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Modulation Studies for IGOSS

HIROSHI AKIMA

BOULDER, COLO.
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INSTITUTE FOR TELECOMMUNICATION SCIENCES
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FOREWORD

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MODULATION STUDIES FOR IGOSS

Hiroshi Akima

Modulation techniques for high frequency (HF) data transmission for the Integrated Global Ocean Station System (IGOSS) are discussed. Noncoherent frequency-shift-keying (NCFSK) and time-differential phase-shift-keying (TDPSK) systems are preliminarily selected as preferred systems. Performances of these preferred systems are discussed both in the presence of noise and/or interfering signal and under adverse propagation conditions. A serious drawback inherent in a TDPSK system under adverse propagation conditions is pointed out. Frequency tolerance, frequency stabilization techniques, and selectivity of receivers are also discussed.

Key Words: Data transmission, frequency stability, high frequency (HF), Integrated Global Ocean Station System (IGOSS), modulation technique, selectivity.

1. INTRODUCTION

This report describes certain aspects of modulation techniques and related topics to be considered for the Integrated Global Ocean Station System (IGOSS). The intention is not to recommend the use of a specific modem (modulator and demodulator), but to provide information necessary for selecting a modem and other related factors that may affect the design of the IGOSS telecommunication network.

Many system parameters of the IGOSS are still undefined; however, the long range plan for the IGOSS is for several platforms (buoys and ships) to transmit data simultaneously to a common shore station in a common 3-kHz high frequency (HF) band. This is similar to multichannel, frequency-division-multiplex (FDM) data modems in point-to-point HF communications. FDM modems, which can tolerate the multipath time-delay spreads associated with HF ionospheric radio propagation, have been studied for more than a decade, and the results were used as guidelines for this study.

Despite the similarity between the IGOSS and FDM point-to-point communications, we must also be aware of the difference between them. In a point-to-point system, signals in different channels are transmitted from a common transmitter, but in the IGOSS they are transmitted from different geographically dispersed platforms. An IGOSS shore station must receive simultaneously, in a common 3-kHz band, several signals that may have different median values and fade independently of each other. Therefore, interchannel interference is more severe than in point-to-point systems, and special attention must be given to the selectivity of the receiver and demodulator.

Independent channel transmitters also mean that synchronization between channels cannot be expected, so that demodulation techniques that rely on orthogonality between channel signals cannot be used.

Since service intervals of up to 1 year are desirable, high reliability and small power consumption of the buoy transmitter are among the criteria for selecting a specific modulation technique and modem.

2. PRELIMINARY SELECTION OF MODULATION SYSTEMS

Among the various modulation systems for digital data transmission, described here in the order of their development, noncoherent frequency-shift-keying (NCFSK) and time-differential phase-shift-keying (TDPSK) systems appear to be the most attractive candidates for the IGOSS.

2.1. On-Off Keying System

Digital information is transmitted by on-off keying of the unmodulated carrier located at the center of the channel. For transmitters limited in peak envelope power, the average power is half that of constant amplitude systems, such as frequency-shift-keying (FSK) or phase-shift-keying (PSK) systems. Since power is transmitted for only one of the two states, a decision threshold must be set artificially. This can be done easily if the signal wave is not fading and the noise is negligible compared with the signal; however, the peak envelope power of received radio wave signals propagated over HF ionospheric paths generally fluctuates considerably, so that setting the decision threshold is a difficult problem. For these reasons, on-off keying of a carrier is thought unsuitable for automated data transmission; therefore, this system is not considered further.

2.2. Noncoherent Frequency-Shift-Keying (NCFSK) System

A certain number of equally spaced frequencies $f_o, f_1, \ldots, f_{n-1}$ are chosen in each channel, where n is the number of chosen modulation conditions. The channel subcarrier is frequency modulated (frequency-shift-keyed) in accordance with the digital information signal, so that the frequency of the subcarrier is equal to f_i during an interval in which the information content is equal to i $(i=0, 1, 2, \ldots, n-1)$. Binary modems $(n=2)$ are most common. Quaternary modems $(n=4)$, sometimes called diplex or twinplex, are also used. NCFSK is one of the most widely used systems for digital data transmission over HF ionospheric paths. It is a strong candidate for the IGOSS.

2.3. Time-Differential Phase-Shift-Keying (TDPSK) System

In a PSK system, a subcarrier is phase modulated (phase-shift-keyed) by the digital information signal in such a way that the difference between the phase of the subcarrier φ and the reference phase φ_0 is equal to

$$\varphi - \varphi_0 = (2i + 1 - n)\,\pi/n, \qquad (i = 0,\ 1,\ 2, \ldots,\ n - 1),$$

where n is the number of modulation conditions. Both binary (n=2) and quaternary (n=4) modems are widely used. In a TDPSK system, both modulation and demodulation are referenced to the phase of the carrier during the previous signalling element. TDPSK is also widely used for digital data transmission over HF ionospheric paths. It is also a strong candidate for the IGOSS.

2.4. Frequency-Differential Phase-Shift-Keying (FDPSK) System

In the FDPSK system, both modulation and demodulation are referenced to the phase of another subcarrier (either modulated or unmodulated), located near the modulated subcarrier (de Haas, 1965). Although this technique is not widely used, perhaps it can yield the highest data signalling rate in a given bandwidth among various techniques of high-speed data transmission over HF ionospheric paths. It has a serious drawback, however, when applied to single-channel circuits. Since a reference-phase carrier must always accompany the modulated carrier, a linear power amplifier, instead of a class-C power amplifier, is required in the transmitter, resulting in a considerably lower transmitter efficiency. Therefore, this system is not considered further in this study.

2.5. Other Techniques and Modems

There are several other modems based on different techniques: the Kathryn modem (Zimmerman and Kirsch, 1965), the Adapticom modem (Di Toro et al., 1965), and one based on a swept-frequency (chirp) modulation technique (Dayton, 1968). These modems are not considered further here, because (a) a too high level of sophistication requires complicated equipment or (b) available operating data are insufficient to adequately predict actual performance characteristics.

3. EFFECTS OF NOISE AND INTERFERENCE ON THE SYSTEM PERFORMANCE

When the desired signal is not fading or experiencing slow, flat fading (i.e., the fading is slow compared with the modulation rate, and different spectrum components are fading simultaneously), the element error probability P_e in a binary NCFSK system is equal to one-half the probability that the instantaneous amplitude of the undesired signal U exceeds the instantaneous amplitude of the desired signal D (Montgomery, 1954), i.e.,

$$P_e = \frac{1}{2} \Pr\{ U > D \}$$

$$= \frac{1}{2} \iint_{U>D} p(D, U)\, dD\, dU, \qquad (1)$$

where $\Pr\{\ \}$ stands for "the probability that," and $p(D, U)$ is the probability density function of D and U. This applies to an NCFSK system where the modulation index is not less than unity and no low-pass filter is used before the decision-making circuit. For proper application of (1), the statistics of the undesired signal must be measured at the input of the demodulator, i. e., at the input to the limiter in a limiter-discriminator demodulator, and in a bandwidth equivalent to the sum of the bandwidths of the two filters in a dual-filter demodulator. Except for a few simple cases, such as a non-fading signal or a Rayleigh-fading signal in the presence of Gaussian noise, this integral cannot be evaluated analytically but must be computed numerically.

The element error probability in a binary TDPSK system is closely approximated by (1); it is exactly equal to the result computed by (1), when the undesired signal consists of Gaussian noise (Cahn, 1959). (Since the bandwidth for a TDPSK system is usually considered one-half that for an NCFSK system, its required ratio of signal energy per bit to noise power density can be 3 dB lower than that for NCFSK to give an equal error probability.)

Amplitudes of signals propagated by the ionosphere vary considerably with time. For short intervals (3 to 7 min), amplitude distribution functions close to the Rayleigh distribution predominate. Over longer intervals (30 to 60 min), on the other hand, amplitude distributions more often follow a log-normal law. Although the form of the measured distribution may differ from the Rayleigh distribution, the observed fading range, defined as the ratio of the upper and lower deciles, is the same order as 13.4 dB expected for the Rayleigh distribution (CCIR, 1967g).

In HF channels noise is mostly atmospheric. The amplitude-probability distribution (APD) of atmospheric noise can be represented, accurately enough for most applications, by an appropriate curve chosen from a family of idealized curves. The choice can be made by specifying a single parameter, defined as the ratio of the rms to the average of the envelope voltage and denoted by V_d in dB (CCIR, 1964).

For simplicity we assume that the interfering signal is either a continuous wave (CW) or an FSK or PSK signal.

The element error probability in each particular case is generally given as a function of signal-to-noise ratio (SNR) and/or signal-to-interference ratio (SIR). The SNR is defined as the ratio of desired signal power to average noise power, and SIR is the ratio of desired signal power to interfering signal power, where both the noise power and interfering signal power are measured at the demodulator input. For a fading signal, either desired or interfering, its median power is used in defining the SNR and SIR.

The integral in (1) has not been computed for very general cases, such as a log-normal-fading signal in the presence of atmospheric noise and a log-normal-fading interfering signal with dual diversity reception. Some particular cases, however, have been studied and described below.

3.1. Nonfading Signal and Atmospheric Noise Without Interference

The element error probability is one-half the value of the APD of the atmospheric noise corresponding to the signal amplitude. The relation between SNR and error probability, parametric in noise parameter V_d (Akima et al., 1969), is shown in figure 1.

3.2. Fading Signal and Atmospheric Noise Without Interference (Nondiversity Reception)

The relation between SNR and element error probability, parametric in noise parameter V_d, computed by numerical integration (Akima et al., 1969), are given in figure 2 for Rayleigh-fading signals and in figure 3 for log-normal-fading signals (fading range = 13.4 dB).

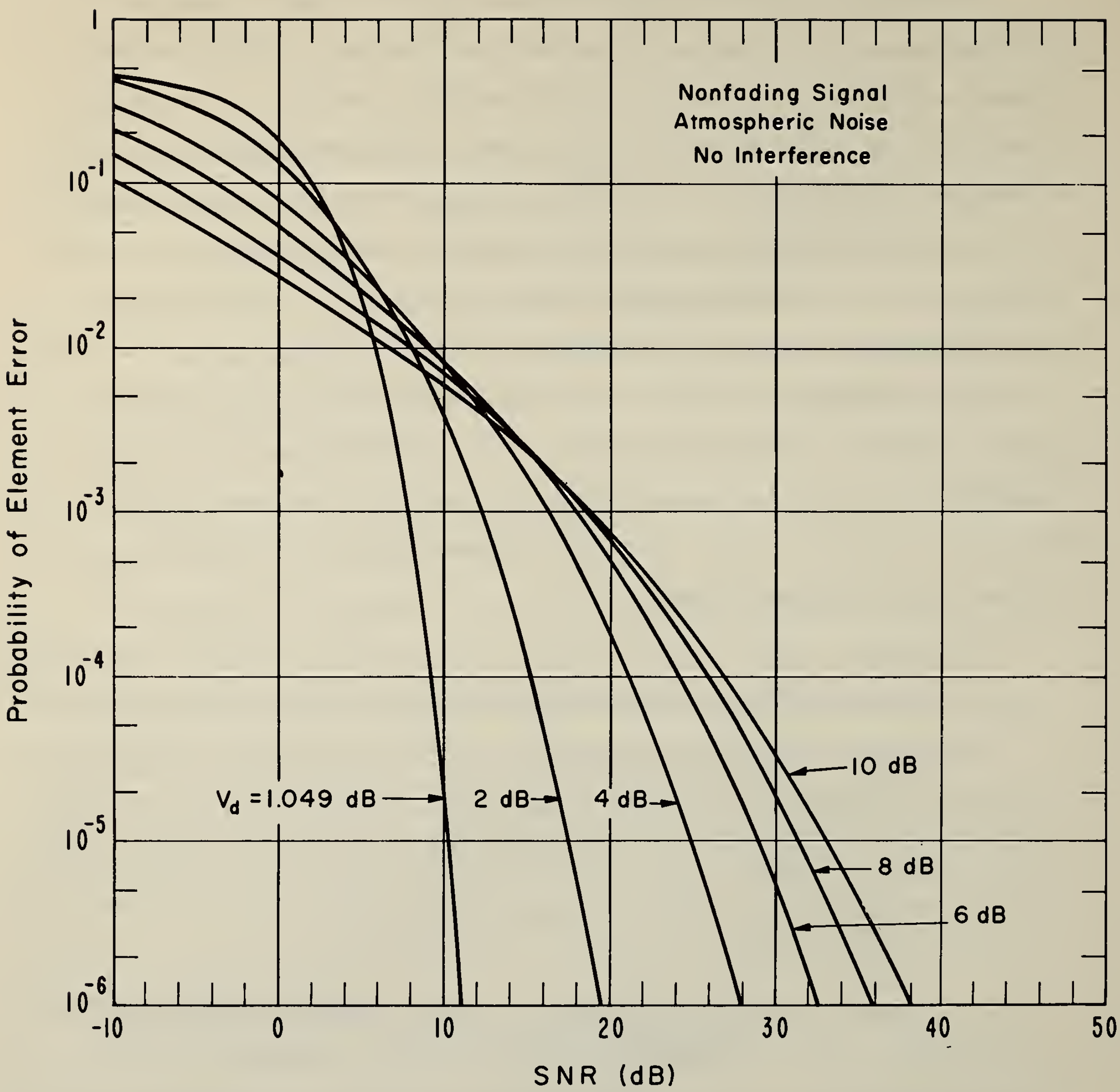

Figure 1. Element error probability in a single-channel NCFSK system under stable (nonfading) conditions with atmospheric noise and no interference.

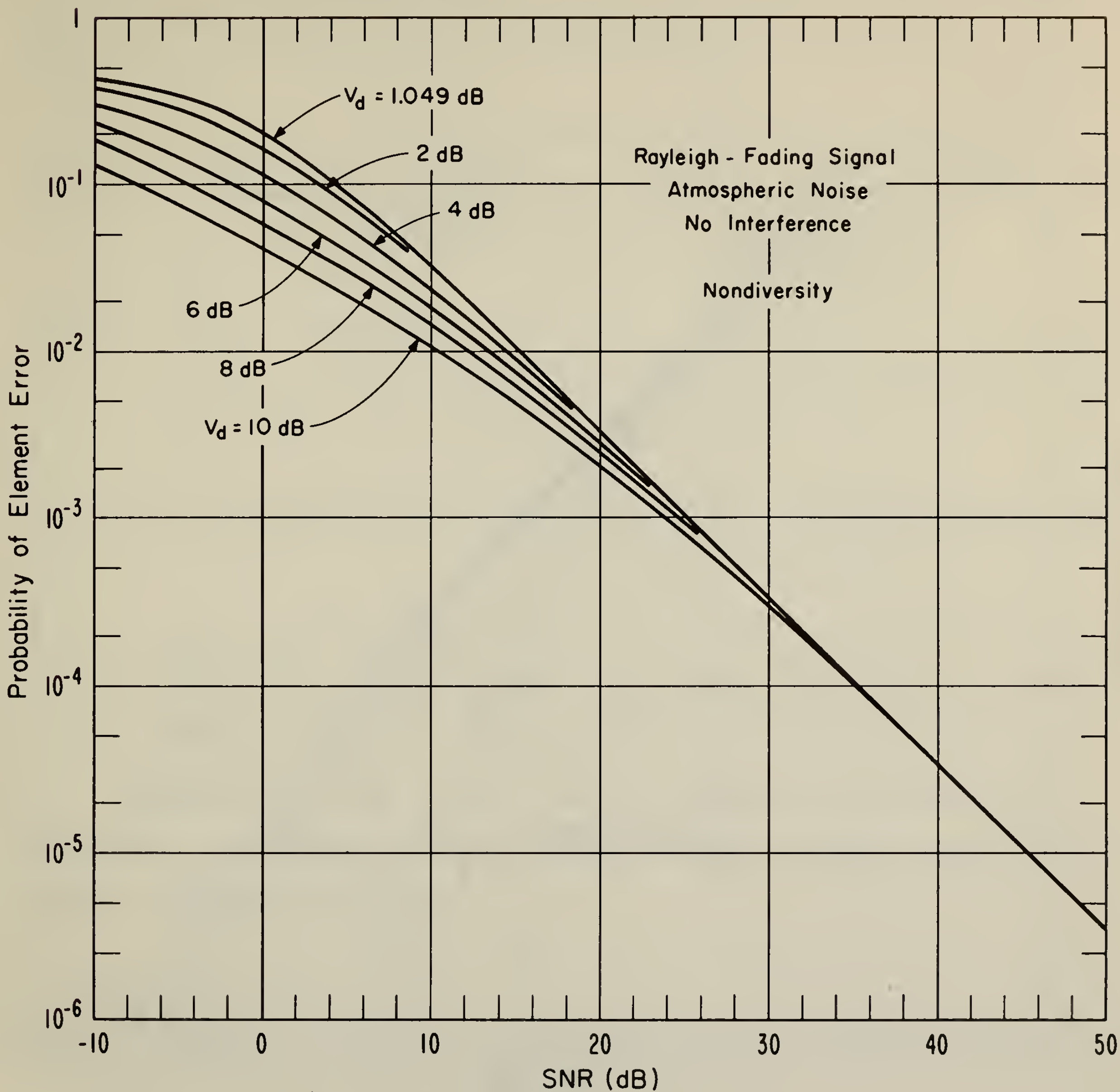

Figure 2. Element error probability in single-channel NCFSK system under Rayleigh-fading conditions with atmospheric noise and no interference, nondiversity reception.

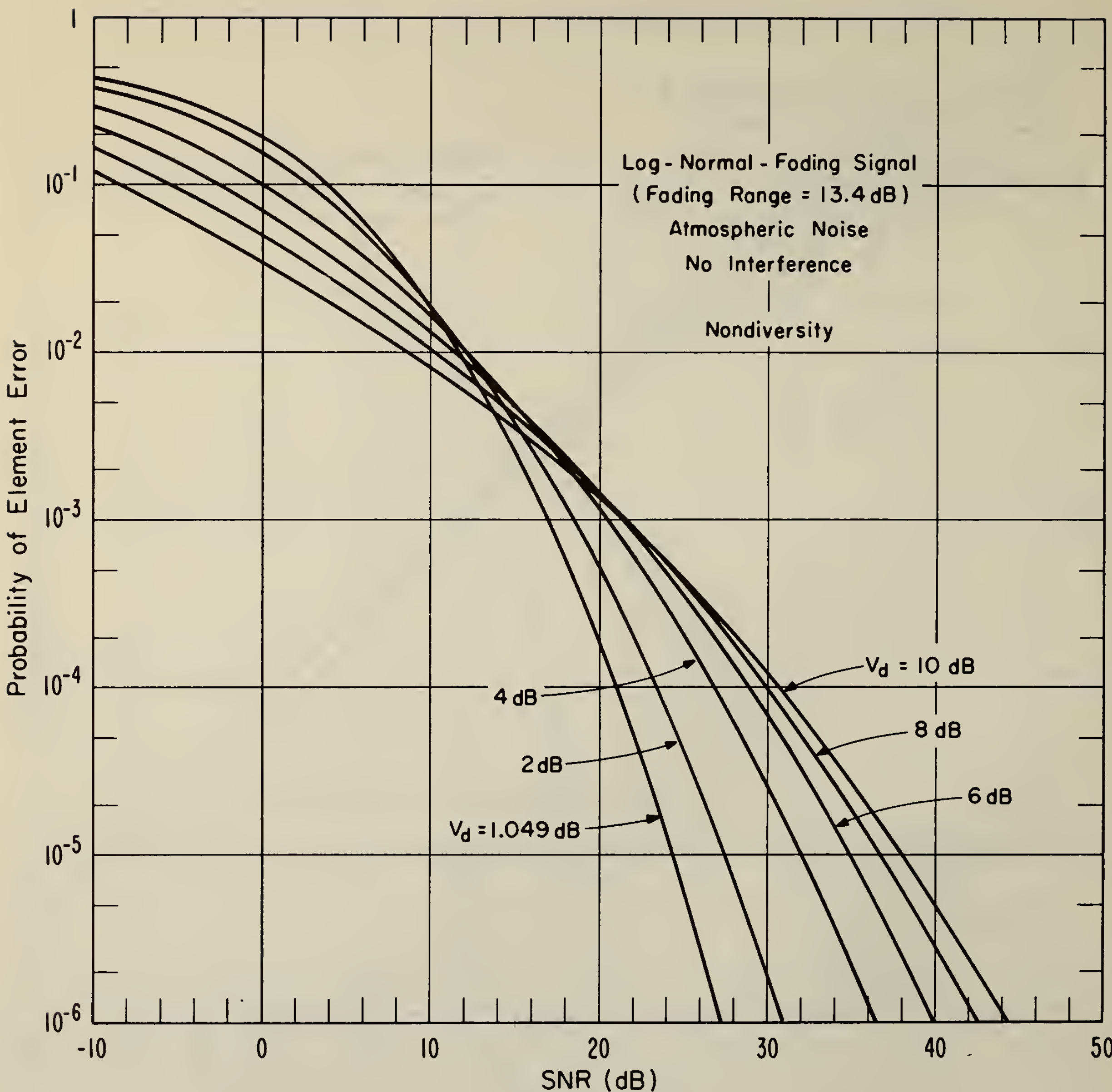

Figure 3. Element error probability in a single-channel NCFSK system under log-normal-fading conditions with atmospheric noise and no interference, nondiversity reception. (The fading range is assumed to be 13.4 dB, which is the same as expected for Rayleigh fading.)

3.3. Fading Signal and Atmospheric Noise Without Interference (Diversity Reception)

When dual diversity with a selection-switching combiner is used, the probability of element error in an NCFSK system is given by

$$P_e = \frac{1}{2} \Pr \left\{ U_1 > D_1 , \; V_1 > V_2 \right\}$$

$$+ \frac{1}{2} \Pr \left\{ U_2 > D_2 , \; V_2 > V_1 \right\} , \qquad (2)$$

where

D_i = voltage of desired signal from antenna i or in channel i (i = 1, 2),

U_i = voltage of undesired signal (noise) from antenna i or channel i (i = 1, 2),

V_i = vector sum of D_i and U_i
= voltage of signal plus noise from antenna i or in channel i (i = 1, 2).

Since D_2 and U_2 follow the same probability distributions as D_1 and U_1, respectively, the second term in (2) is equal to the first term. Therefore, we have

$$P_e = \Pr \left\{ U_1 > D_1 , \; V_1 > V_2 \right\} \qquad (3)$$

as an expression of element error probability in an NCFSK system with dual selection-switching diversity reception.

This result can easily be extended to cover higher order diversity with selection-switching combining. For example,

$$Pe = 2Pr \left\{ U_1 > D_1 \, , \; V_1 > V_2 \, , \; V_1 > V_3 \, , \; V_1 > V_4 \right\} \qquad (4)$$

is the expression of el ment error probability for quadruple diversity reception.

Evaluating P_e from (3) or (4) is the evaluation of a definite integral of a joint probability density function and can best be done by the Monte Carlo method or numerical simulation. One big advantage of the Monte Carlo method is that it does not become much more complicated when the order of diversity is increased. When we use this method, sequences of random numbers are chosen to follow the probability distribution functions of the signal, the noise, and the phase difference between the signal and the noise. The necessary condition for the occurrence of an error, as given by (3) or (4), is then tested with successive sets of elements; each element is picked from the corresponding sequence of random numbers, and the number of times an error occurs is counted.

We assumed statistical independence among the signal and noise amplitudes and a random phase relationship between signal and noise. We then tested one million sets of elements for each combination of two types of signal fading, dual and quadruple diversity, six values of V_d, and several values of SNR in 2-dB steps. Results of these computations for dual diversity are shown in figures 4 and 5 for Rayleigh fading signal and log-normal-fading signal, respectively, and for quadruple diversity in figures 6 and 7. Values of the required SNR for element error probabilities of 10^{-3} and 10^{-4} are read from figures 2 through 7 and shown in table 1. This table indicates that the gain in required SNR obtained either by the use of dual diversity instead of nondiversity or by the use of quadruple diversity instead of dual diversity depends on the type of fading, the value of V_d, and the allowable error

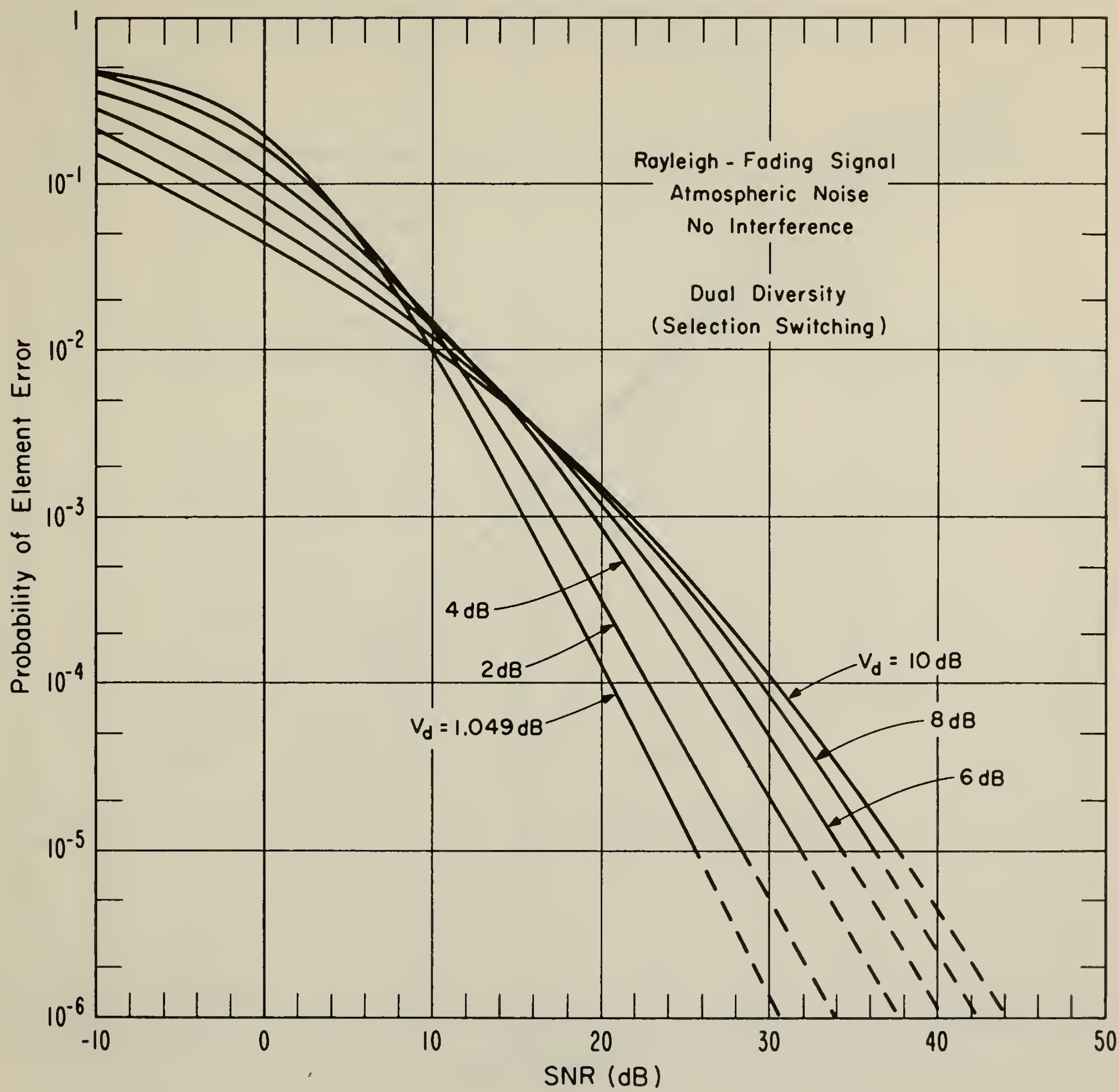

Figure 4. Element error probability in a single-channel NCFSK system under Rayleigh-fading conditions with atmospheric noise and no interference, dual (selection-switching) diversity reception.

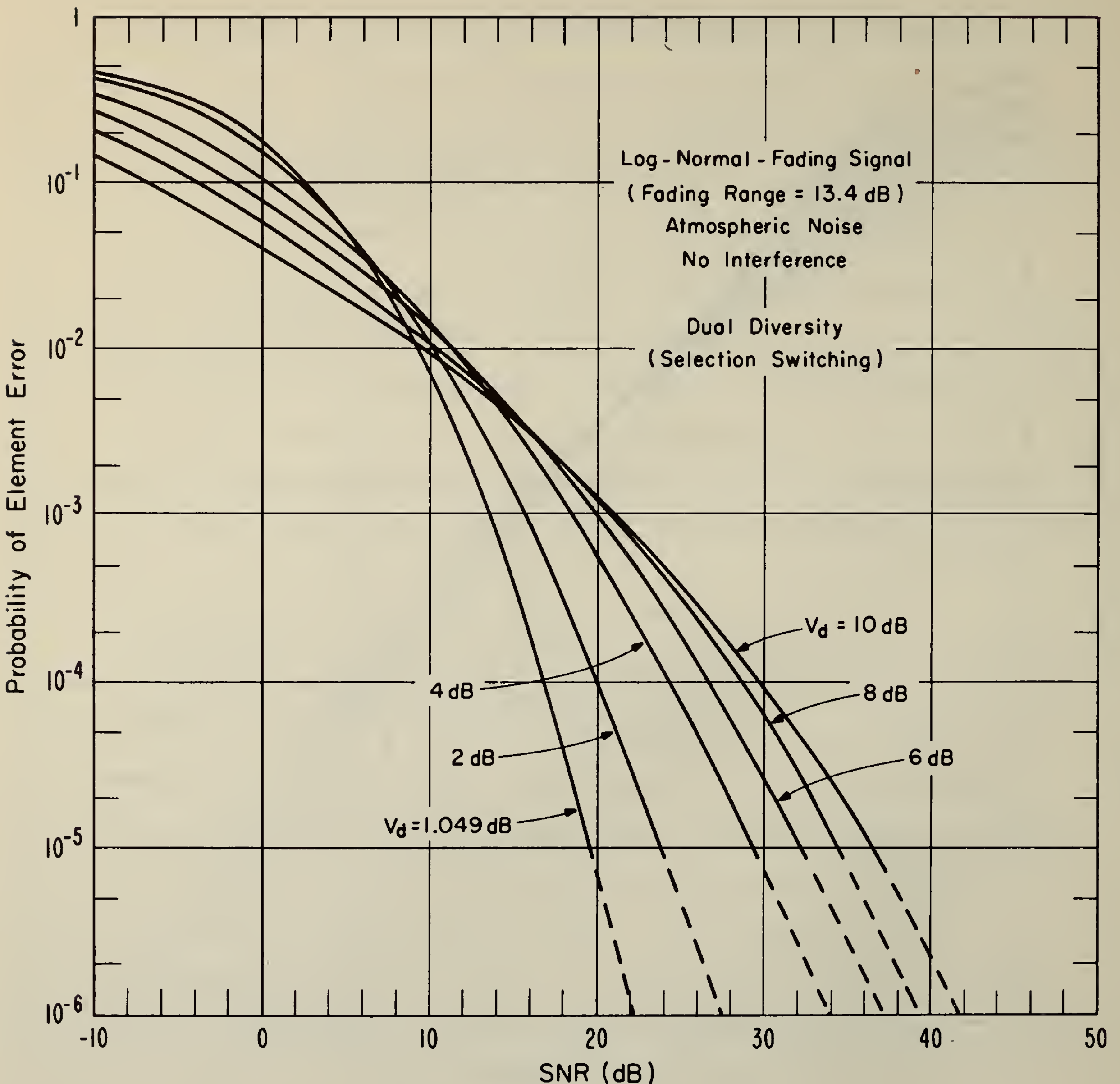

Figure 5. Element error probability in a single-channel NCFSK system under log-normal-fading conditons with atmospheric noise and no interference, dual (selection-switching) diversity reception. (The fading range is assumed to be 13.4 dB.)

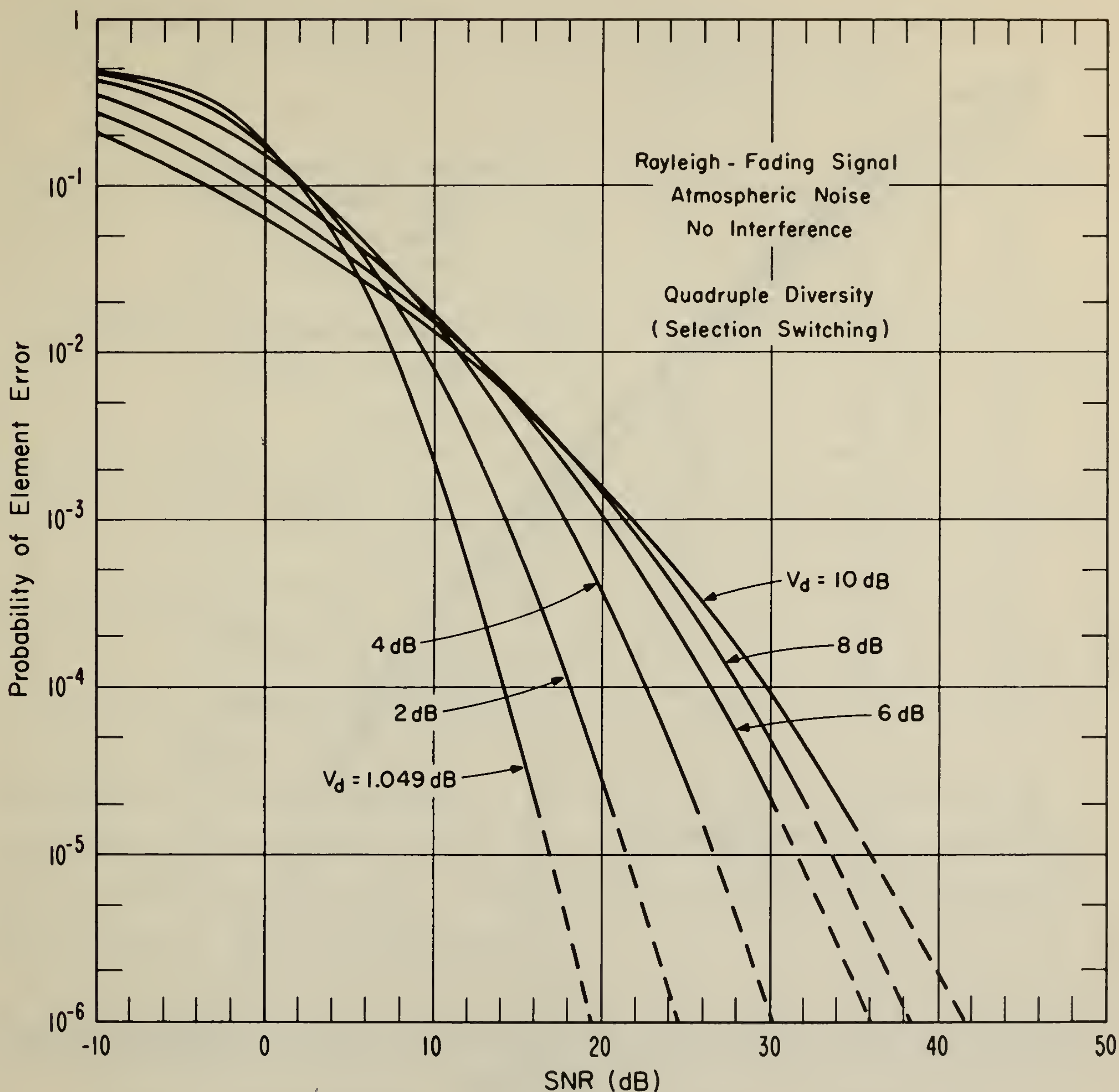

Figure 6. Element error probability in a single-channel NCFSK system under Rayleigh-fading conditions with atmospheric noise and no interference, quadruple (selection-switching) diversity reception.

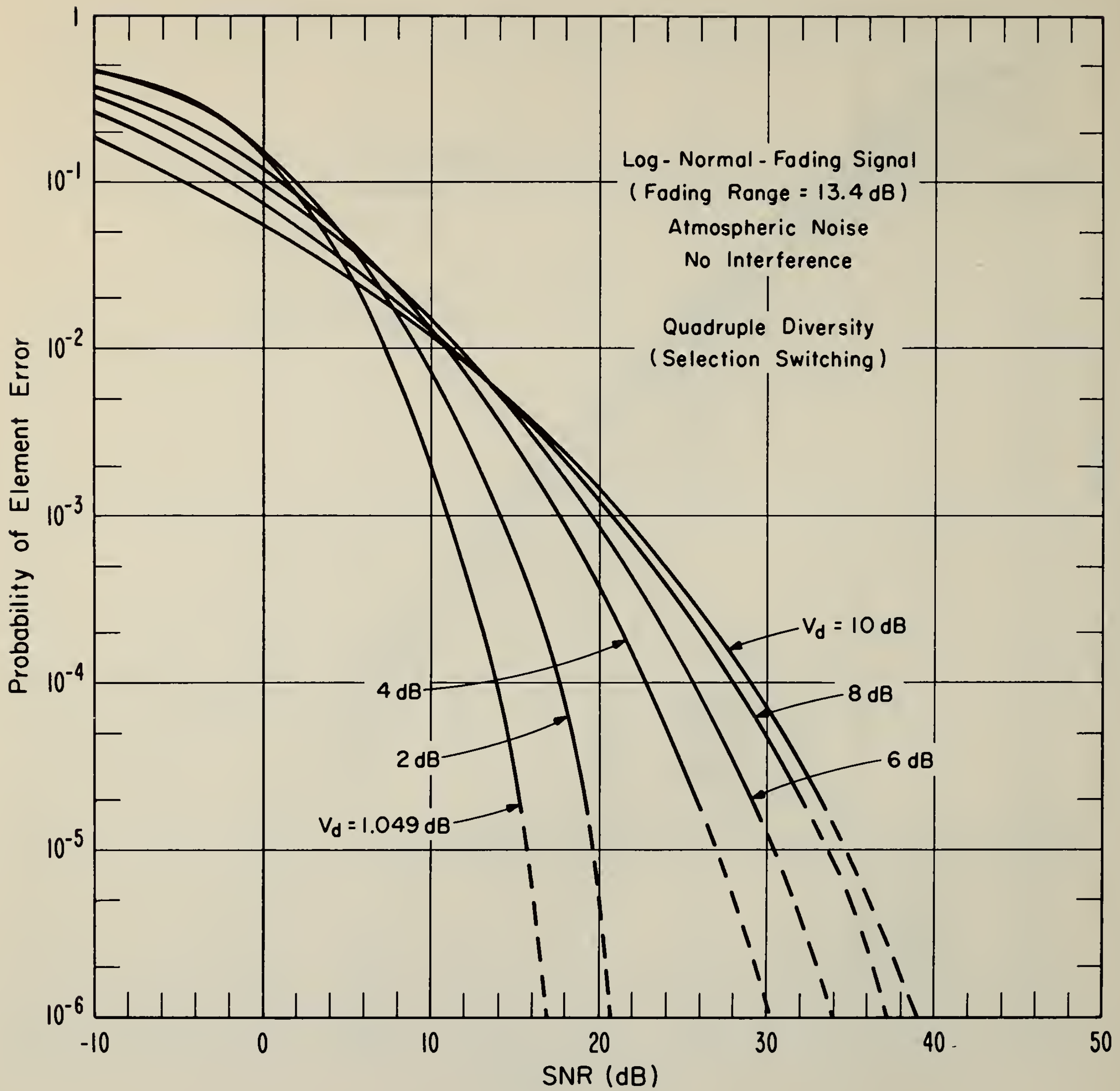

Figure 7. Element error probability in a single-channel NCFSK system under log-normal-fading conditions with atmospheric noise and no interference, quadruple (selection-switching) diversity reception. (The fading range is assumed to be 13.4 dB.)

Table 1. Required SNR for a Single-Channel NCFSK System Under Fading Conditions With Atmospheric Noise and No Interference. (P_e is the element error probability and V_d is the CCIR noise parameter in dB. Selection-switching diversity is assumed.)

Signal Fading	P_e	Diversity	Required SNR in dB					
			V_d = 1.049	2	4	6	8	10
Rayleigh	10^{-3}	None	25	25	25	25	24	24
		Dual	15	17	19	20	21	22
		Quadruple	11	14	18	20	21	22
	10^{-4}	None	35	35	35	35	35	35
		Dual	21	23	26	28	29	30
		Quadruple	14	18	23	26	28	30
Log-Normal	10^{-3}	None	17	18	20	21	21	21
		Dual	14	16	18	20	21	21
		Quadruple	11	14	18	19	21	21
	10^{-4}	None	21	23	27	29	30	31
		Dual	17	20	24	27	29	30
		Quadruple	14	18	23	26	28	29

probability. The gain is generally less for a log-normal-fading signal than for a Rayleigh-fading signal. The higher the V_d of the noise, the less the gain from diversity. The lower the allowable error probability, the greater the gain from diversity.

3.4. Nonfading Signal and Nonfading Interference With Gaussian Noise

Since the distribution of an undesired signal consisting of a nonfading interfering signal and Gaussian noise follows the Nakagami-Rice distribution, the error probability P_e can be expressed as

$$P_e = \frac{1}{2} Q(a, b), \tag{5}$$

where Q(a, b) is a Marcum's Q function (Marcum, 1950), a is the ratio of the amplitude of the interfering signal to the rms noise voltage, and b is the ratio of the amplitude of the desired signal to the rms noise voltage. Both a and b are measured at the demodulator input. The relation between SNR and error probability, parametric in SIR, is given in figure 8.

As expected, interference necessitates an increase in required SNR. At $P_e = 10^{-3}$, for example, SIR of 3 dB, 6 dB, and 10 dB necessitate increases of 9 dB, 5 dB, and 2 dB, respectively, in required SNR.

3.5. Nonfading Signal and Nonfading Interference With Atmospheric Noise

The element error probability is equal to one-half the value of the APD, for the level corresponding to the signal amplitude, of the composite waveform which consists of the interfering signal and atmospheric noise. The APD of the composite waveform for a specified level is equal to a two-dimensional integral of the probability density function of the composite waveform outside a circle whose center is the origin of the coordinate system and radius equals the specified level. Assuming randomness between the phases of the interfering signal and atmospheric noise, we can evaluate the integral by numerical integration. The relations between SNR and error probability, parametric in SIR, are given in figures 9, 10, and 11 for the V_d values of 4, 6, and 10 dB, respectively. The effect of interference is less severe in the presence of atmospheric noise than in Gaussian noise. At $P_e = 10^{-3}$ and $V_d = 4$ dB, SIRs of 3, 6, and 10 dB necessitate increases of 7, 3, and 1 dB, respectively, in required SNR,

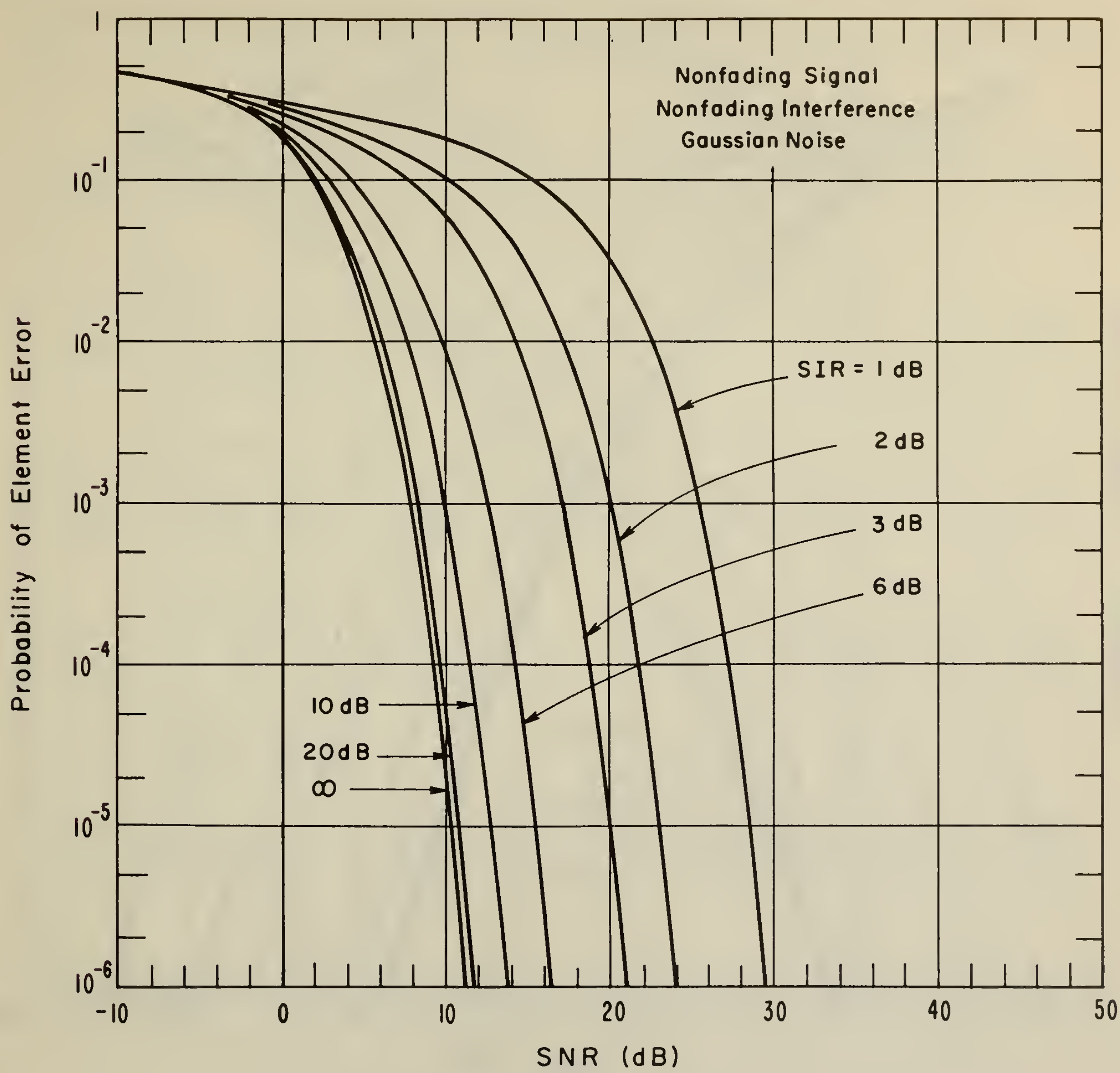

Figure 8. Element error probability in a single-channel NCFSK system under stable conditions with Gaussian noise and nonfading interference.

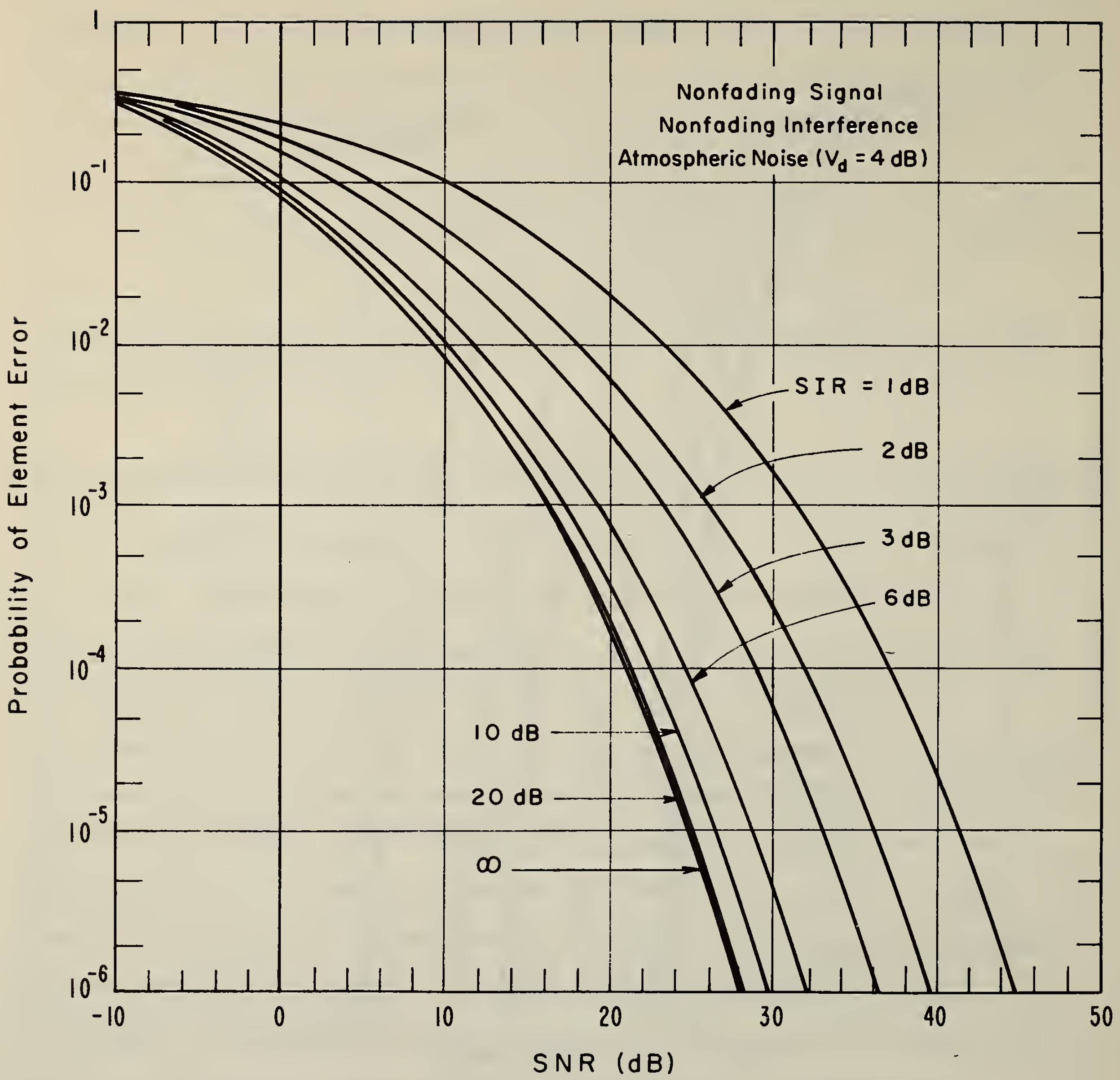

Figure 9. Element error probability in a single-channel NCFSK system under stable conditions with atmospheric noise (V_d = 4 dB) and nonfading interference.

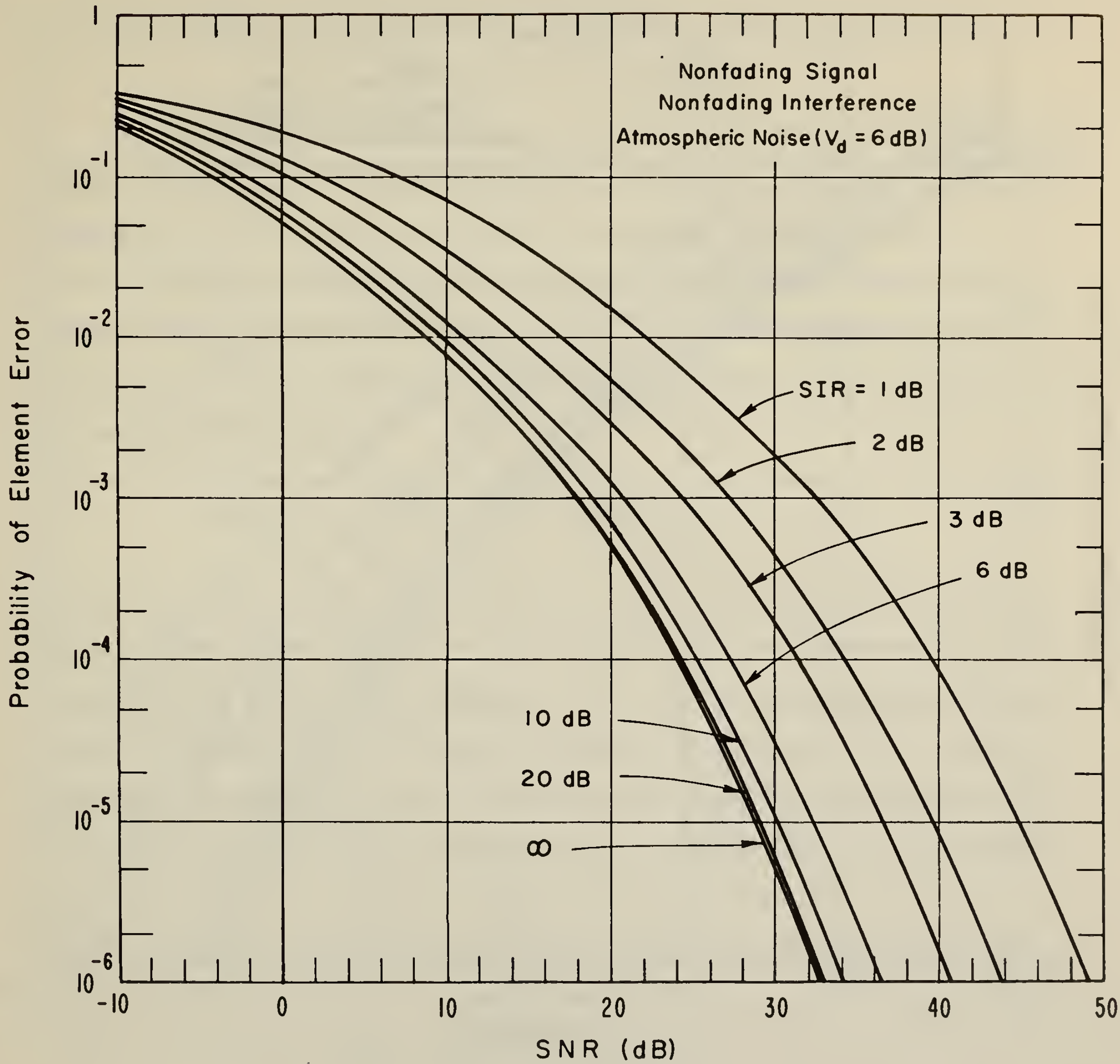

Figure 10. Element error probability in a single-channel NCFSK system under stable conditions with atmospheric noise (V_d = 6 dB) and nonfading interference.

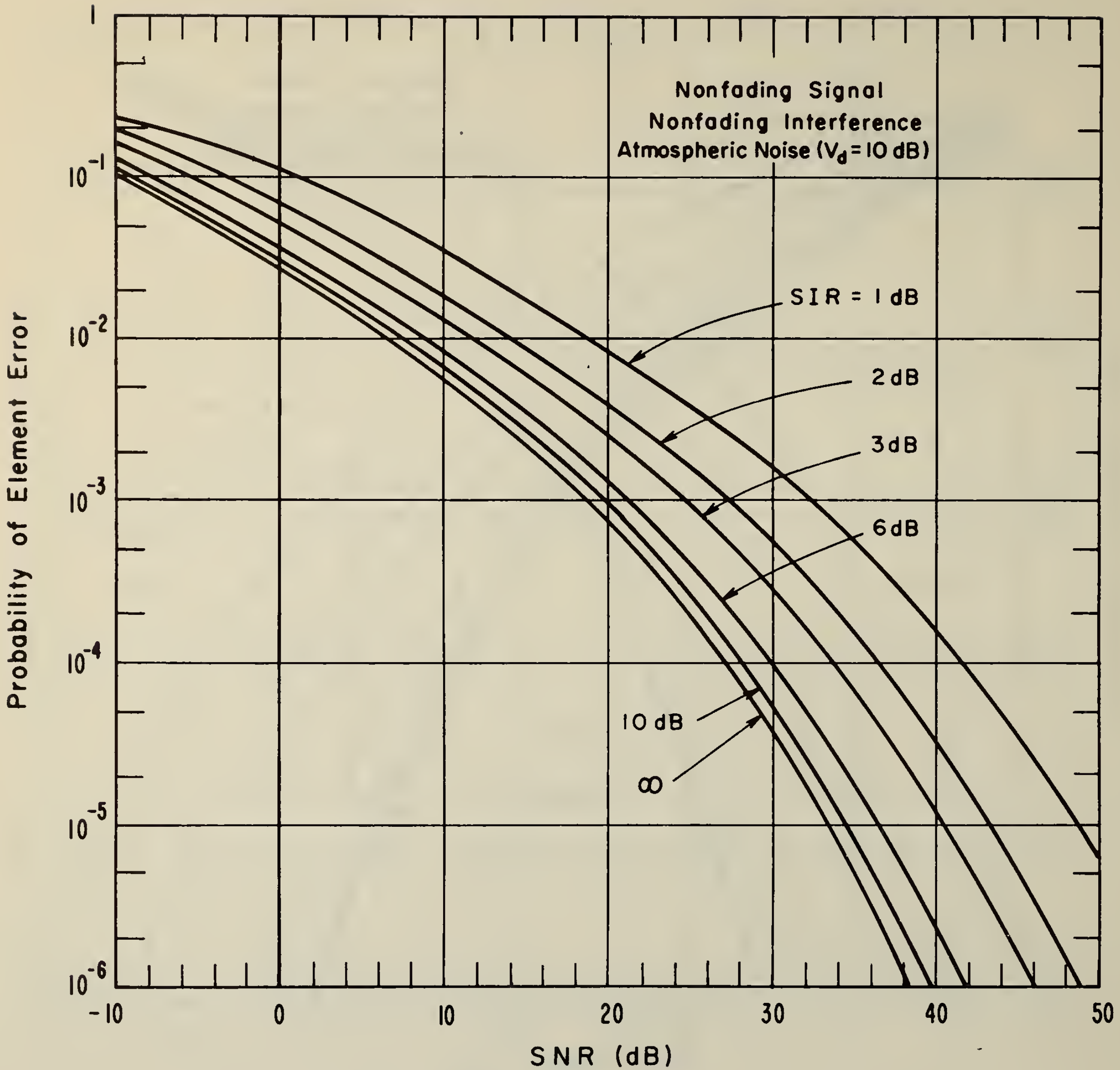

Figure 11. Element error probability in a single-channel NCFSK system under stable conditions with atmospheric noise (V_d = 10 dB) and nonfading interference.

compared with 9, 5, and 2 dB for Gaussian noise. The corresponding values are 6, 3, and 1 dB for $V_d = 6$ dB, and 6, 2, and 1 dB for $V_d = 10$ dB.

3.6. Fading Signal and Fading Interference Without Noise (Nondiversity Reception)

When both the desired signal and the interference are Rayleigh-fading the element error probability is the same as that for a Rayleigh-fading desired signal in the presence of Gaussian noise and no interference, because the APD of the Rayleigh-fading interference is the same as that of Gaussian noise. For a Rayleigh-fading signal and Gaussian noise, Montgomery (1954) showed that the element error probability is given by

$$P_e = \frac{1}{2} \cdot \frac{1}{1+R}, \tag{6}$$

where R is the ratio of average signal power to average noise power. This relation applies to our case, if R is interpreted as the SIR. (Although the SIR is defined here as the ratio of median desired signal power to median interfering signal power, it is equal to the ratio of average desired signal power to average interfering signal power when both signals follow an identical distribution law.) This relation for nondiversity is plotted in figure 12, section 3.7.

For log-normal-fading signals, desired and interfering, a linear transformation of the coordinate system can reduce the double integral in (1) to a single integral of the form

$$P_e = \frac{1}{2} \cdot F\left(\frac{m_s - m_i}{\sqrt{\sigma_s^2 + \sigma_i^2}} \right), \tag{7}$$

where

$$F(x) = \frac{1}{\sqrt{2\pi}} \int_{x}^{\infty} \exp\left(-\frac{t^2}{2}\right) dt, \tag{8}$$

m_s and m_i are the median levels in dB of the desired and interfering signals, respectively, and σ_s and σ_i are the respective standard deviations, also in dB. The $\sigma_{s,i}$ is related to the fading range R_f by

$$R_f = 2.563\,\sigma. \tag{9}$$

The relation between SIR and element error probability for log-normal fading (fading range = 13.4 dB) and nondiversity is shown in figure 13, section 3.7.

3.7. Fading Signal and Fading Interference Without Noise (Diversity Reception)

We start with quadruple space and frequency diversity with a selection-switching combiner: two modulated signals (both corresponding to an identical digital signal) are transmitted simultaneously in two channels separated in frequency, channel 1 and channel 2, and are received on two antennas separated in space, antenna 1 and antenna 2; the strongest signal among the four received signals is selected at the diversity combiner. The probability of element error in an NCFSK system is then given by

$$P_e = \frac{1}{2} \Pr\left\{ U_{a_1c_1} > D_{a_1c_1},\ V_{a_1c_1} = \max(V_{a_ic_j}) \right\}$$

$$+ \frac{1}{2} \Pr\left\{ U_{a_1c_2} > D_{a_1c_2},\ V_{a_1c_2} = \max(V_{a_ic_j}) \right\}$$

$$+ \frac{1}{2} \Pr \left\{ U_{a_2c_1} > D_{a_2c_1},\ V_{a_2c_1} = \max(V_{a_ic_j}) \right\}$$

$$+ \frac{1}{2} \Pr \left\{ U_{a_2c_2} > D_{a_2c_2},\ V_{a_2c_2} = \max(V_{a_ic_j}) \right\}, \qquad (10)$$

where

$D_{a_ic_j}$ = voltage of desired signal from antenna i in channel j ($i = 1, 2; j = 1, 2$),

$U_{a_ic_j}$ = voltage of undesired signal (interference) from antenna i in channel j ($i = 1, 2; j = 1, 2$),

$V_{a_ic_j}$ = vector sum of $D_{a_ic_j}$ and $U_{a_ic_j}$

= voltage of signal plus interference from antenna i in channel j ($i = 1, 2; j = 1, 2$),

and $\max(V_{a_ic_j})$ is a voltage having a maximum amplitude among the four voltages, $V_{a_1c_1}$, $V_{a_1c_2}$, $V_{a_2c_1}$, and $V_{a_2c_2}$. From this general form we shall derive three useful relations.

First, we assume quadruple space and frequency diversity reception with an interfering signal in channel 1 only. Since no interfering signal exists in channel 2, both $U_{a_1c_2}$ and $U_{a_2c_2}$ are zero and cannot exceed $D_{a_1c_2}$ and $D_{a_2c_2}$, respectively; therefore, the second and the fourth terms of (10) are zero. Since, in general, $D_{a_2c_1}$ and $U_{a_2c_1}$ follow the same probability distributions as $D_{a_1c_1}$ and $U_{a_1c_1}$, respectively, the third term of (10) should be equal to the first term. Therefore, we have

$$P_e = \Pr\left\{U_{a_1c_1} > D_{a_1c_1},\ V_{a_1c_1} = \max(V_{a_ic_j})\right\}$$

$$= \Pr\left\{U_{a_1c_1} > D_{a_1c_1},\ V_{a_1c_1} > D_{a_1c_2},\ V_{a_1c_1} > V_{a_2c_1},\ V_{a_1c_1} > D_{a_2c_2}\right\} \tag{11}$$

as an expression of element error probability in this case.

Second, simplifying the reception scheme, we assume dual frequency diversity reception with an interfering signal in channel 1 only. Since we do not have antenna 2, the third and the fourth terms of (10), based on voltages from antenna 2, are zero. In addition, since no interference is present in channel 2, the second term of (10) is zero. Therefore, we obtain

$$P_e = \frac{1}{2} \Pr\left\{U_{a_1c_1} > D_{a_1c_1},\ V_{a_1c_1} = \max(V_{a_ic_j})\right\}$$

$$= \frac{1}{2} \Pr\left\{U_{a_1c_1} > D_{a_1c_1},\ V_{a_1c_1} > D_{a_1c_2}\right\} \tag{12}$$

for this case.

Finally, we consider dual space diversity with an interfering signal. Since we do not have channel 2, the second and the fourth terms of (10), based on voltages in channel 2, are zero. In addition, as in quadruple diversity reception, the third term of (10) is equal to the first term; therefore, we have

$$P_e = \Pr\left\{U_{a_1c_1} > D_{a_1c_1},\ V_{a_1c_1} = \max(V_{a_ic_j})\right\}$$

$$= \Pr\left\{U_{a_1c_1} > D_{a_1c_1},\ V_{a_1c_1} > V_{a_2c_1}\right\} \tag{13}$$

for this case.

Using (11), (12), and (13), we evaluated the probability of element error for quadruple space and frequency diversity, dual frequency diversity, and dual space diversity, respectively. A Monte Carlo method was used for the evaluation, which simulated error performance tests with a total of 10^6 bits for each type of diversity. Results of these evaluations are shown for Rayleigh fading in figure 12 and for log-normal fading in figure 13. They are also compared with the previously derived results for nondiversity reception. Note that the gain in required SIR expected from the use of diversity depends on the type of fading of signal and interference. Only a small gain can be expected when both desired and interfering signals are fading with a log-normal distribution; for example the gain expected from quadruple space and frequency diversity instead of nondiversity is 6 dB at an element error probability of 10^{-3}.

3.8. Fading Signal and Fading Interference With Noise

Except for some special cases, the element error probability must be computed by numerically integrating the joint probability density function of the desired signal, interfering signal, and noise. Perhaps, the Monte Carlo method would be useful for this and might be mandatory for diversity reception. Even with the Monte Carlo method, the computation is expensive and time consuming, if we try to obtain data for numerous combinations of input parameters, such as types of fading, diversity techniques, noise parameter, SNR, and SIR.

For Rayleigh-fading desired and interfering signals and Gaussian noise with nondiversity reception, the element error probability can be calculated analytically. Since the probability distribution function of the composite wave of a Rayleigh-fading interference and Gaussian noise is the same as that of Gaussian noise, the element error probability can be computed by (6). In this case, however, we must use,

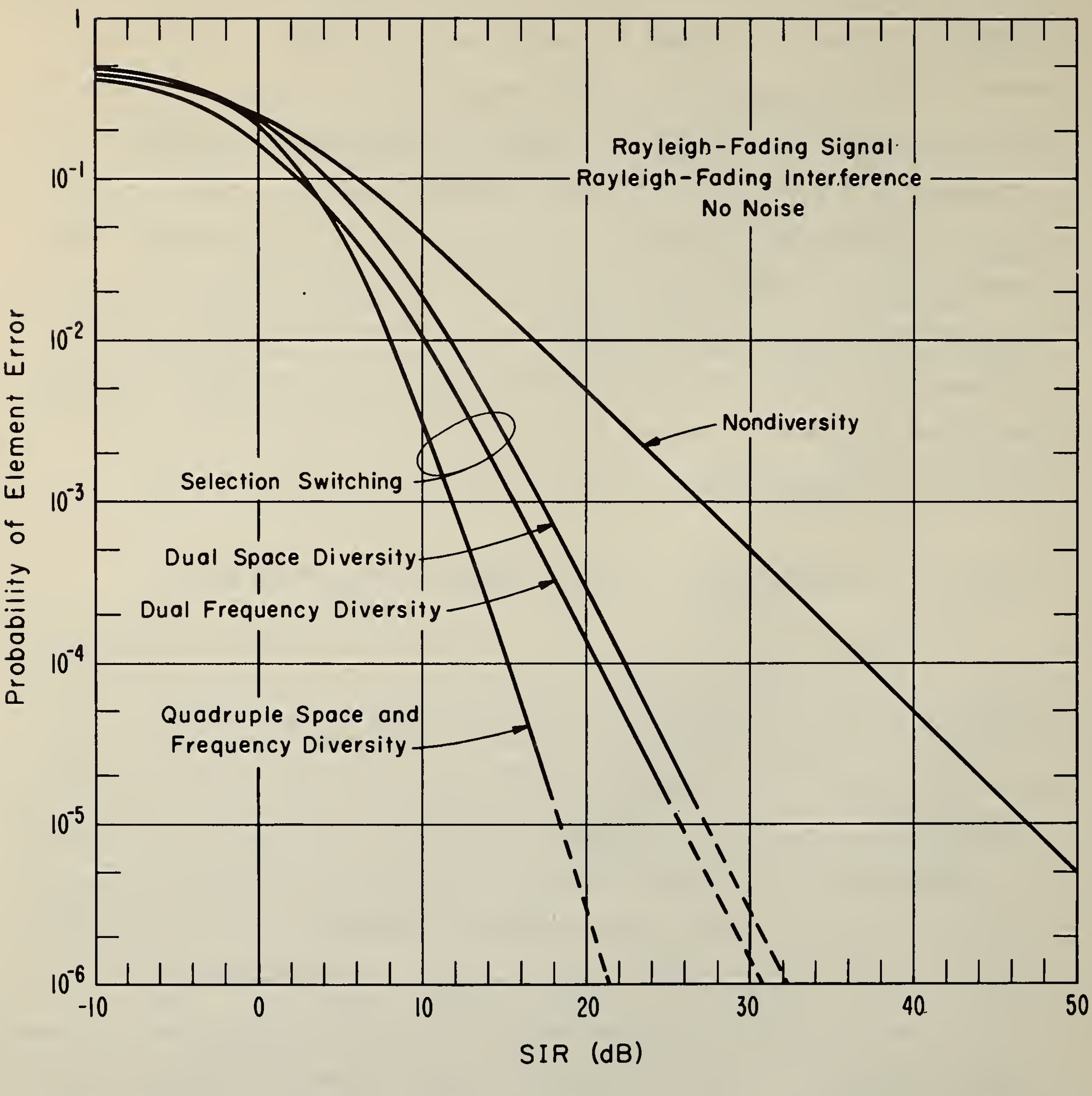

Figure 12. Element error probability in a single-channel NCFSK system under Rayleigh-fading conditions with Rayleigh-fading interference and no noise.

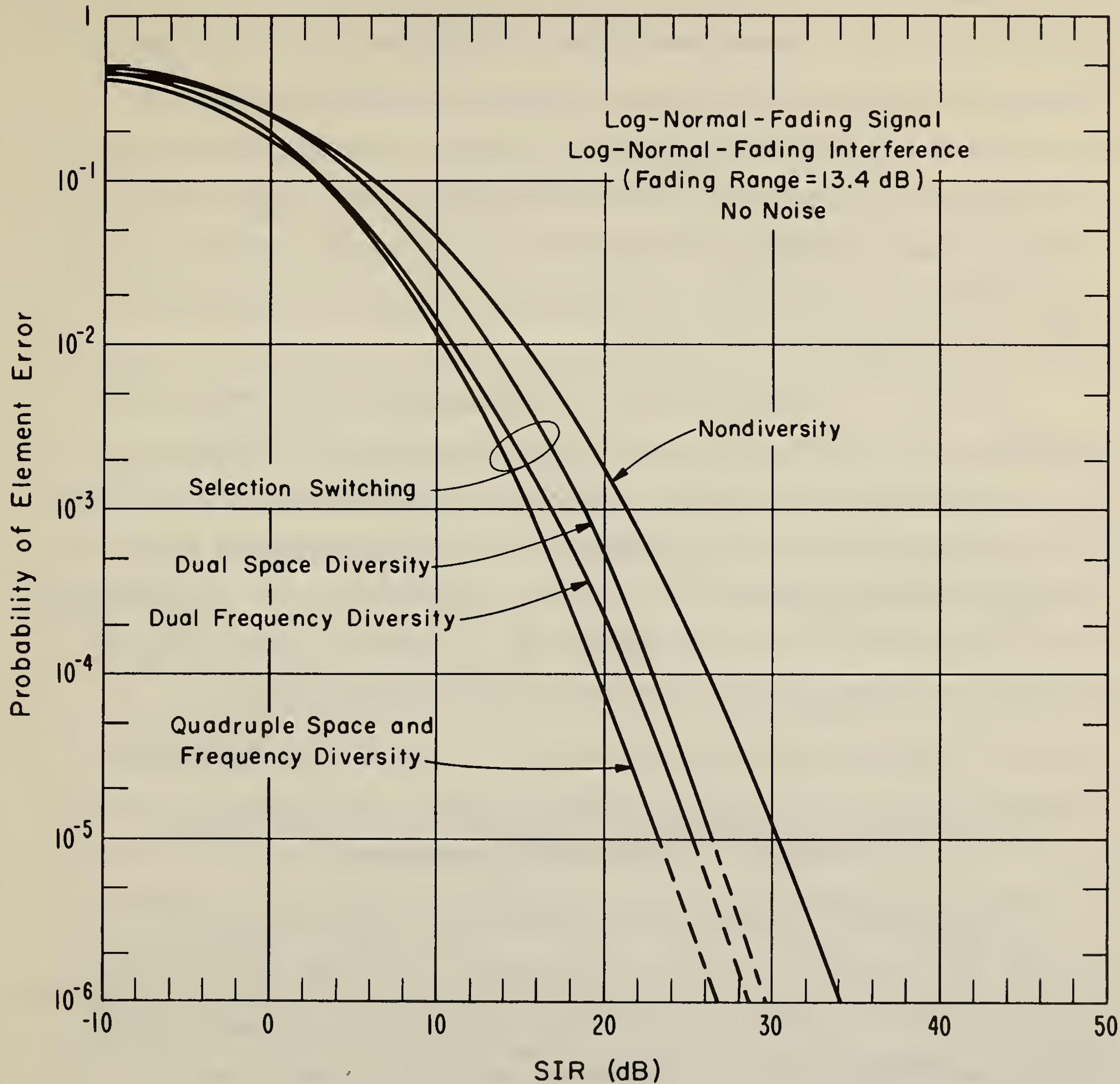

Figure 13. Element error probability in a single-channel NCFSK system under log-normal fading conditions with log-normal fading interference and no noise.

as the value of R, the ratio of the average desired signal power to the average power of the composite wave of the interference and noise.

For a Rayleigh-fading desired signal with nondiversity reception, the element error probability is almost independent of the probability distribution function of undesired signals in the lower error probability region. It depends only on the ratio of the desired signal power to the total power of undesired signals. Therefore, the element error probability in the lower error probability region can be read from figure 2, if we use, as the SNR in the abscissa of this figure, the ratio of the median desired signal power to the average power of the composite signal of the interference and noise.

Except for a Rayleigh-fading desired signal with nondiversity reception, the element error probability is higher with a higher V_d value in the range of error probability of our interest, say 10^{-3} or less. Since a composite wave of interference and noise is generally less impulsive than the noise alone, each curve in figures 3 to 7 gives an upper bound of the element error probability for a particular value of V_d, if the SNR used is the same as used in the preceding paragraph.

4. EFFECTS OF ADVERSE PROPAGATION CONDITIONS ON THE SYSTEM PERFORMANCES

HF signals propagated by the ionosphere suffer a number of peculiar distortions that affect the selection of a modem and the design of the telecommunication network. For some types of distortion both NCFSK and TDPSK systems deteriorate similarly; other types of distortion affect these systems in different ways. The effects of ionospheric distortions on NCFSK and TDPSK systems have been analyzed in detail by Bello and Nelin (1962, 1963, 1964) and comprehensively surveyed from the standpoint of FDM data transmission systems design (Akima, 1967). A discussion of these effects in relation to the IGOSS, and of a serious drawback inherent in a TDPSK system, follows.

4.1. Multipath Time-Delay Spread

Complex paths with both low and high rays having a different number of ionospheric and ground reflections are a major source of multipath long-time delay spread. Multipath spread is a function of the distance between the transmitter and the receiver and of the multipath reduction factor, defined as the ratio of the operating frequency to the highest of the classical MUFs (maximum useful frequencies) for simple E, F_1, and F_2 modes (Salaman, 1962). A spread of 1 to 2 ms is fairly common. The spread becomes smaller and smaller as the multipath reduction factor approaches unity. However, in a scheduled operation with a limited choice of operating frequencies, as for the IGOSS, it is virtually impossible always to achieve operation with a multipath reduction factor close to unity.

To minimize the deleterious effect of multipath time-delay spread in both NCFSK and TDPSK systems, we must make the element length relatively long compared with the multipath time-delay spread. For this reason the minimum element length is usually about 10 ms, resulting in a maximum modulation rate of approximately 100 bauds. For the IGOSS, the final recommended choice of element length will depend on the above factors and on considerations of overall system efficiency, i.e., maximizing the total amount of usable information that can be obtained in a given time and with a given number of operating frequencies and channels.

4.2. Frequency Fluctuations

The instantaneous frequency of a received signal fluctuates due to propagation medium characteristics (Doppler effects) regardless of the stability of the transmitter frequency. Long-term changes in frequency (10 min or longer) are relatively small, usually less than a fraction of 1 Hz, but short-term changes (a few minutes or shorter) can be much greater and may be as much as tens of Hz under very severe conditions (Davies and Baker, 1966).

Frequency changes in signals in adjacent channels (originating from different locations) are not necessarily in the same direction, because the propagation paths for the two signals are different. Thus, in either an NCFSK system or a TDPSK system, channel spacings wider than those commonly used in FDM modems are recommended for circuits where such severe conditions may be expected (e.g., auroral and transequatorial circuits).

Another logical solution is that propagation paths exhibiting severe frequency fluctuations should be avoided as much as possible when the configuration of the IGOSS is designed. In practice, this is limited to choosing shore station locations and to considering these effects in preparing operational schedules.

To combat the effect of frequency fluctuations on an NCFSK system, a wider frequency shift and, thus, a wider channel bandwidth than those required by the modulation rate are necessary. Although they require a higher signal level for maintaining the same SNR, these requirements do not conflict with other constraints, such as the longer element length dictated by multipath time-delay spread. Thus, for an NCFSK system, adverse propagation condition can be accommodated at the cost of a decreased data signalling capacity in the 3-kHz bandwidth and increased transmitter power. For this system, one possibility would be for severe circuits to use twice the values of deviation and channel bandwidth used for less severe circuits.

Frequency fluctuation created by transmitter and receiver frequency instabilities is discussed in section 5.

4.3. Phase Fluctuations

Several observations of phase fluctuation of signals propagated by the ionosphere have been reported (Koch and Beery, 1962; Porter et al., 1963; David et al., 1965). They indicate an increase in phase fluctuations with an increase in the fading rate for a given sampling interval (e.g., 5 or 10 ms) and with an increase in the sampling interval for a given fading rate. For example, the probabilities that phase fluctuations exceeded 90° for sampling intervals of 8 ms and 20 ms over a transauroral path were 1 percent and 5 percent, respectively, when the fading rate was 7 Hz (Koch and Beery, 1962).

Phase fluctuations limit the maximum element length that can be used in TDPSK systems, because in these systems decisions are made by comparing the phases of two successive elements. To avoid the harmful effect of phase fluctuations in TDPSK systems, element length must be kept as short as possible. The minimum length, however, is limited by the multipath time-delay spread in data transmission channels over HF ionospheric paths, as noted in section 4.2. The fact that adverse propagation conditions impose these basically conflicting requirements is perhaps the most serious drawback of TDPSK systems for data transmission over HF ionospheric paths. It is very difficult, if even possible, to find an element length suitable for the range of propagation conditions that may be encountered in the IGOSS. Furthermore, we cannot escape from these two contradictory requirements, even if we are willing to reduce the data signalling rate in the 3-kHz bandwidth.

Frequency fluctuation or offset caused by unstable or inaccurate transmitter and receiver frequencies can be translated into phase fluctuation and phase offset. This is discussed in section 5.

5. FREQUENCY STABILITY

Recent advances in frequency stabilization techniques allow us to obtain, at an increased cost for a higher frequency stability, oscillators that are as stable as desired for most communication applications. Therefore, frequency stability to be specified for a transmitter or a receiver should be determined as a compromise between the loss in bandwidth use (or system performance) due to increased frequency tolerances and the increased cost for higher stability of the oscillator frequency.

In this study we will consider only frequency stability of a platform transmitter, because the frequency of a shore-station receiver can be stabilized to the desired extent at a small cost compared with the total cost of the shore station facilities.

The following discussions suggest that the frequency stability to be specified for the platform transmitters lies between approximately one part in 10^7 and one part in 10^6. The oscillator for 10^{-7} stability is expensive but achieves high efficiency in bandwidth use. The oscillator for 10^{-6} stability is less expensive but can only be used with a penalty in bandwidth and, therefore, in system transmission time. A final decision should be based on many factors, some of which are beyond the scope of this study.

5.1. Frequency Tolerance for NCFSK Systems

The CCIR (1967d) gives a value of 3 Hz as the frequency tolerance required for reception without automatic frequency control of A7 emissions (multichannel voice-frequency telegraphy) in the 4 to 29.7 MHz band. This tolerance corresponds to a frequency stability of 1.4×10^{-7} at 22 MHz, which is the highest frequency allocated for the IGOSS, by the World Administrative Radio Conference (WARC) held in Geneva in 1967. A frequency stability of 10^{-7} can be achieved within state-of-the-art techniques; therefore, this is one of the prospective values for the IGOSS.

Even though it can be achieved, a frequency stability of 10^{-7} requires a high quality frequency source. We shall discuss whether it is possible, and if so to what extent, to relax the tolerance beyond 3 Hz for the IGOSS.

Montgomery (1954) showed that, as long as the instantaneous signal amplitude exceeds the instantaneous noise amplitude, the discriminator input frequency cannot deviate from the signal frequency by more than 0.5 R Hz, where R is the modulation rate in bauds. The CCIR (1967b) has reported the optimum deviation for a modulation rate of R bauds to be ± 0.4 R Hz for HF radio circuits, with a required minimum bandwidth (at the -3 dB points) of R Hz. This indicates that the probability that the discriminator input frequency deviates from the signal frequency by more than 0.4 R Hz is negligibly small, though not zero, as long as the instantaneous signal amplitude exceeds the instantaneous noise amplitude. Using numerical examples we now discuss the possibility of relaxing the tolerance.

The first example we assume is an NCFSK system with a frequency deviation of ± 42.5 Hz, a channel separation of 170 Hz, and a channel filter bandwidth of 120 Hz at the -3 dB points. These values of frequency deviation and channel separation are the same as in FDM NCFSK systems, as recommended by the CCIR (1967b), operating at a modulation rate of approximately 100 bauds over HF radio circuits. In the assumed system with a deviation of ± 42.5 Hz and a bandwidth of 120 Hz, the noise bandwidth is slightly wider than that of the optimum system, but since the difference is small, we may assume, to a first approximation, that a frequency error of $42.5 - 0.4\,R = 42.5 - 40 = 2.5$ Hz would result in negligible degradation of the performance of our system. This value is close to the value of 3 Hz given by the CCIR (1967d).

To relax the frequency tolerance of this system without increasing the error probability, we must reduce the modulation rate while keeping the bandwidth the same. Since the modulation index is now very different from the optimum system it is better to rely on the criterion derived by Montgomery (1954). In this case, the frequency tolerance is equal to $42.5 - 0.5\,R$ Hz, i.e., $42.5 - 37.5 = 5$ Hz and $42.5 - 25 = 17.5$ Hz, when the modulation rate is reduced to 75 and 50 bauds, respectively.

These values of the frequency tolerance apply when the signal frequency deviates toward the center of the channel. However, when the signal frequency deviates toward the edge of the channel filter, the signal also degrades. The tolerance against the frequency error in this direction can be estimated as follows. As quoted earlier, the required minimum bandwidth (at the -3 dB points) is R Hz for the optimum deviation of $\pm 0.4\,R$ Hz. Therefore, the difference between the -3 dB point of the filter and the signal frequency must be at least $R/2 - 0.4\,R = 0.1\,R$ Hz. In the assumed system with a 120-Hz bandwidth, the frequency error cannot exceed $120/2 - 0.1\,R - 42.5 = 17.5 - 0.1\,R$ Hz; therefore, we have 7.5, 10, and 12.5 Hz as the values of the frequency tolerance for R = 100, 75, and 50 bauds, respectively.

Combining these values with those obtained in the preceding paragraph and taking the smaller one, we have 2.5, 5, and 12.5 Hz as the values of frequency tolerance for the modulation rates of 100, 75, and 50 bauds, respectively. These tolerances roughly correspond to frequency stabilities of 10^{-7}, 2×10^{-7}, and 5×10^{-7} at the highest transmitting frequency of 22 MHz .

The second example we assume is an NCFSK system with a frequency deviation of ± 85 Hz, a channel separation of 340 Hz (twice the values given in CCIR, 1967b), and a channel filter bandwidth of 240 Hz at the -3 dB point. In this system the bandwidth use is halved and thus the total system transmission time is doubled, as compared with the system having a ± 42.5 Hz frequency deviation. Since the channel bandwidth is doubled, the transmitter power should be approximately 3 dB higher to keep the error probability the same for an equal modulation rate. Applying the same argument as in the first example, we have 85 - 0.5 R Hz as the tolerance for the frequency error toward the center of the channel filter and 240/2 - 0.1 R - 85 = 35 - 0.1 R Hz for the error toward the edge of the channel filter. The first equals 35, 47.5, and 60 Hz for the modulation rates of 100, 75, and 50 bauds, respectively, and the other equals 25, 27.5, and 30 Hz for the same modulation rates. Thus, the frequency tolerance in this example is determined only by the latter and is equal to 25, 27.5, and 30 Hz for the modulation rates of 100, 75, and 50 bauds, respectively. These values of frequency tolerance roughly correspond to frequency stabilities of 1.1×10^{-6}, 1.2×10^{-6}, and 1.4×10^{-6} at the highest transmitter frequency of 22 MHz.

In summary, the frequency tolerance of 3 Hz can be relaxed to 12.5 Hz at the cost of halving the modulation rate and thus doubling the transmission time in the first system, and relaxed to 25 Hz at the cost of doubling the bandwidth in the second system, which results in doubling the total system transmission time and requires twice the required transmitter power.

5.2. Frequency Tolerance for TDPSK Systems

In a binary TDPSK system the difference between the received frequency and the allocated frequency results in a loss in the signal amplitude by a factor of $\cos(2\pi \cdot \Delta f \cdot T)$ and, accordingly, a loss in the required SNR by the same amount, where Δf is the frequency difference and T is the element length of the digital information signal. If we assume that a 1-dB increase in the required SNR is allowed, $\Delta f \cdot T$ can be as large as 0.075, and from this we obtain 7.5, 5.6, and 3.8 Hz as frequency tolerances for modulation rates of 100, 75, and 50 bauds, respectively. Thus, in a TDPSK system the frequency tolerance becomes more stringent when the modulation rate is reduced. When a 3-dB increase in the required SNR is allowed, $\Delta f \cdot T$ can be 0.125, and the frequency tolerances can be relaxed to 12.5, 9.4, and 6.25 Hz for modulation rates of 100, 75, and 50 bauds, respectively. However, $\Delta f \cdot T$ increases as only a square root of the increase in the required SNR in dB; thus frequency tolerance cannot be relaxed greatly by allowing a large increase in the required SNR.

In a quaternary TDPSK system the frequency difference not only causes a loss in the signal amplitude of the in-phase component by a factor of $\cos(2\pi \cdot \Delta f \cdot T)$, but also produces a quadrature-component response of magnitude $\sin(2\pi \cdot \Delta f \cdot T)$. An error occurs in this system when the amplitude of the quadrature component of signal plus noise exceeds the amplitude of the in-phase component of signal plus noise. The error in this system can be compared to the error in an NCFSK system with a dual-filter demodulator, where an error occurs when the amplitude of an interference plus noise in a filter exceeds the amplitude of the desired signal plus noise in the other filter. Therefore, the error probability in a quaternary TDPSK system can be discussed in approximately the same manner as in an NCFSK system in the presence of interference and noise. At $\Delta f \cdot T = 0.05$ the loss in the signal amplitude of the in-phase component is 0.4 dB, and the equivalent SIR (that is the ratio of the in-phase component of the signal to the quadrature component) is about 10 dB. Since an increase of 1 to 2 dB is necessary in the required SNR by an interfering signal at the level of SIR = 10 dB (see figs. 8 through 11), this value of $\Delta f \cdot T$ is about the maximum tolerance allowable in a quaternary TDPSK system. This corresponds to frequency tolerances of 5, 3.75, and 2.5 Hz at modulation rates of 100, 75, and 50 bauds, respectively.

In summary, the frequency tolerance for a TDPSK system operating at a modulation rate of 50 to 100 bauds lies in a range between 2.5 and 12.5 Hz, depending on the modulation rate, the loss in the required SNR that can be tolerated, and the number of phases used in modulation. These values correspond to frequency stabilities of 10^{-7} to 5×10^{-7} for a transmitter frequency of 22 MHz.

5.3 Frequency Stabilization

Quartz crystal oscillators are most widely used for stabilizing transmitter frequencies. Rubidium-beam or cesium-beam oscillators are also used when higher stabilities are required. The CCIR (1967c) gives data on the present status of frequency stabilization techniques that rely on quartz crystals.

There are two ways of generating a necessary frequency, one by using a crystal cut for that frequency and the other by using a frequency synthesizer. A frequency synthesizer has a master oscillator of high stability and synthesizes the desired frequency from the single frequency source by multiplication, division, and addition processes; the stability of the synthesized frequency thus equals that of the source.

In services that require changes over a number of transmitter frequencies, all frequencies can be (1) generated with individual crystals, (2) synthesized from a master oscillator frequency by a common variable-frequency synthesizing circuit, or (3) synthesized from a master oscillator frequency by individual fixed-frequency synthesizing circuits. The first method has been exclusively used and is still widely used even now but becomes progressively more expensive if a high stability is required. For a frequency stability of 10^{-7} per year the estimated current cost is approximately \$800 to \$1,000 per required frequency. The second method costs \$7,000 to \$11,000 including \$2,000 for the master oscillator; additional circuitry for remote control of the frequency is generally required when this method is used in an unmanned station. The third method costs approximately \$2,400 + \$200N including \$2,000 for the master oscillator and \$400 for the power supply for the synthesizing circuits, where N is the number of frequencies: a remote control circuit for this method is much simpler than that for the second method. The third method, i.e., a master oscillator and fixed-frequency synthesizing circuits, seems to be the most promising for our application.

In an FSK transmitter, frequency stability also depends on the way of modulation. If two stabilized frequencies are generated and one of them is selected according to the binary information signal, frequency stability of the FSK signal is equal to that of the stabilized frequencies. But, if a variable frequency oscillator (VFO) output, frequency-shift-keyed by the binary signal, is heterodyned with a stabilized oscillator output, the stability of the FSK signal may be worse than that of a stabilized oscillator. Because of a possible service interval of 1 year, this last method does not look very attractive for the IGOSS.

Assuming that an NCFSK signal is to be transmitted in one of 15 channels (such as recommended by the CCIR, 1967b) in one of the six HF bands allocated by the WARC in 1967, we shall briefly discuss several possible schemes of generating the platform transmitter frequencies. Generating an NCFSK signal is equivalent to generating one of $6 \times 15 \times 2 = 180$ possible frequencies f_{ijk} ($i = 1, 2, \ldots, 6$; $j = 1, 2, \ldots, 15$; $k = 1, 2$) according to the band-selecting signal i, the channel-selecting signal j, and the binary data signal k.

One extreme scheme is to generate all possible frequencies simultaneously and to select one of them. All the frequencies could be generated separately from individual crystals or synthesized by appropriate fixed-frequency synthesizing networks. But, in any case, this scheme has redundant circuit construction and does not seem to be practical.

The other extreme scheme is to use a single variable-frequency synthesizer. Although only one master oscillator is required in this scheme, a variable-frequency synthesizing network is rather expensive ($7,000 - $11,000, including the master oscillator), and the control circuitry is also fairly complicated.

Another scheme can be as follows: one frequency is selected by a band-selecting switch out of six that correspond to the six HF bands; one by a channel-selecting switch out of 15 frequencies corresponding to the 15 channels; one by a data signalling switch (frequency-shift-keyer) out of two frequencies corresponding to "0" and "1" of the binary data signal. The output signals from the three switches are then mixed so that the frequency of the output signal is the sum of the frequencies of the three signals. The necessary 6 + 15 + 2 = 23 frequencies for this scheme can be synthesized from a master oscillator frequency by 23 fixed-frequency synthesizing networks. The cost is estimated at $7, 000, which includes $2, 000 for a master oscillator, $200 for each synthesizing network, and $400 for a power supply.

The scheme described in the preceding paragraph is not the only possible one; it can be modified in several ways. Since the 15 channel frequencies are separated by equal intervals, the number of necessary frequencies can be reduced by generating the 15 channel frequencies in two steps. For example, we can generate a subgroup of five frequencies and heterodyne the selected frequency with one of three frequencies. The number of frequencies required is then 6 + 5 + 3 + 2 = 16, reducing the estimated cost to approximately $5, 600, a substantial saving over $7, 000.

6. SELECTIVITY OF RECEIVERS

The selectivity of a receiver is a measure of its ability to discriminate between wanted and unwanted signals. Criteria for establishing the selectivity of a receiver and definitions to be used for the purpose of studying the selectivity are recommended by the CCIR (1967a). Some data for the selectivity characteristics of various classes of receivers and several methods of measuring selectivity are also described by the CCIR (1967a, e, f).

This selectivity is expressed in terms of the single-signal selectivity and the effective selectivity. The single-signal selectivity represents the selectivity in the linear region; it can be studied by measuring, with one signal, the passband, attenuation-slope, image-rejection ratio, intermediate-frequency rejection ratio, and other spurious-response rejection ratios. The effective selectivity includes the effects of amplitude nonlinearity; it can be investigated by measuring, with the wanted signal and an unwanted signal both present, blocking and adjacent-signal selectivity (adjacent-channel selectivity, if there is regular channelling); and by measuring, with two unwanted signals present, radio-frequency intermodulation distortion.

In this section we discuss requirements for the selectivity of IGOSS shore-station receivers and design factors necessary to meet those requirements.

6.1. Requirements for the Selectivity of IGOSS Shore-Station Receivers

Since the HF bands allocated to the IGOSS have a 250-Hz guard band on each side of them, it is easy to protect the IGOSS signals against unwanted signals of other services.

In the IGOSS, however, there is a peculiar problem of inter-channel interference. An IGOSS shore station must receive simultaneously, in a common 3-kHz bandwidth, as many signals as there are channels in the band. Since these signals come from geographically dispersed platforms, they fade independently and also may differ considerably in their median levels. Therefore, the requirements for selectivity are more severe in the IGOSS than in a frequency-division-multiplexed (FDM) data transmission system, where the signals in the different channels have an equal median level.

The difference in the median level of signals received in a common 3-kHz HF band is thus an important factor for specifying the selectivity of IGOSS shore-station receivers. As an example we have the following result computed by Hatfield and Adams (private communication). They assume a system operation plan in which a total of 5501 platforms are randomly located all over the world and each platform transmits observed data to one of 18 shore stations, in one of 13 channels preassigned to each platform, in one of the six HF bands. Using the method described by Barghausen et al. (1969), they computed the reliability for each possible combination of platform, shore station, and HF band; on the basis of this, they selected a systems assignment plan that would provide maximum total system reliability. Circuits using the same shore station, frequency band, and channel were assigned sequential time slots as the assignments occurred. With the system operation plan thus determined, statistics were taken of the median level difference between pairs of adjacent channels. The result was that the probabilities that the level difference exceeds 6, 9, 12, and 15 dB are approximately 15, 5, 2, and 1 percent, respectively.

This result indicates that significant level differences can occur if time slots are assigned sequentially without any precautions. However, we can limit large level differences by reassigning time slots, with a possible penalty of increased number of time slots and, accordingly, increased total transmission time. Since the probability of occurrence of large level differences, such as 15 dB or greater, is relatively small, we can expect that the possible increase in the number of time slots resulting from the elimination of level differences greater than 15 dB is also small. Therefore, we assume that the maximum difference in the median signal levels between a pair of adjacent channels will not exceed 20 dB, including 5 dB for the uncertainty in predicting median signal levels.

Another important factor to be considered in specifying the selectivity of a receiver is the maximum tolerable effect of interference.

Although the CCIR (1967a) uses the output power from the receiver to represent the effect of interference in the definition of the selectivity of a receiver, this is valid only for a receiver used for analog-signal transmission. For a receiver used for digital data transmission, the power at a point just before the decision-making circuit could be used. But here we use the power at the demodulator input to represent the effect of interference, because, from a practical standpoint, it is better suited for this purpose. Consequently the maximum tolerable amount of interference is in terms of the minimum SIR required at the demodulator input.

The minimum required SIR is a function of the allowable error probability and can be determined from figures 12 and 13. These figures indicate that the SIR at the demodulator input should be approximately 20 and 25 dB for element error probabilities of 10^{-3} and 10^{-4}, respectively, when dual space diversity is used.

If all the wanted and unwanted signal levels were always in the linear region of the receiver, 45-dB attenuation of signals in other channels would be sufficient for specifying the selectivity of an IGOSS shore-station receiver, because it would secure an SIR of 25 dB against an unwanted signal 20 dB stronger than the wanted signal. Since very strong signals may sometimes occur, this is insufficient, and the selectivity should be specified in terms of effective selectivity.

To specify effective selectivity, we must specify the level of the wanted signal at which to make the measurements. In our case the level of the wanted signal should be 20 dB less than the maximum signal level expected. This, in turn, depends on the overall plan of the IGOSS, and no reliable data are now available.

Adjacent-channel selectivity and intermodulation distortion are the most important factors governing systems performance, and as a result of the above discussion the following two requirements are recommended to specify the selectivity of IGOSS shore-station receivers:

(1) The SIR at the demodulator input shall be 25 dB or higher, when a wanted signal is applied to one channel and an unwanted signal is applied to another channel at a level of 20 dB stronger than that of the wanted signal.

(2) The SIR at the demodulator input shall be 25 dB or higher when a wanted signal is applied to one channel and two unwanted signals are applied, one to the adjacent channel on one side and the other next to the adjacent channel on the same side, both 20 dB stronger than the wanted signal.

The wanted signal level in both cases will be specified as soon as the overall plan of the IGOSS has been established in sufficient detail.

6.2. Receiver Structure

In general, the selectivity of a receiver is not only limited by the characteristics of the filter but also limited by unavoidable amplitude nonlinearities, e.g., cross-modulation of the wanted signal by strong unwanted signals. Since the signal level generally increases as it comes closer to the receiver output, there is more chance of a loss of effective selectivity caused by amplitude nonlinearities in the later stages of a receiver. As a general rule, the CCIR (1967a) recommends that the filters that determine the selectivity shall be included as near as possible to the receiver input, and the amplifying stages preceding the filter shall be sufficiently linear, to avoid significant loss of selectivity. For IGOSS shore-station receivers, this principle implies that a signal in each channel should be separated from signals in other channels as early as possible. This also implies that the number of common circuits shared by the signals in the different channels should be restricted as much as possible.

There are several choices; in principle, a signal in each channel can be separated at any stage in the receiver. In practice, however, we have two practical schemes from which we must choose one. One choice would be to separate each channel immediately after the first frequency converter, i. e., at the input of the intermediate-frequency amplifier (IFA); the other choice would be to separate the channels at audio frequency, as done in most FDM data-transmission receivers. The first choice may be called the multiple IFA scheme, because each channel has its own IFA; the second may be called the common IFA scheme, because all channels share a common IFA.

Although it would be difficult to prove the necessity of the multiple IFA scheme for the IGOSS shore-station receivers, there is no question that better effective selectivity can be achieved with this scheme than with the common IFA scheme. Moreover, adopting the multiple IFA scheme will allow the design of automatic gain control (AGC) circuitry providing independent control of each channel. Therefore, we recommend adopting the multiple IFA scheme.

The difference in cost between the two schemes is mainly the difference in the type of filters and the number of IFAs required. Since the center frequency of the filters for the multiple IFA scheme is about 1 MHz and the bandwidth is about 100 Hz, a crystal filter or a mechanical filter is required. The required selectivity suggests using a five-pole or six-pole crystal filter; its cost is roughly estimated at $150 to $200, compared with the cost of an audio-frequency filter for the common IFA scheme roughly estimated at $70 to $100. The second factor is that a multiple IFA receiver must have the same number of IFAs as the number of channels. This extra cost is roughly estimated at $50 to $70 per channel.

Besides the protection against possible interchannel interference within the IGOSS, we must protect the IGOSS signals against other unwanted signals. Therefore, we recommend that crystal filters of approximately 3- to 4-kHz bandwidth be used at the input of the receiver for each of the six HF bands, regardless of the schemes used for the receiver. Modern techniques enable designing such filters to have an insertion loss of only 1 to 2 dB.

The discussion given in this section applies equally to NCFSK or TDPSK.

7. CONCLUSIONS

To provide basic data relevant to selecting a modem and the design of the IGOSS telecommunication network, we have discussed modulation techniques for digital data transmission over HF ionospheric paths and some related topics.

On a preliminary basis, NCFSK and TDPSK systems were preferred; the performance of these systems in the presence of noise and/or interfering signal under normal propagation conditions has been analyzed. These analyses were made on a fairly general basis: nonfading and fading signals (either desired or undesired), Gaussian and atmospheric noise, nondiversity and diversity reception.

The effects of adverse propagation conditions, peculiar to HF ionospheric paths, on these preferred systems were also considered. The results indicate that a TDPSK system has a serious drawback when applied to the IGOSS, where propagation conditions may be widely different and adverse conditions are sometimes likely to occur.

Frequency tolerances allowed for the preferred modulation systems were studied, and some frequency stabilization techniques that can be used for the IGOSS platform transmitters are given. Requirements for the selectivity of the IGOSS shore-station receivers were determined, and the multiple IFA scheme is recommended as a receiver structure that satisfies these requirements.

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