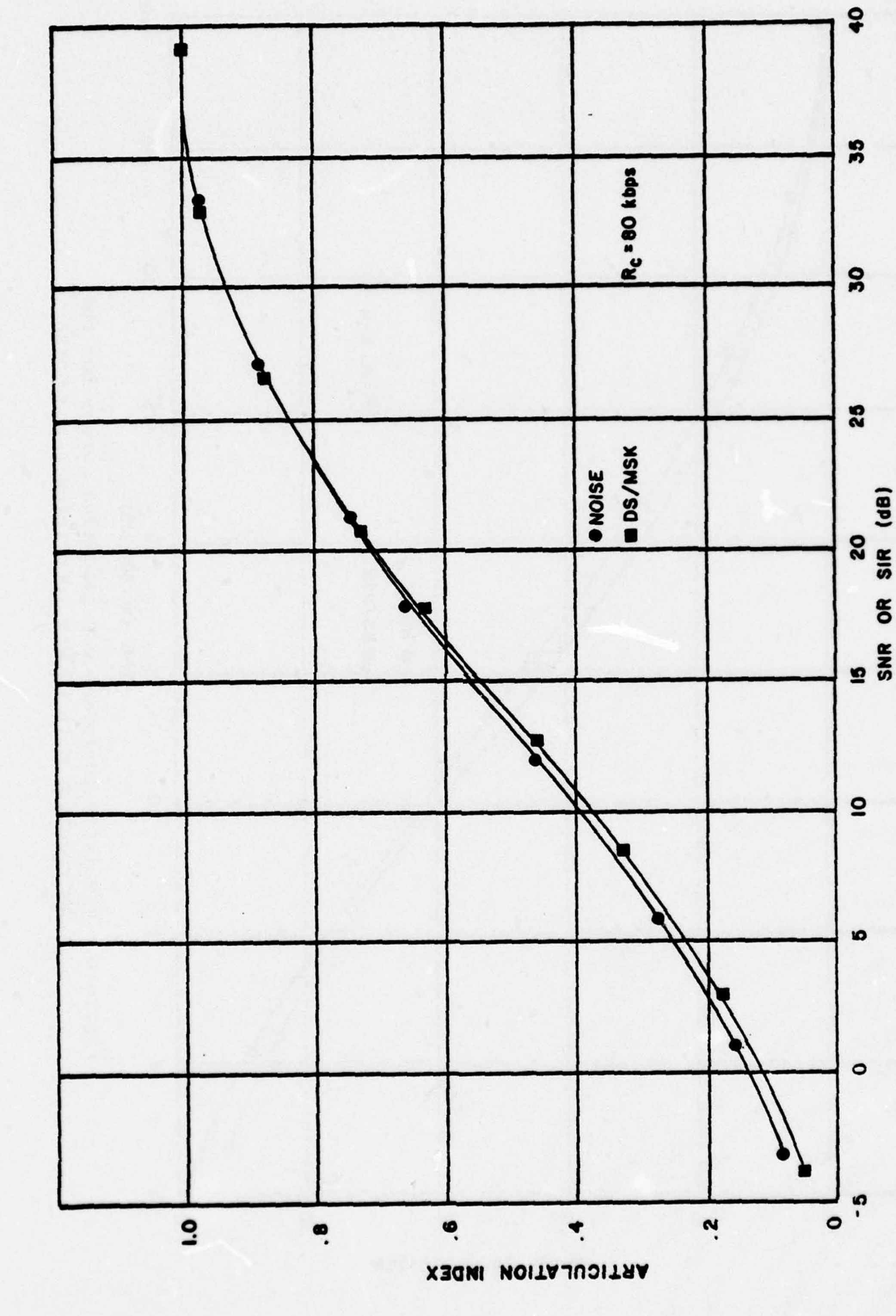


Figure 7. Simulated performance degradation curves for the AM receiver with DS/MSK interference and noise.

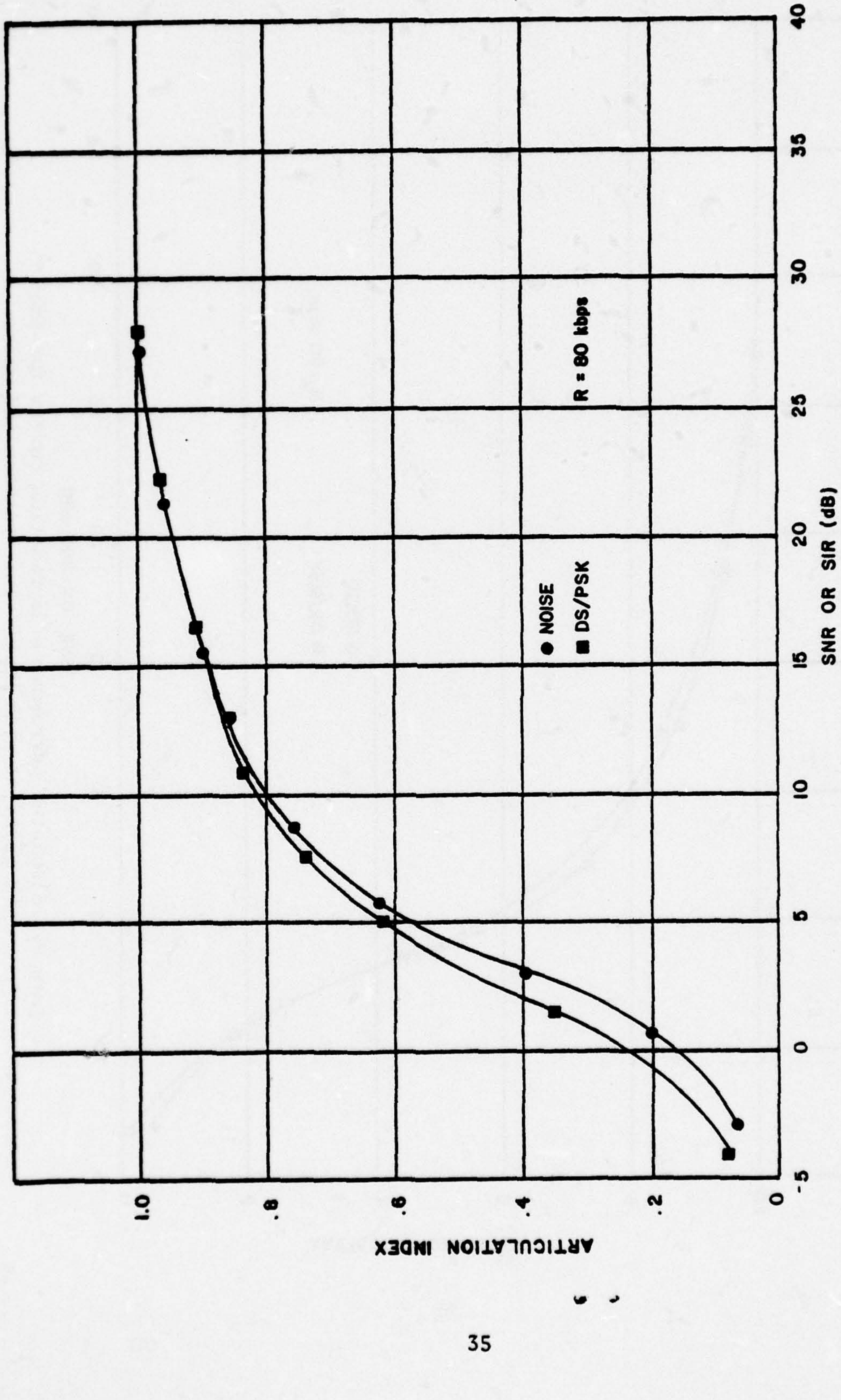


Figure 8. Simulated performance degradation curves for the FM receiver with DS/PSK interference and noise.

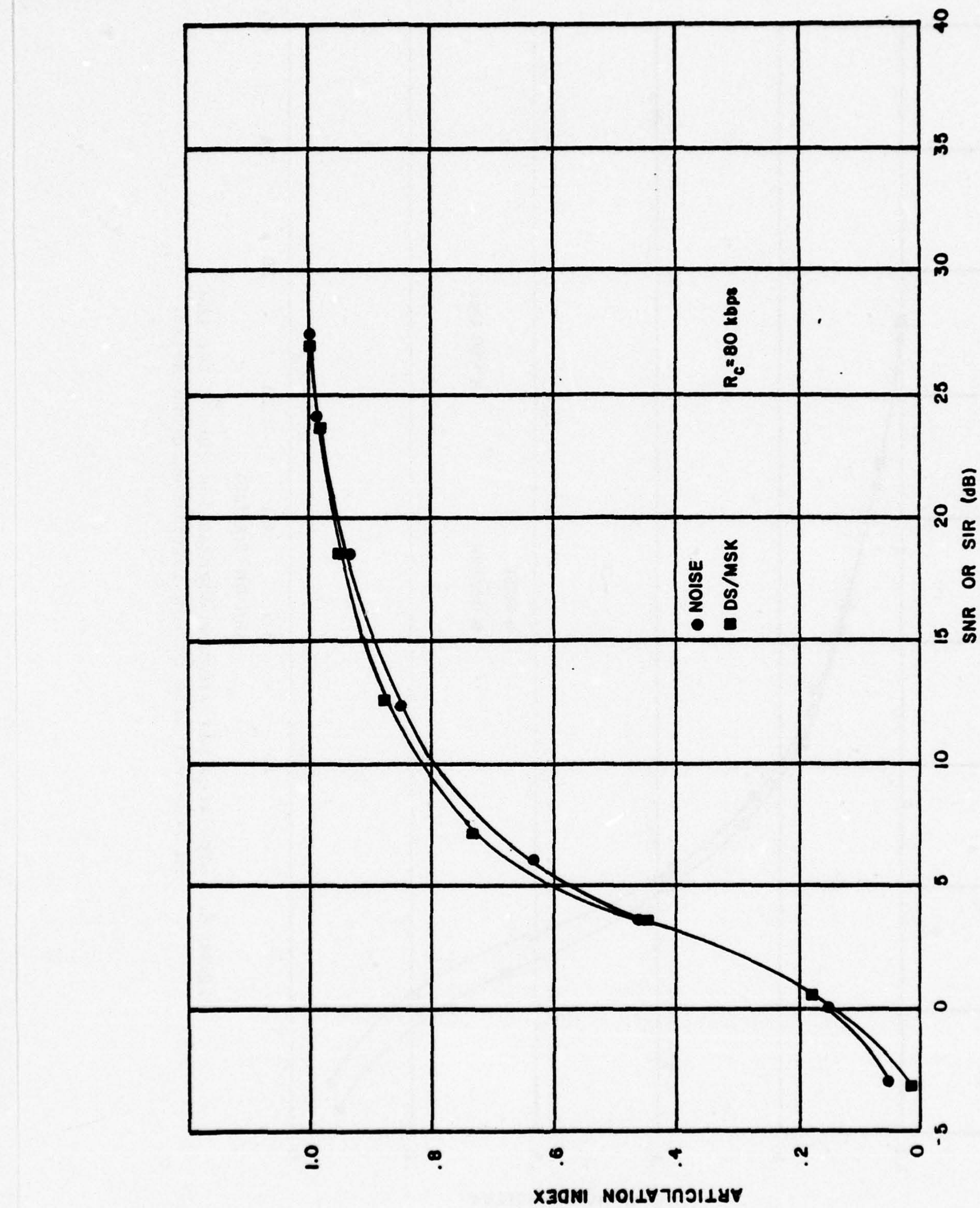


Figure 9. Simulated performance degradation curves for the FM receiver with DS/MSK interference and noise.

performance was obtained for two cases. In the first case, only the on-tune frequency of the FH signal was processed whenever it occurred, with the input interference signal set to zero whenever any other frequency occurred. In the second case, the on-tune frequency and the adjacent frequencies 60 kHz above and below the receiver tuned frequency were processed whenever they occurred, with the input interference signal set to zero whenever any other frequency occurred. The difference between the results for the two cases was negligible, i.e., the contributions of the FH signal when using the frequencies 60 kHz above and below the receiver tuned frequency to performance degradation were negligible compared to the contribution of the FH signal when using the on-tune frequency. Therefore, only the on-tune frequency was processed. While the entire FH signal could have been processed, a much larger sampling frequency would have been required and program execution time would have been corresponsingly greater.

For each receiver, the performance was also obtained for pulsed signals with 200-μs and 50-μs pulse widths and pulse repetition frequencies of 250 pps and 1000 pps, respectively. The performance of the AM receiver with the undesired FH signal having a 200-μs dwell time (250 ops FOR) and the results for the undesired pulsed signal having a 250 pps PRF are shown in Figure 10. The corresponding comparison for the AM receiver using the 50-μs dwell time (1000 ops FOR) for the FH signal and the 1000 pps PRF for the pulsed signal is depicted in Figure 11. The same two performance comparisons for the FM system are illustrated in Figures 12 and 13, respectively. These results indicate that, at typical operating levels, the performance of conventional narrowband AM and FM voice communication systems would be approximately the same for a given amount of undesired power within the IF filter resulting from an FH signal or a pulsed signal, such that the dwell time and mean frequency occurrence rate of the FH signal are, respectively, equal to the pulse width and pulse repetition frequency of the pulsed signal.

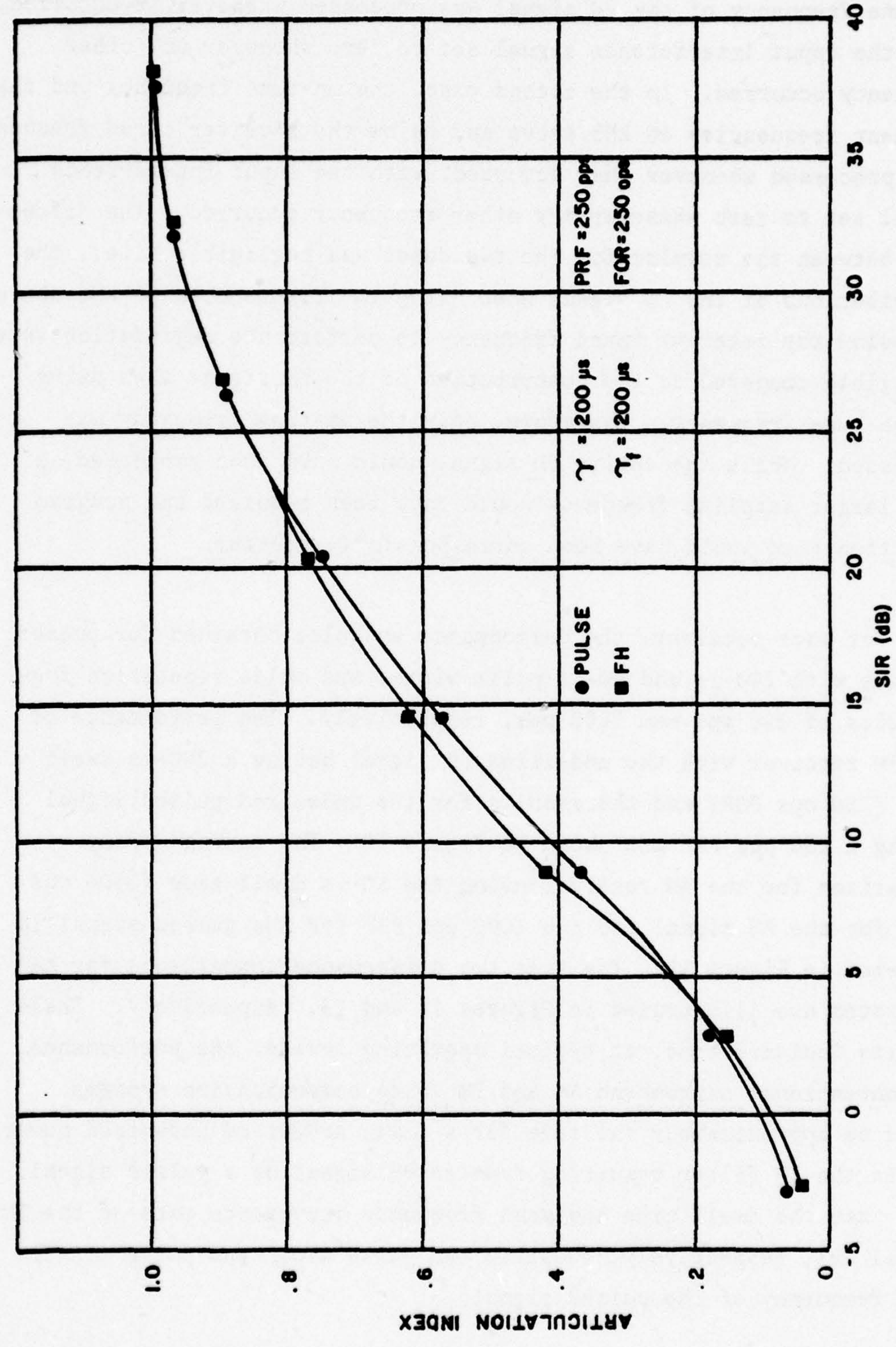


Figure 10. Simulated performance degradation curves for the AM receiver with FH and pulsed interference.

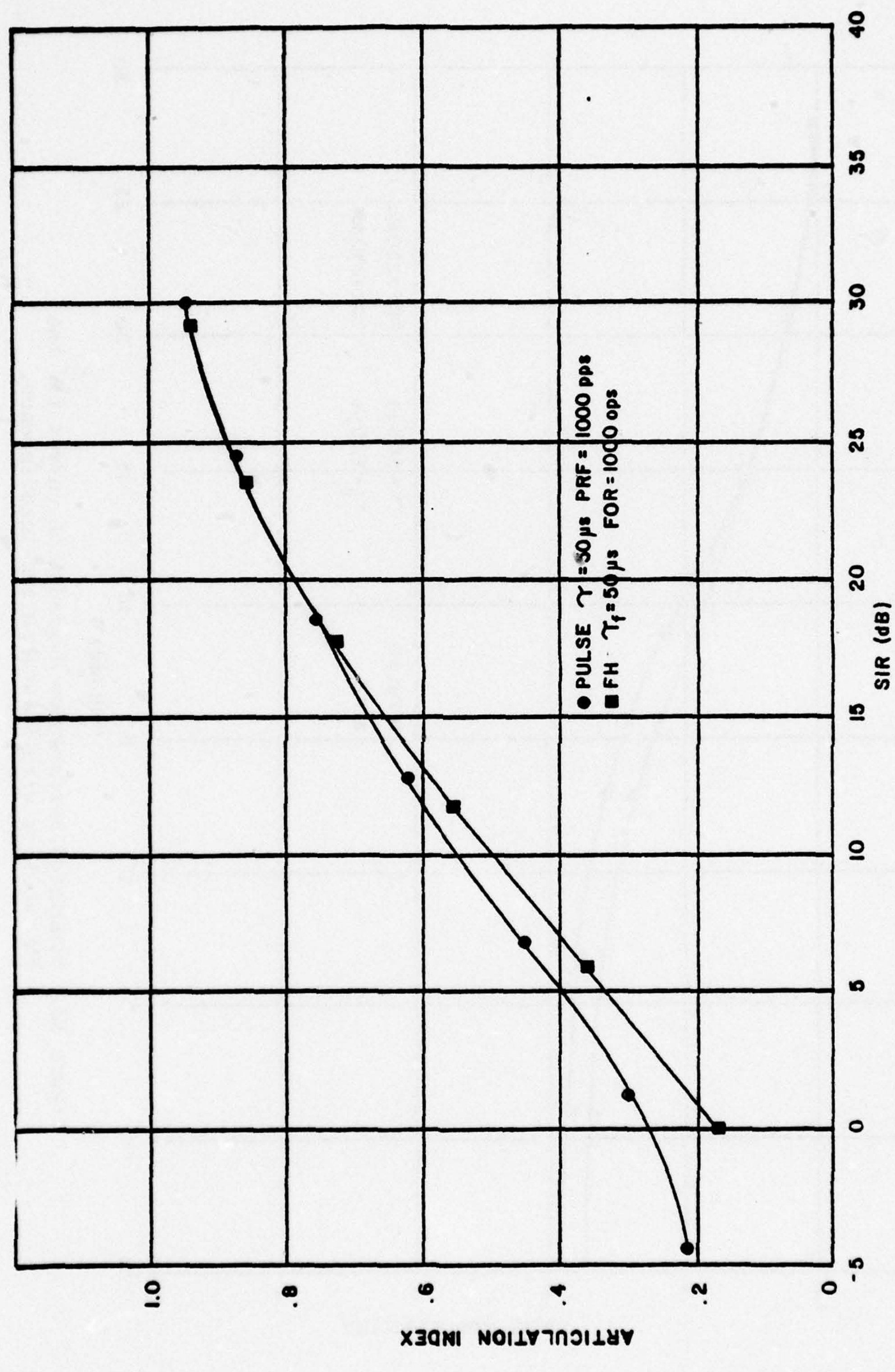


Figure 11. Simulated performance degradation curves for the AM receiver with FH and pulsed interference.

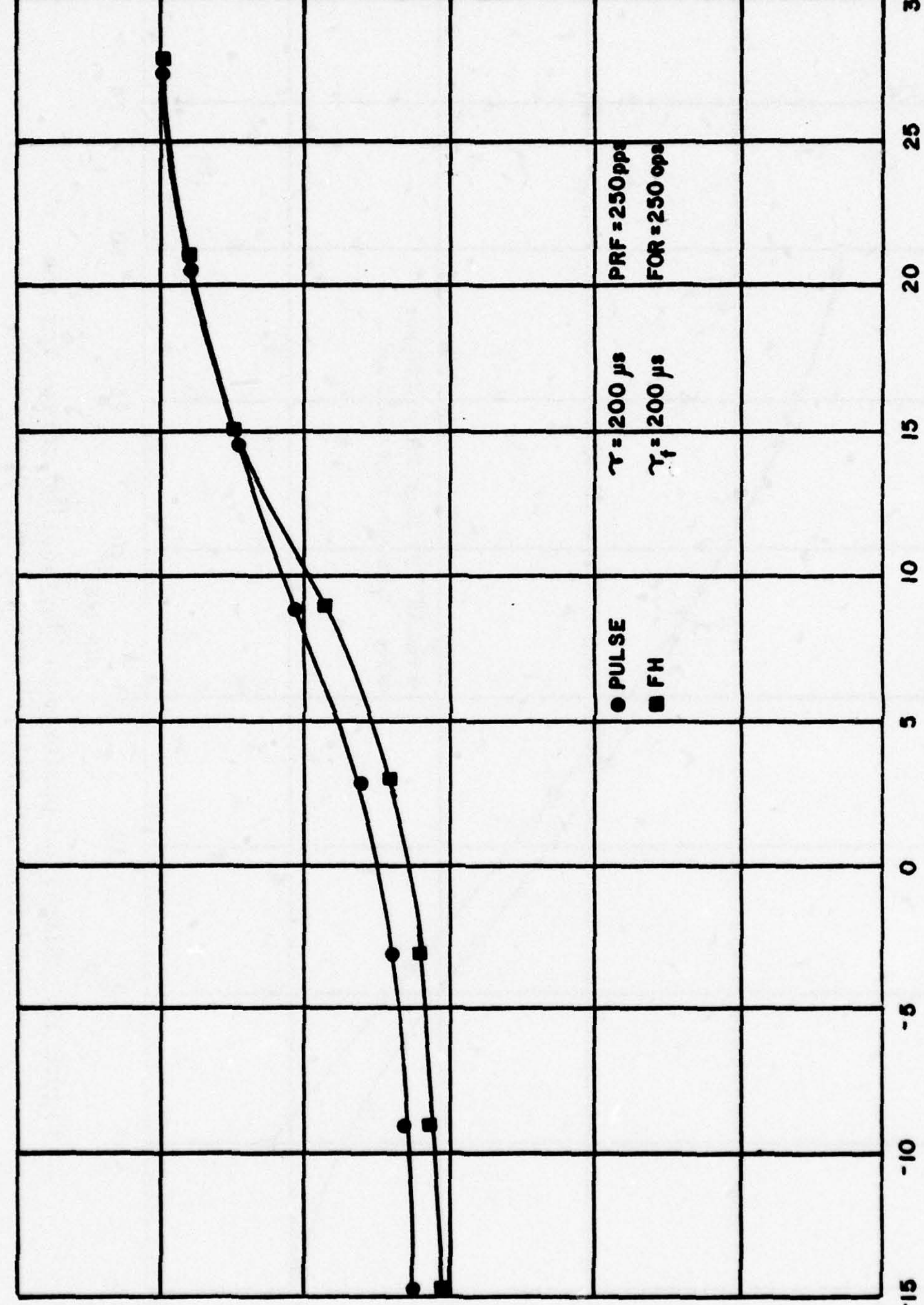


Figure 12. Simulated performance degradation curves for the FM receiver with FH and pulsed interference.

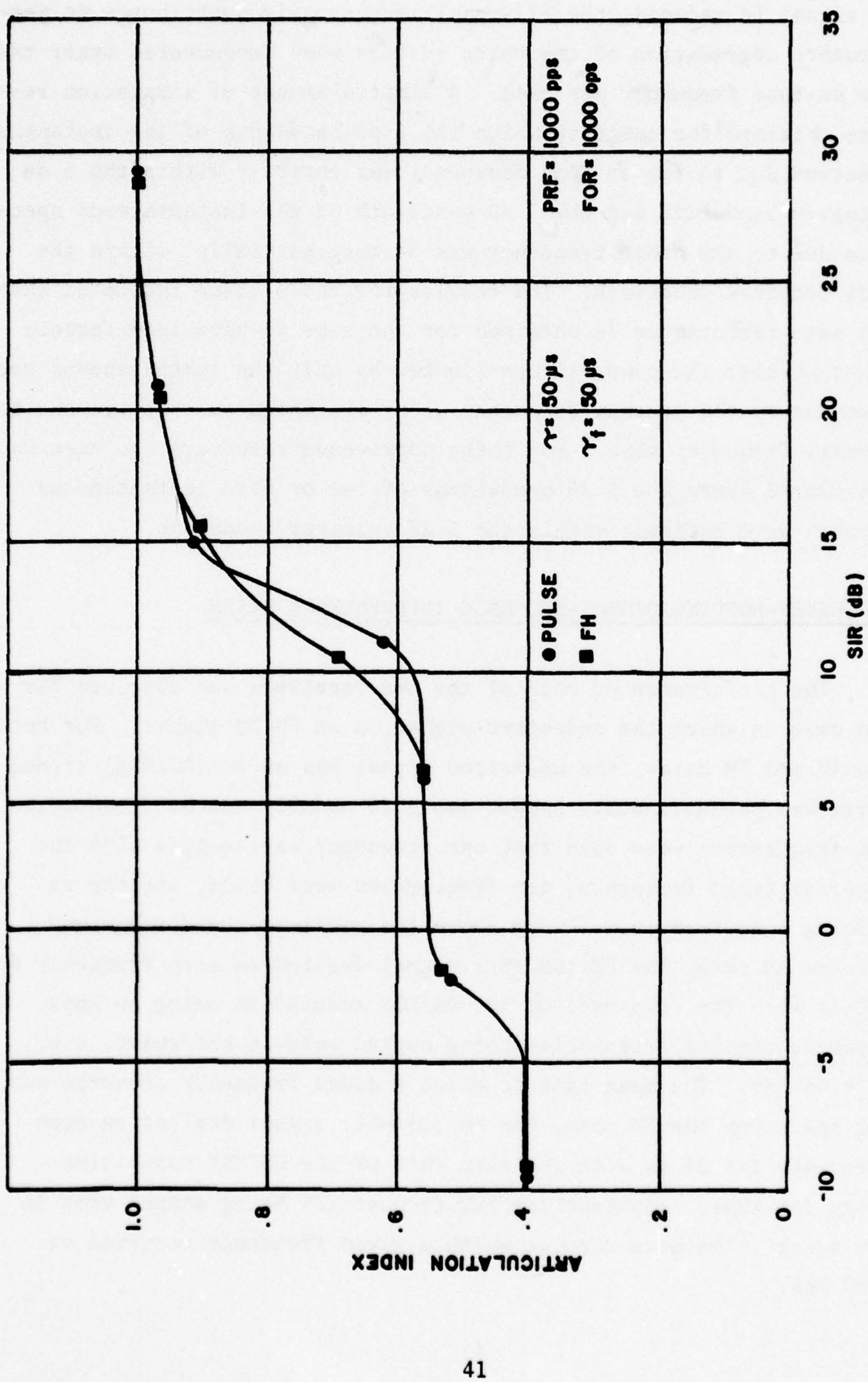


Figure 13. Simulated performance degradation curves for the FM receiver with FH and pulsed interference.

As the frequency separation between adjacent frequencies of the FH signal is reduced, the FH signal increasingly contributes to performance degradation of the voice systems when frequencies other than the on-tune frequency are used. A limited amount of simulation results were obtained for cases in which the 3 dB bandwidth of the instantaneous spectrum due to the on-tune frequency was entirely within the 3 dB receiver bandwidth and the 3 dB bandwidth of the instantaneous spectrum due to any other frequency was at most partially within the 3 dB receiver bandwidth. The results for these cases indicated that the same performance is obtained for the same in-band interference power whether the power is contributed by only the instantaneous spectrum due to the on-tune frequency or by instantaneous spectra due to several frequency hops. For these narrowband receivers, no case was considered where the 3 dB bandwidths of two or more instantaneous spectra were entirely within the 3 dB receiver bandwidth.

FREQUENCY-HOPPING/DIRECT-SEQUENCE INTERFERENCE CASES

The performance of each of the two receivers was obtained for the case in which the undesired signal is an FH/DS signal. For both the AM and FM cases, the undesired signal was an FH/(DS/MSK) signal which was pseudorandomly hopped among 20 equally spaced frequencies. The frequencies were such that one frequency was on-tune with the receiver tuned frequency, ten frequencies were below, and the remaining nine frequencies were above the receiver tuned frequency. For the AM case, the FH/(DS/MSK) signal dwelled on each frequency for 200 μs with the chip rate of the DS/MSK modulation being 80 kbps. Adjacent carrier frequencies being hopped were 48 kHz apart, i.e., f_s = 48 kHz. The mean rate at which a given frequency occurred was 250 ops. For the FM case, the FH/(DS/MSK) signal dwelled on each frequency for 50 μs with the chip rate of the DS/MSK modulation being 160 kbps. Adjacent carrier frequencies being hopped were 96 kHz apart. The mean rate at which a given frequency occurred was 1000 ops.

For both the AM and FM cases, the input interference signal was set to zero whenever any frequency other than the on-tune frequency or one of the immediately adjacent frequencies occurred. It was determined, by considering the theoretical spectrum, that this results in a negligible (approximately 0.01 dB) reduction in the average interference power contained in the bandwidth of the receiver. Also, the average power within the receiver bandwidth due to the on-tune frequency and the immediately adjacent frequencies above and below the on-tune frequency is approximately 0.25 dB above that due to only the on-tune frequency. Therefore the contributions to the average inband undesired power are small when the interfering signal is using any carrier frequency other than the on-tune frequency.

The performance of the AM receiver with the undesired FH/(DS/MSK) signal having a 200-μs dwell time, a 250-ops frequency occurrence rate and a 80-kbps DS chip rate is shown in Figure 14, along with the performance of the receiver for the case of an undesired pulsed signal with 200-μs pulse width and a PRF of 250 pps. Figure 15 illustrates the performance of the FM receiver with the undesired FH/(DS/MSK) signal having a 50-μs dwell time, a 1000-ops frequency occurrence rate and a DS chip rate of 160 kbps, along with the performance of the receiver for an undesired pulsed signal having a 50-μs pulse width and a PRF of 1000 pps. These results indicate that, at typical operating levels, the performance of conventional narrowband AM and FM voice communication systems would be approximately the same for a given amount of undesired power within the IF filter resulting from an FH/DS signal with a short dwell time or a pulsed signal, such that the dwell time and mean frequency occurrence rate of the FH/DS signal are, respectively, equal to the pulse width and pulse repetition frequency of the pulsed signal.

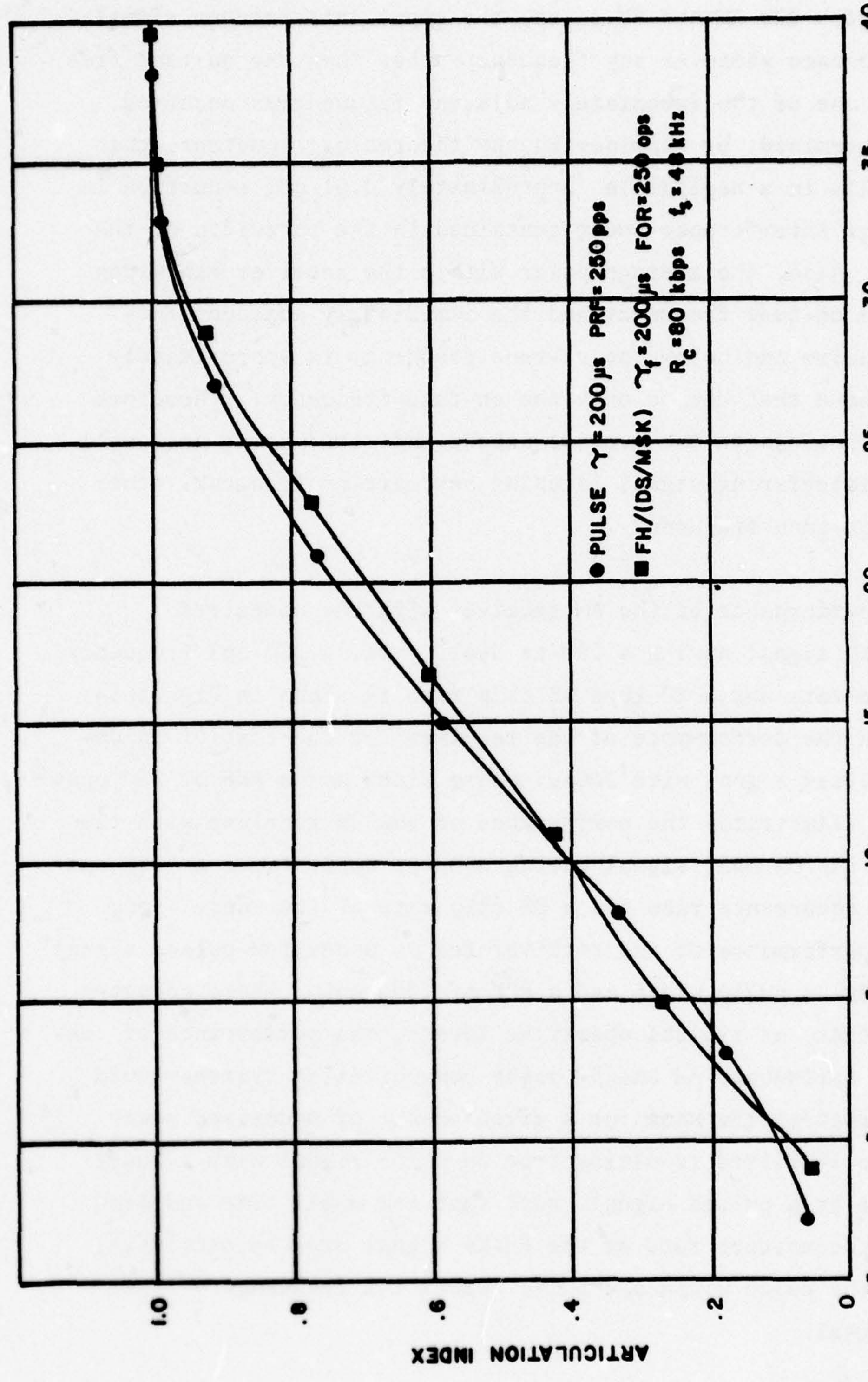


Figure 14. Simulated performance degradation curves for the AM receiver with FH/(DS/MSK) and pulsed interference.

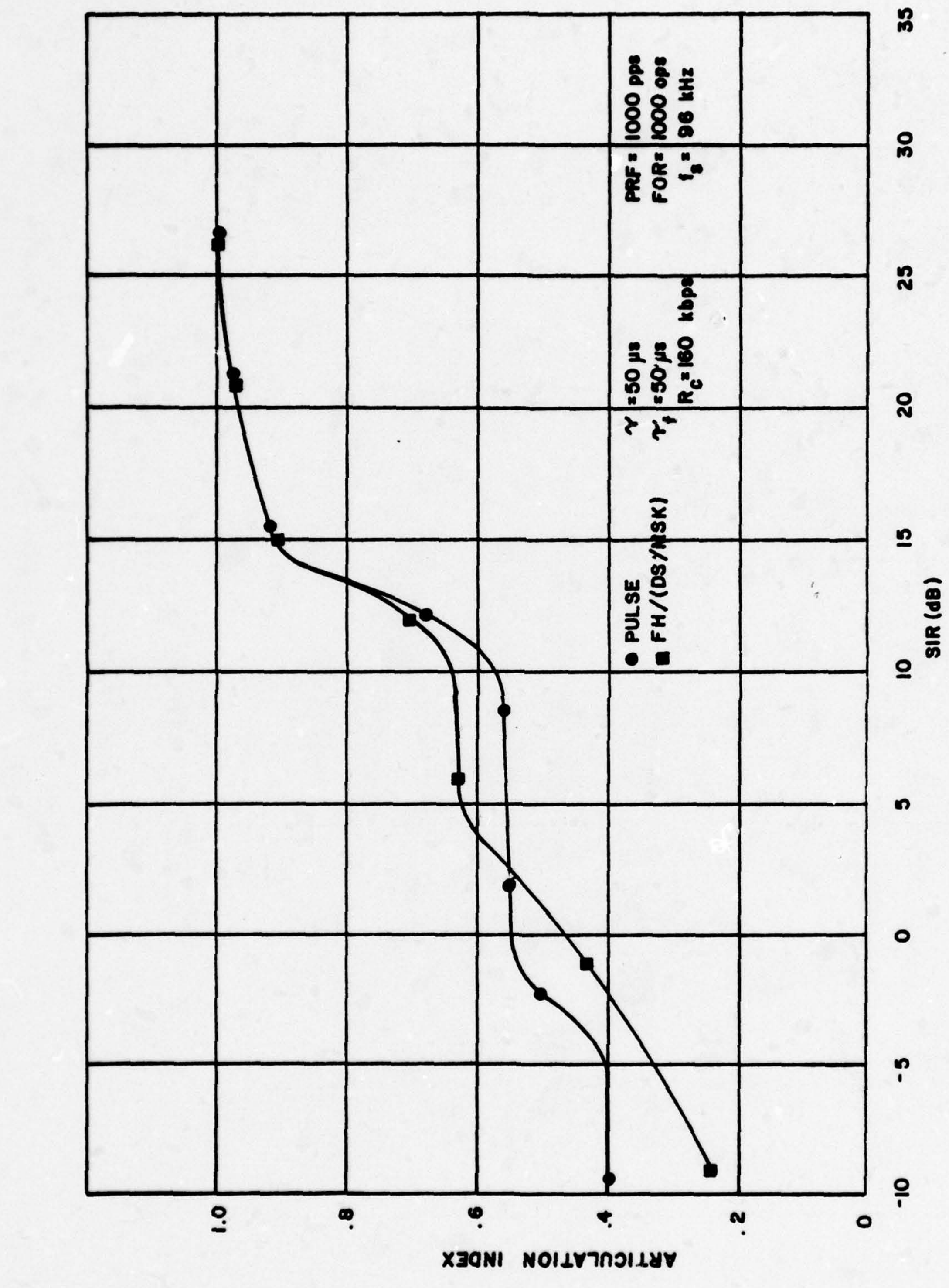


Figure 15. Simulated performance degradation curves for the FM receiver with FH/(DS/MSK) and pulsed interference.

SECTION 4

CONCLUSIONS

The simulation results obtained indicate the following, for conventional narrowband AM and FM voice communications systems at typical operating performance levels and a given amount of undesired power produced within the IF filter:

1. A direct-sequence interfering signal would result in approximately the same system performance as white Gaussian noise of the same bandwidth.

2. A frequency-hopping or short-dwell-time, frequency-hopping/direct-sequence interfering signal having only one of its frequencies hop within the IF filter would result in approximately the same system performance as a periodic pulsed signal in which the dwell time of the hopper is equivalent to the pulse width of the periodic signal and the hopper's mean frequency occurrence rate is equivalent to the PRF of the pulsed signal.

APPENDIX A

RECEIVER PERFORMANCE EVALUATION

INTRODUCTION

If one considers a system communicating information as including the listener or device for using the information, an evaluation of system performance can be made in terms of how well this user is able to discern the information or signal being delivered by the system. Noise will be introduced during the communicating process and will appear superimposed on the desired signal. The signal may be distorted during processing. Unwanted information, called interference, may also be communicated to the user and he must be able to distinguish this from the desired signal. One means of system performance evaluation is to observe the system in operation and measure the percentage of correctly communicated information. Another is to model the essential features of the system and measure the performance of the model. This appendix will consider both techniques but should not be considered an exhaustive study of the performance evaluation. A general discussion of this topic is presented by Hawthorne, et. al.[9]

The mathematical evaluation of a system's performance is the end objective of a prediction analysis. However, there is no one complete mathematical model for analyzing all types of systems. Although considerable research has been conducted on performance evaluation, the basic difficulty, which remains, is to determine what type of evaluation should be applied to systems in the presence of interference. In particular, for voice systems, Articulation Score (AS) (the percent of words correctly received) is the main intelligibility

[9]Hawthorne, G., et. al., *Performance of Communications Systems in the Presence of Interference,* Final Report - Volume 1, RADC-TR-59-133A.

standard. Articulation Index (AI) values closely approximate Articulation Scores in many applications. AI is much simpler and cheaper to apply and can be easily automated. In many cases, a simple power ratio is sufficient to evaluate system performance.

ARTICULATION SCORE

The basic measure of the intelligibility of a voice system is in terms of the percent of words correctly identified over a channel perturbed by interference. This measure has been designated as Articulation Score (AS) and is usually conducted with specific types of words or syllables, as well as specific system parameters. In an attempt to define the main voice parameters that are involved, workers in the field have conducted experiments by varying the word content, bandwidth, audio SNR and the type of talkers and listeners that are involved. As one would expect, the scores increased with increasing bandwidth, number of syllables in the words, speaker-listener familiarity, and audio SNR.

If the receiving system is subjected to a range of distortion or masking conditions, the AS may then be determined as a function of the interfering condition. Figure A-1 presents typical AS curves for different PB (phonetically balanced) word groupings in which the interference was white noise of various bandwidths. White noise, which is characterized by a continuous uniform spectrum, is one of the most effective maskers of speech and is often used in speech intelligibility studies as a standard or reference interference.

The articulation testing procedure is not simple nor has it always been standardized. Because it deals with the performance of human beings, the test can yield variable results in individual cases when proper statistical safeguards are not taken. It is generally necessary to use a number of listeners in order to obtain statistically meaningful results. Proper conduct of the test is

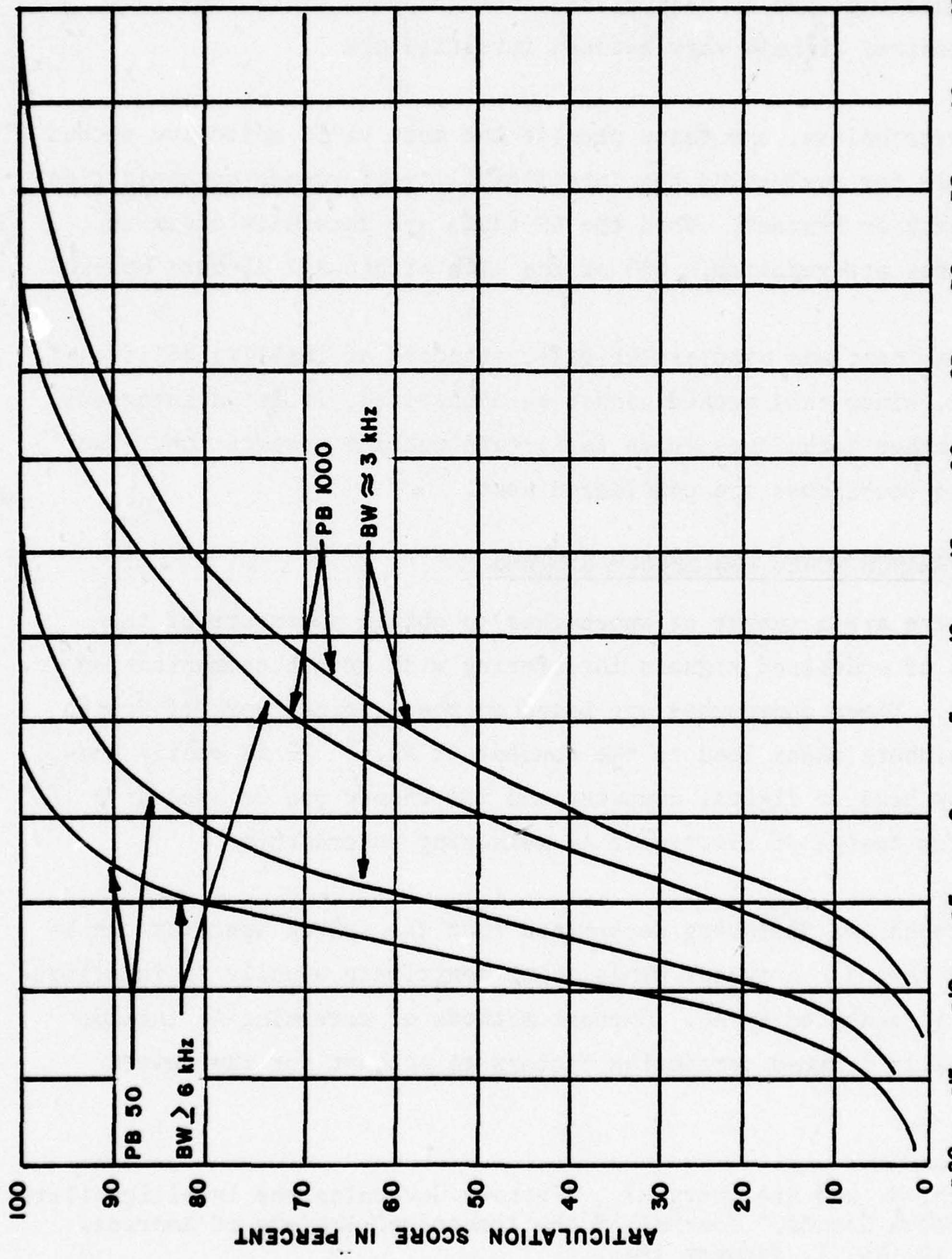


Figure A-1. Articulation score versus noise interference.

tedious and time-consuming. The situation is aggravated for most word lists by the necessity of a training program for the listeners in order to eliminate the improvement noted for successive tests. The test procedures, the material that is used, as well as the techniques employed to measure the average power of the desired and undesired signals vary between investigators.

Nevertheless, the tests provide the most valid objective method available for evaluating the intelligibility of speech communication components or systems. When the AS tests are carefully organized, the scores are repeatable 68% of the time within a 2 dB data spread.

This test was used as the basic standard of intelligibility. However, since this method cannot be mechanized, it is advantageous to use other techniques which facilitate machine computation. Two of these techniques are considered next.

A PERFORMANCE MODEL FOR SPEECH SYSTEMS

There are a number of approaches to obtain a measure of the effects of undesired signals interfering with speech communication systems. These approaches are based on the original work of French and Steinberg which lead to the concept of AI.[10] AI is easily computed by hand or digital computer and the theory can be used as a basis for design of electronic AI measuring instruments.

French and Steinberg determined that the speech spectrum can be divided into N contiguous bands which contribute equally to intelligibility as measured by AS. Current methods of determing AI include empirically derived correction factors to account for the upward

[10] French, N. and Steinberg, J., "Factors Governing the Intelligibility of Speech Sounds," *Journal of the Acoustical Society of America*, Vol. 19, No. 1, January 1947.

spread of masking. This is the phenomenon in which interference at a low frequency masks a higher frequency portion of the voice spectrum. Effects of noise and other factors (interference, distortion) prevent the N bands from making their full contribution to intelligibility. Since intensity of speech and interference may vary depending on the band, a weighting factor is determined for each band which reflects the fact that some bands do not make their maximum possible contribution to the speech intelligibility. The weighting factors vary for each band according to the ratio of the speech energy in the band to the hearing threshold. When the speech energy level in the band is 30 dB or more above the threshold level, it contributes its maximum value and, hence, has a unit weighting factor. When the speech energy level is between 0 and 30 dB above the threshold, the band's contribution is in proportion to its maximum as its level is to 30 dB. When the energy level is below the threshold, there is no contribution and the weighting factor vanishes. These weighting factors are additive and the sum can be used with empirical curves to determine the corresponding AS.

VOICE INTELLIGIBILITY ANALYSIS SET

An instrument produced by the General Electronics Laboratory (GEL) to mechanically measure voice intelligibility is called the Voice Intelligibility Analysis Set (VIAS).[11,12] This device divides the spectrum into a number of unequal bandwidths (14) and measures the contribution to intelligibility for each band. The contributions from the bands are then averaged over all 14 bands to produce the composite AI. The 14 VIAS frequency bands are shown in Figure A-2

[11] Thompson, A., *The Application of the Voice Interference Analysis System to the Prediction of Voice Intelligibility-Part 1*, Bell Report No. A70009-280, November 1967.

[12] Fitts, R., *Electronic Evaluation of Voice Communications Systems*, RADC-TDR-63-355, August 1963.

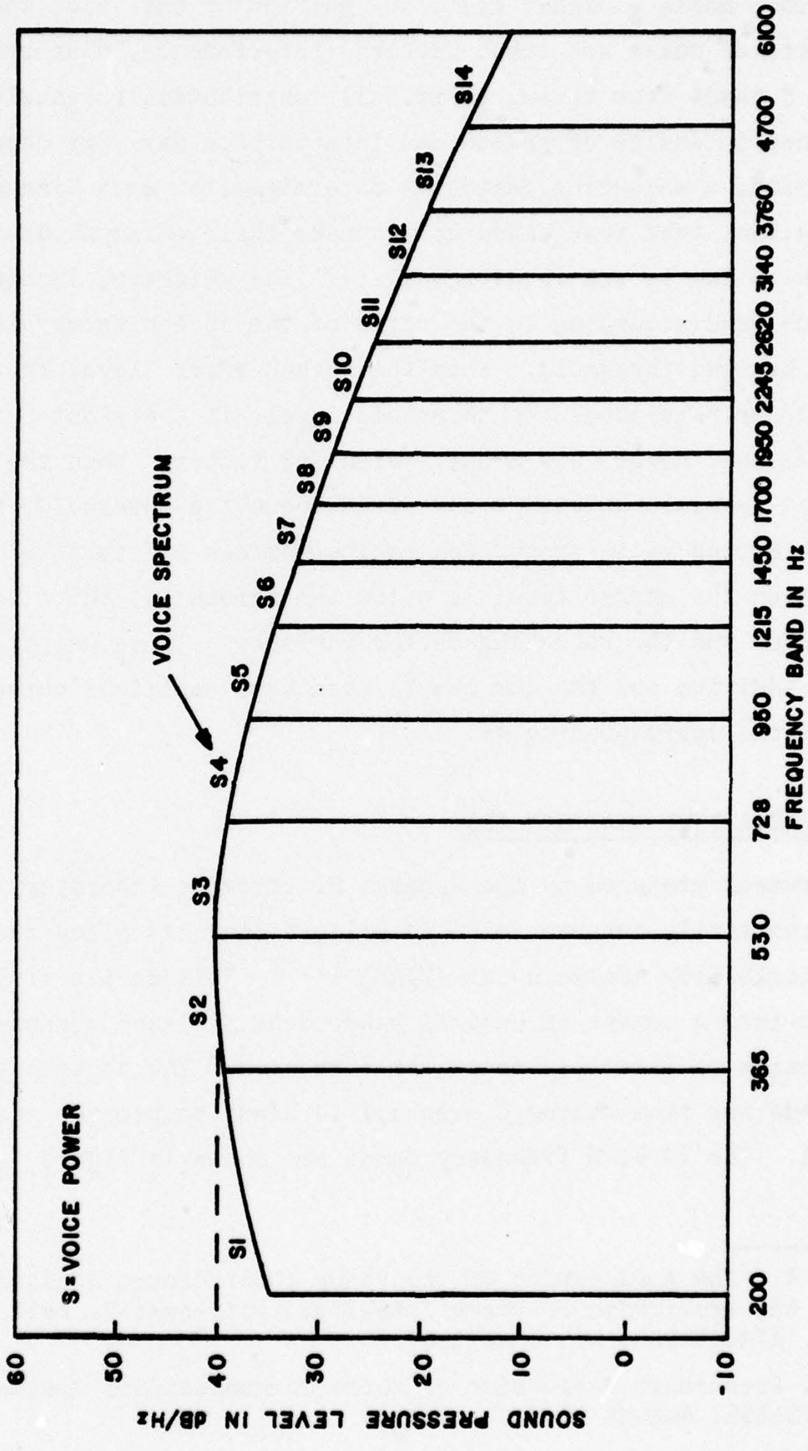


Figure A-2. Long-term speech spectrum and associated AI bands.

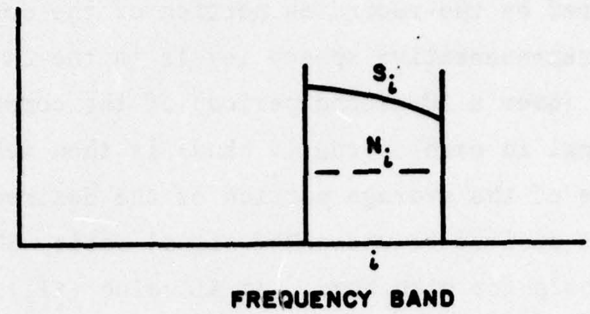
and the calculation of AI is depicted graphically in Figure A-3. A synthetic desired speech signal, which consists of a triangle-modulated 950-Hz tone, is transmitted over the test channel and is then measured by the recording portion of the device, in order to establish representative speech levels in the 14 bands. The average power (over a 10-second period) of the components of undesired signal in each of the 14 bands is then measured and, from knowledge of the average portion of the desired signal in each band, the desired-to-undesired signal ratio, SNR_i, is computed in decibels for each band. An AI value (AI_i) is assigned to each frequency band by the relation:

$$AI_i = \begin{cases} 1 & \text{for } SNR_i \geq 18 \\ (SNR_i + 12)/30 & \text{for } -12 < SNR_i < 18 \\ 0 & \text{for } SNR_i \leq -12 \end{cases} \qquad \text{(A-1)}$$

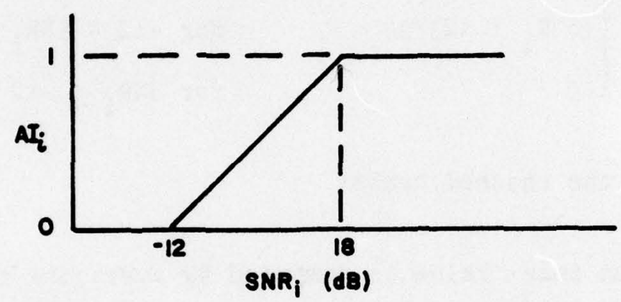
where

i = the channel number

The articulation index value is computed by averaging the AI_i values for each of the 14 bands. VIAS incorporates empirically derived correction factors to account for the upward spread of masking. This is the phenomenon in which interference at a low frequency masks a higher frequency portion of the voice spectrum. Correction must also be inserted for the receiver's frequency characteristics. The degree of frequency correction must be determined by measurement on the system and is then inserted manually. Reference 11, Appendix II, discusses this correction factor in detail. This method implies that interfering effects are independent and, consequently, additive. This last statement is especially critical since the use of performance numbers measured by this technique requires validation for situations in which the undesired signal is not additive.



$$SNR_i = 10 \, LOG_{10} \frac{S_i}{N_i}$$



$$AI = \frac{\sum_{i=1}^{14} (AI)_i}{14}$$

Figure A-3. Theoretical calculation of AI given the power in each frequency band.

COMPUTERIZED MODEL

The design relationships for the VIAS can be applied to mathematical expressions of a system waveform to determine AI values analytically. This approach can be easily programmed for use with computerized system simulation models.

It will be assumed that the desired signal has been approximated by an audio frequency tone and that the output power due to this tone is known and given by P_s. In addition, the power spectrum of the system audio output less the desired tone power must be known.

The frequency range from 200 Hz to 6100 Hz is divided into 14 frequency bands defined in TABLE A-1.[13] Though of course the desired tone power, P_s, only appears in one frequency band, this power represents a speech spectrum. Therefore, P_s is divided into 14 components, S_i, which represent the desired-signal power in each of the 14 frequency bands. These components are adjusted to conform with the typical long-term speech spectrum given in Figure A-2 and their values are given for each frequency band in TABLE A-1. That is, the desired signal power in the ith frequency band is:

$$S_i = 10 \log_{10} (\alpha_i P_s) \qquad \text{(A-2)}$$

The interference power, I_i, in the ith frequency band is calculated from the audio power spectrum data. The first step is to calculate the power contained in each frequency band by summing the power of each spectral component within the band. Let P_i be the power in the ith frequency band. The effective interference in the ith band may be greater than P_i because power in lower frequency bands will also interfere with reception of signals in this band. This effect is called

[13] *The Interference Prediction Model Theory and Computer Programs,* Bell Aerosystems Company, Report No. 60009-435, Revision B, Vols. I and II, February 1965.

TABLE A-1

AI FREQUENCY BANDS AND SIGNAL POWER PROPORTIONS

Band, i	Frequency (Hz)	Arithmetic Center, F (Hz)	Proportion of Desired Signal in Each Band, α	Upward Spread of Masking Falloff, K(dB/decade)
1	200 to 365	283	0.10915	50
2	365 to 530	448	0.20715	50
3	530 to 728	629	0.23315	50
4	728 to 950	839	0.15715	200/3
5	950 to 1225	1083	0.08715	200/3
6	1225 to 1450	1333	0.06115	200/3
7	1450 to 1700	1575	0.04115	200/3
8	1700 to 1950	1825	0.03395	100
9	1950 to 2245	2098	0.02185	100
10	2245 to 2620	2433	0.01625	100
11	2620 to 3140	2880	0.01385	100
12	3140 to 3760	3450	0.00825	100
13	3760 to 4700	4230	0.00625	100
14	4700 to 6100	5400	0.00355	100

upward spread of masking. The effective interference can be calculated for the ith band by calculating the masking power in band i due to each lower band and then choosing the largest one of the masking and in-band powers. The interference, I_i, in band i is the maximum power of the following:

$$I_i = \max \left[P_1 + a_1, P_2 + a_2, \ldots, P_j + a_j, \ldots, P_i\right] \quad \text{(A-3)}$$

and

$$a_j = -3 - K_j \log_{10} (F_i/F_j), \text{ dB} \quad \text{(A-4)}$$

The "a_j" factor indicates that the effect of power P_j in the jth band on the ith band is 3 decibels less than P_j and further reduced by K decibels per decade of frequency. The F and K factors for each frequency band are given in TABLE A-1.

Once the signal power, S_i, and the interference power, I_i, are determined for each frequency band, the articulation index value for each band can be calculated by:

$$AI_i = \frac{S_i - I_i}{30} + 0.4 \quad \text{(A-5)}$$

where

S_i and I_i = the powers having the same logarithmic units.

The characteristics of the system being evaluated may indicate that certain frequency bands should not be allowed to contribute to the overall AI. This arises when the audio spectrum is bandpass-filtered above 200 and below 6100 Hz. Since the desired-signal spectrum is artificially generated from 200 to 6100 Hz, the highest and lowest frequencies are actually attenuated and make no contribution to AI. The easiest solution is to disregard certain AI_i contributions by using a contribution factor C_i (in VIAS, the switch settings). The total AI value is then calculated by:

$$AI = \frac{\sum_{i=1}^{14} C_i(AI_i)}{14} \qquad \text{(A-6)}$$

where

C_i = depends on the audio filter characteristics.

The above parameters, while being those used in the design of the VIAS, may not be optimum. The object is, however, to evaluate system performance in the same way as VIAS so that the measured relationships between VIAS-measured AI and AS can be used to obtain AS values from the mathematical model AI values. More sophisticated methods of evaluating system performance could be devised for use with digital computer simulation models.

APPENDIX B

DIGITAL FILTERS

The voltage transfer function of an nth order, normalized (i.e., unity 3 dB cutoff frequency, $\omega_c = 1$), analog lowpass Butterworth filter is given by:

$$G(s') = \frac{1}{B_n(s')} \qquad \text{(B-1)}$$

where $B_n(s')$ is the nth order Butterworth polynominal.[14] The transfer function of the lowpass filter having a cutoff frequency of $\omega_c \neq 1$ is obtained by applying the lowpass-to-lowpass transformation, $s' \rightarrow s/\omega_c$, to the transfer function of the normalized filter. For an IF filter of bandwidth B, $\omega_c = \pi B$. The resulting transfer function may be expressed as:

$$G(s) = \frac{\omega_c^{\,n}}{\prod_{i=1}^{n} (s - s_i)} \qquad \text{(B-2)}$$

where s_i is the ith pole of the transfer function. The transfer function of the bandpass filter having lower and upper 3 dB cutoff frequencies of ω_1 and ω_2, respectively, is obtained by applying the lowpass-to-bandpass transformation $s' \rightarrow (s^2 + \omega_1 \ \omega_2)/[(\omega_2 - \omega_1)s]$ to the transfer function of the normalized filter. The resulting transfer function can be written as:

[14]Weinberg, L., *Network Analysis and Synthesis*, McGraw-Hill, New York, 1962.

$$G(s) = \frac{(\omega_2 - \omega_1)^n s^n}{\prod_{i=1}^{2n} (s - s_i)} \qquad \text{(B-3)}$$

Since G(s) for either the lowpass case in Equation B-2 or the bandpass case in Equation B-3 is a proper fraction, it can be expanded into a sum of partial fractions of the form:

$$G(s) = \sum_{i=1}^{p} \frac{r_i}{s - s_i} \qquad \text{(B-4)}$$

where $p = n$ or $p = 2n$ for the lowpass or bandpass case, respectively, and r_i is the residue evaluated at the ith pole and is given by:

$$r_i = (s - s_i)\, G(s) \Big|_{s = s_i} \qquad \text{(B-5)}$$

The digital transfer function, G(z), is obtained by applying the standard z- transformation (see Reference 8) to G(s) in Equation B-4 as follows:

$$G(s) = \sum_{i=1}^{p} \frac{r_i}{s - s_i} \rightarrow \sum_{i=1}^{p} \frac{r_i T}{1 - \exp(s_i T) z^{-1}} = G(z) \qquad \text{(B-6)}$$

where z^{-1} is the unit delay operator and T is the sampling interval. By combining pairs of terms which result from complex conjugate poles, G(z) can be written in the form:

$$G(z) = \sum_{i=1}^{m} \frac{a_{i,0} + a_{i,1}z^{-1}}{b_{i,0} + b_{i,1}z^{-1} + b_{i,2}z^{-2}} \qquad \text{(B-7)}$$

where m is the integer part of (n + 1)/2 for the lowpass case, m = n for the bandpass case, and the $a_{i,k}$, $b_{i,k}$ are real coefficients. It is convenient to divide the $a_{i,k}$ and $b_{i,k}$ by $b_{i,0} \neq 0$, which leads to the final form of the transfer function:

$$G(z) = \sum_{i=1}^{m} G_i(z) = \sum_{i=1}^{m} \frac{A_{i,0} + A_{i,1}z^{-1}}{1 + B_{i,1}z^{-1} + B_{i,2}z^{-2}} \qquad \text{(B-8)}$$

where $G_i(z)$ is termed the ith subfilter and the form of G(z) in Equation B-8 is called the parallel form and is illustrated in Figure B-1.

The input-output relationship of ith parallel digital subfilter having $G_i(z)$ as its transfer function is given by:

$$y_i(kT) = \sum_{j=0}^{1} A_{i,j}\, x_i(kT - jT) - \sum_{j=1}^{2} B_{i,j}\, y_i(kT - jT) \qquad \text{(B-9)}$$

In Equation B-9, $x_i(kT - jT)$ and $y_i(kT - jT)$ denote, respectively, the jth previous input and output sample values of the ith subfilter. The 0th previous values are the present values. Since the same input sample is applied to each parallel subfilter, the subscript i could be dropped in $x_i(kT - jT)$. Using the recursive relation in Equation B-9, the present output sample value of each subfilter can be computed. The present output sample value of the filter, y(kT), is then calculated as:

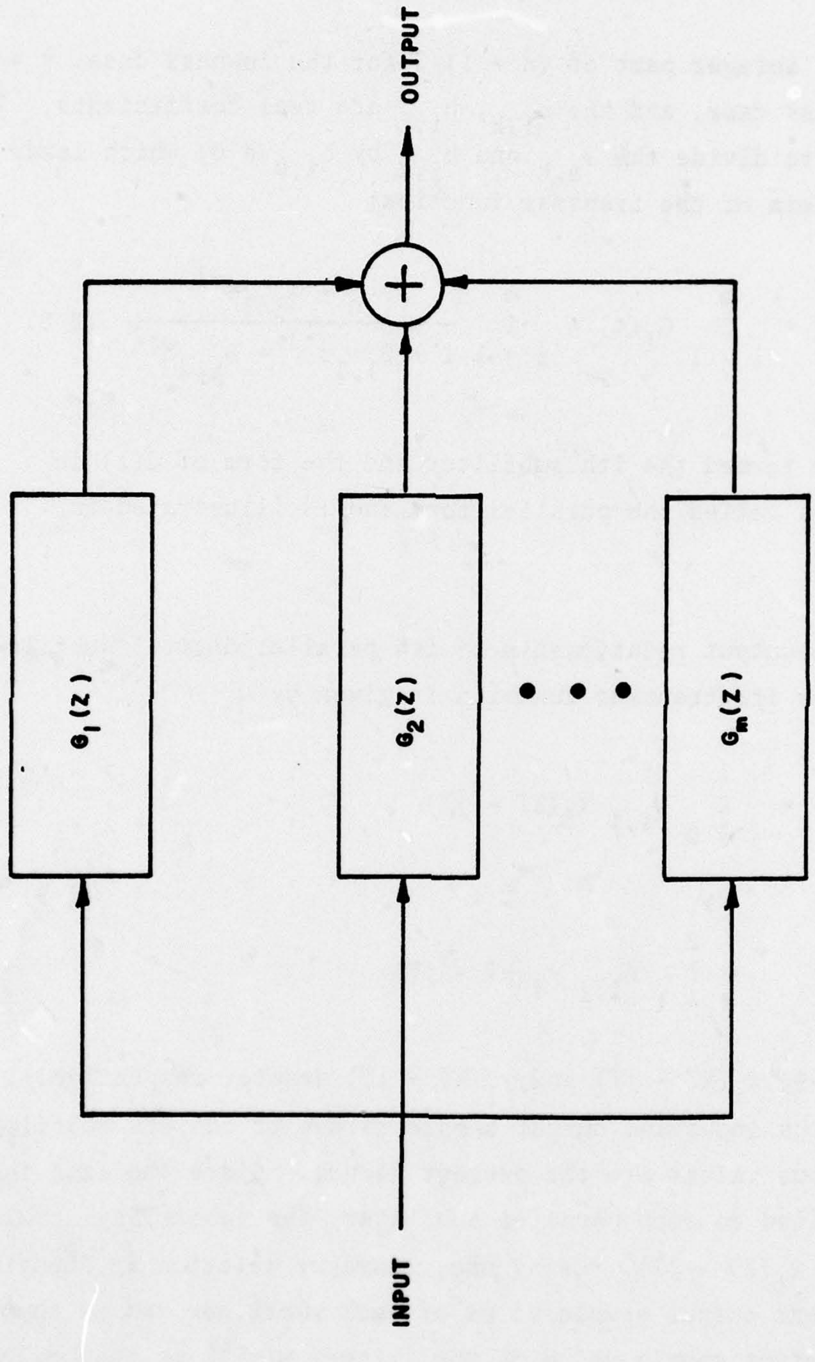


Figure B-1. A digital filter in parallel form.

$$y(kT) = \sum_{i=1}^{m} y_i(kT) \qquad \text{(B-10)}$$

The implementation of the digital filter in parallel form is shown in Figure B-2.

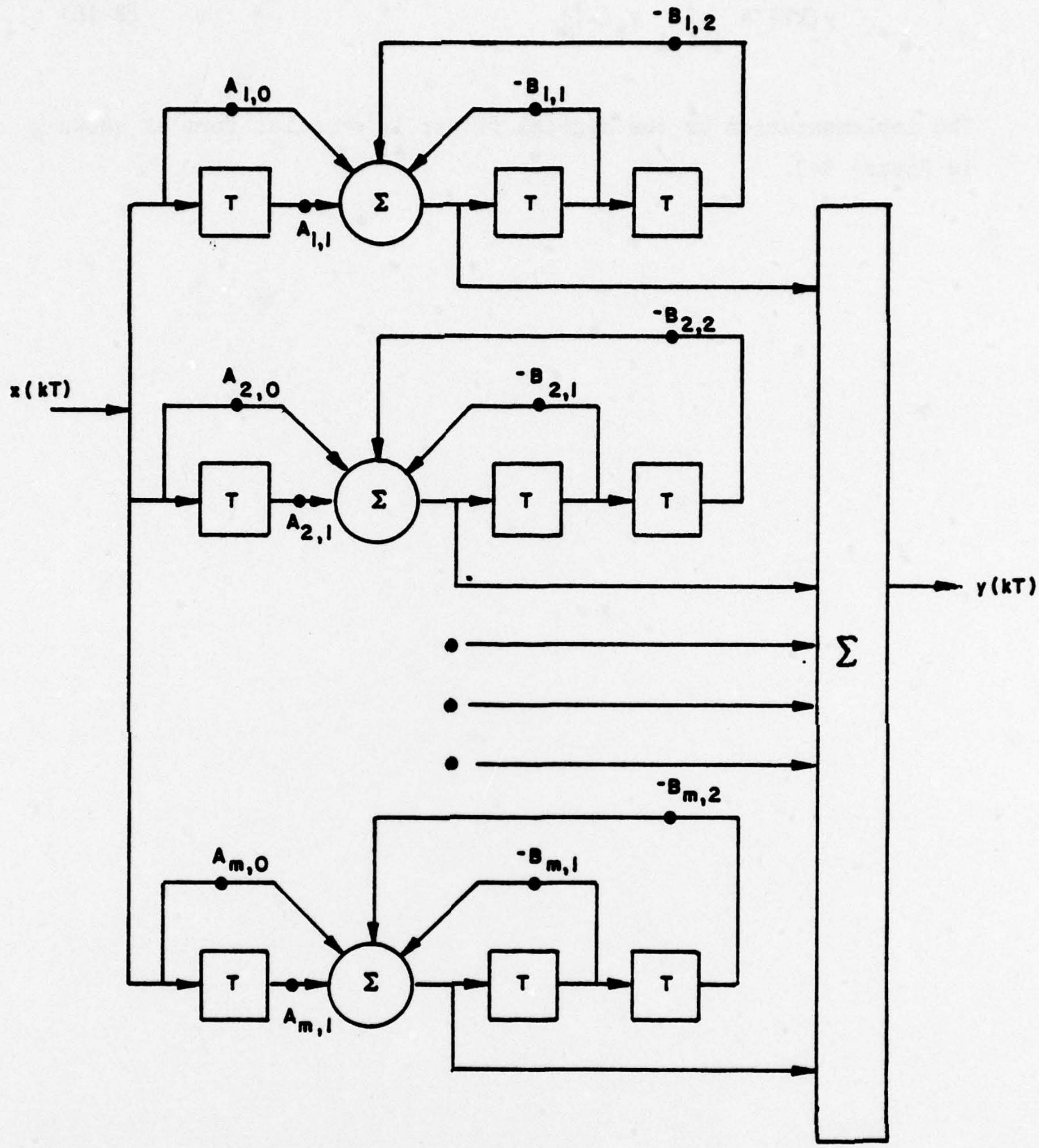


Figure B-2. Digital filter implementation.

REFERENCES

1. Farber, L., *Some Criteria For Frequency Sharing Between Voice Communication and Spread Spectrum Systems,* ECAC-CR-77-020, ECAC, Annapolis, MD, March 1977.

2. Meyers, R., *A Receiver Waveform Simulation Model,* ESD-TR-71-099, ECAC, Annapolis, MD, August 1971.

3. Dixon, R. C., *Spread Spectrum Systems,* John Wiley and Sons, New York, 1976.

4. Kravitz, F., Lemke, M., *Communication/Electronics Receiver Performance Degradation Handbook* (Second Edition), ESD-TR-75-013, ECAC, Annapolis, MD, August 1975.

5. Kivett, J. A., *Wideband Command and Control Modem (Waveform and Modem Conceptual Design Study),* RADC-TR-73-12, December 1972, Appendix A.

6. Hull, T. E., and Dobell, A. R., "Random Number Generators," *SIAM Review,* Vol. 4, No. 3, July 1962.

7. *A Handbook on Electrical Filters,* White Electromagnetics, Inc., Rockville, MD, 1963.

8. Kuo, F. F. and Kaiser, J. F., Eds., *System Analysis by Digital Computer,* John Wiley and Sons, New York, 1966, Chapter 7.

9. Hawthorne, G., et. al., *Performance of Communications Systems in the Presence of Interference,* Final Report - Volume 1, RADC-TR-59-133A.

10. French, N. and Steinberg, J., "Factors Governing the Intelligibility of Speech Sounds," *Journal of the Acoustical Society of America,* Vol. 19, No. 1, January 1947.

11. Thompson, A., *The Application of the Voice Interference Analysis System to the Prediction of Voice Intelligibility-Part 1,* Bell Report No. A70009-280, November 1967.

12. Fitts, R., *Electronic Evaluation of Voice Communications Systems,* RADC-TDR-63-355, August 1963.

13. *The Interference Prediction Model Theory and Computer Programs,* Bell Aerosystems Company, Report No. 60009-435, Revision B, Vols. I and II, February 1965.

14. Weinberg L., *Network Analysis and Synthesis,* McGraw-Hill, New York, 1962.

DISTRIBUTION LIST FOR PERFORMANCE OF VOICE COMMUNICATIONS SYSTEMS IN THE PRESENCE OF SPREAD SPECTRUM INTERFERENCE ESD-TR-77-005

DoD AND OTHERS	No. of Copies
DDC Cameron Station Alexandria, VA 22314	12
Office of Telecommunications Attn: R. Mayher 179 Admiral Cochrane Drive Annapolis, MD 21401	1
Director Defense Communications Agency Attn: Code 404A Washington, DC 20305	1
Director National Security Agency Attn: W36/Mr. V. McConnell Ft. George G. Meade, MD 20755	2
Department of Transportation Federal Aviation Administration Attn: ARD-62 2100 Second St., S.W. Washington, DC 20591	1
National Aeronautics & Space Administration Goddard Space Flight Center Attn: Code 801/Jim Scott Greenbelt, MD 20771	1
Commander in Chief U.S. European Command Attn: ECJ6-PF APO, NY 09128	1
Commander in Chief, Pacific Attn: J611 FPO, San Francisco 96610	1
DCA EUR (Code E-400) APO, NY 09131	1
Commander in Chief, Atlantic Attn: N612B Norfolk, VA 23511	1
Commander in Chief U.S. Readiness Command Attn: RCJ6-T MacDill AFB, FL 33608	1
Commander in Chief Strategic Air Command Offutt AFB, NE 68113	1

DoD AND OTHERS (Continued)	No. of Copies
Commander in Chief Continental Air Defense Command Ent AFB Colorado Springs, CO 80912	1
Capt. T. True, USAF Assistant U.S. Member ARFA NATO/OTAN APO, NY 09667	1
Dr. Robert Kahn DARPA 1400 Wilson Blvd. Arlington, VA 22209	1

NAVY	
Chief of Naval Operations (OP-941F) Navy Department Washington, DC 20350	1
Chief of Naval Operations (OP-941E) Navy Department Washington, DC 20350	1
Director Navy Electromagnetic Spectrum Center Naval Communication Unit Washington Washington, DC 20390	2
Commander (ELEX-008E) Naval Electronic Systems Command Washington, DC 20360	2
Commander (ELEX-034/Sweet) Naval Electronic Systems Command Washington, DC 20360	1
Commander (ELEX-095) Naval Electronic Systems Command Washington, DC 20360	1
Commander (ELEX-51023) Naval Electronic Systems Command Washington, DC 20360	1
Commander (ELEX-51024/Neill) Naval Electronic Systems Command Washington, DC 20360	1
Commander (AIR-360) Naval Air Systems Command Washington, DC 20361	1
Commander Naval Air Development Center Attn: T. Long (Code IJ) Warminster, PA 18974	1

DISTRIBUTION LIST (Continued)

NAVY (Continued)	No. of Copies
Commander Naval Ocean Systems Center Attn: I. Olsen San Diego, CA 92152	1
AIR FORCE	
Hq. ESD/DRI Hanscom AFB, MA 01731	7
Hq. ESD/DRT Hanscom AFB, MA 01731	1
Hq. USAF/FMO Washington, DC 20330	1
Hq. USAF/RDPE Washington, DC 20330	1
Hq. AFSC/SDDE Andrews AFB Washington, DC 20334	1
Hq. AFSC/DOKP Andrews AFB Washington, DC 20334	1
Hq. RADC/RBC Griffiss AFB, NY 13441	1
Hq. RADC/RBCT Griffiss AFB, NY 13441	1
SAMSO/AWST PO Box 92960 Worldway Postal Center Los Angeles, CA 90009	1
Hq. AFCS/DOYF Richards-Gebaur AFB, MO 64030	1
Hq. AFEWC/SUR San Antonio, TX 78243	1
Hq. ADTC/DC Eglin AFB, FL 32542	1
ARMY	
HQDA (DAMO-TCF) (A. Anderson) Washington, DC 20310	1
HQDA (DAMA-CS) Washington, DC 20310	1
CDR, USAECOM Attn: DRSEL-NL-RY-5 Fort Monmouth, NJ 07703	1
CDR, USAECOM Attn: DRSEL-CE-ES Fort Monmouth, NJ 07703	1

ARMY (Continued)	No. of Copies
CDR, USAMIRADCOM Attn: DRDMI-TE (D. Smith) Redstone Arsenal, AL 35807	1
CDR, USATECOM Attn: DRSTE-EL (L. Doughty) Aberdeen Proving Ground, MD 21005	1
USACC CH, C-E Services Division Attn: CC-OPS-CE (Col. Falbo) Room 1B269 Washington, DC 20301	1
CDR, USACC Attn: CC-OPS Fort Huachuca, AZ 85613	1
CDR, USACEEIA Attn: CCC-EMEO-ECD (Mr. B. Allen) Fort Huachuca, AZ 85613	1
CDR, USACEEIA Attn: CCC-EMEO Fort Huachuca, AZ 85613	1
Office of the Area Frequency Coordinator Attn: DoD AFC Arizona (Mr. Harry Berger) Fort Huachuca, AZ 85613	1
DoD Area Frequency Coordinator Attn: Mr. H. Hungate WSMR, NM 88002	1
Commander U.S. Army Element, DoD ECAC Annapolis, MD 21402	1
CDR, U.S. Army Signal School Attn: ATZHDC Fort Gordon, GA 30905	1
CDR, U.S. Army Signal School Attn: ATSN-CTD (W. Mannel) Fort Gordon, GA 30905	1
CDR, USAINSCOM Arlington Hall Station Attn: IATEL-ZS (A. Casey) Arlington, VA 22212	1
CDR, USAINSCOM Arlington Hall Station Attn: IARD-MA (Dr. F. Williams) Arlington, VA 22212	1
CDR, USAINSCOM Arlington Hall Station Attn: IARD-MA (P. Ambrose) Arlington, VA 22212	1

DISTRIBUTION LIST (Continued)

ARMY (Continued)	No. of Copies
CDR, USATRADOC Attn: ATCE Fort Monroe, VA 23651	1
USCINCEUR RAFLO BONN Box 365 US Embassy APO, NY 09080	1
CDR US Forces Korea Attn: J-6 Freq. Coordinator APO, SF 96301	1
CDR 5th Signal Command Attn: CCE-PO-TS APO, NY 09056	1
CDR 5th Signal Command Attn: CCE-PO-F APO, NY 09056	1
Hq. Berlin Brigade Radio System Branch CE Division APO, NY 09205	1
CDR, 7th-Signal Command Attn: CCN-PO-OP Fort Ritchie, MD 21719	1
MARINE CORPS	
Director, Development Center Marine Corps Development and Education Command (Attn: C^3 Div) Quantico, VA 22134	5
Commandant of the Marine Corps Headquarters, U.S. Marine Corps (CCT) (Attn: Captain Franklin) Washington, DC 20380	1
Commander Naval Ocean Systems Center (Marine corps Liaison Office) San Diego, CA 92152	1
Commanding Officer Marine Corps Tactical Systems Support Activity Camp Pendleton, CA 92055	1
Hq. ESD/DCBE Attn: JTIDS JPO/Major McCalley Hanscom AFB, MA 01731	1
Office of the Project Manager, PLRS DPM, PLRS/TIDS (Lt. Col. W. H. Bond) Fort Monmouth, NJ 07703	1

INTERNAL	No. of Copies
ACL	1
ACV	1
ACW	1
ACY	1
ACZ	1
ACOE-E	1
ACOA-T	15
L. Farber	2
A. Hernandez	1
P. Newhouse	1
S. Cameron	1
M. Ellis	1
M. Lustgarten	1
M. Maiuzzo	1
D. Madison	1
M. Sanchez	1