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ALE Frequency Selection for Navy Modems: Analysis

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ADMINISTRATIVE INFORMATION

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EXECUTIVE SUMMARY

OBJECTIVE

Develop a simple model for HF (2–30 MHz) modem performance as a function of various channel parameters. Develop techniques for predicting those parameters.

RESULTS

Modem performance was modeled based on three parameters—signal-to-noise ratio (SNR), delay spread, and Doppler spread. Each of these parameters is measurable using known techniques. The model demonstrates the inadequacy of using only SNR and bit-error rate estimates for frequency selection in automatic link establishment (ALE) algorithms.

RECOMMENDATIONS

The benefits from incorporating delay and Doppler spreads in the ALE frequency selection algorithm are sufficiently clear to justify extension of the algorithms to include these factors. Further research is required to verify the sufficiency of SNR, delay spread, and Doppler spread for frequency selection.

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INTRODUCTION

BACKGROUND

Standards for high frequency radio automatic link establishment (ALE) are being developed at the federal government level, within the Federal Standards series. These same standards are being accepted and promulgated, in slightly different format, by the Department of Defense as MIL-STD-188-141A. The standards support automated means for selecting the "best" operating frequency for HF communications between two locations. The standards define both the transmission waveforms and frequency selection algorithms.

The waveform selected for the ALE standard (8-ary FSK) is not the same as either of the contemporary waveform types used by the Navy (multitone DQPSK; equalized, high data-rate single-tone DQPSK). It is not clear that the frequency that's best for the ALE modem will also be best for a Navy data modem. Early ALE on-the-air testing (DCA, 1988) suggests the possibility of such a performance disparity between ALE and the Harris RF3466, a 39-tone DQPSK modem.

The military standard does not fully specify the frequency selection algorithm. The standard provides the potential to exchange both SNR data and supplementary channel quality data. The standards developers recognized the potential future need to exchange channel measurement data such as multipath spread and Doppler spread.

OBJECTIVE

The objective of the effort described was to identify a small set of parameters to augment SNR for link quality assessment; and a link quality measure based on received SNR plus those parameters, such that ALE will select the best frequency for a given Navy modem. We anticipated that the link quality algorithm will be a function of which modem is to operate over the selected frequency. A further objective is these parameters should be measurable using the ALE hardware and waveforms.

APPROACH

The investigation relied upon existing analytic results for the modulations implemented in existing Navy modems and for the modulation defined for ALE. Due to time and funding limitations, we further narrowed the scope to focus on Navy parallel tone DQPSK modems. We considered two channel parameters; root-mean-square (rms) Doppler spread and rms-delay spread. These parameters were chosen because they have a strong connection to the analytical results, relatively simple techniques have been developed for estimating them on real channels (Perl and Kagan, 1986), and they can be easily accommodated within the ALE data-transfer structure already proposed for exchanging channel information. Existing analytic results were modified to describe the Navy and ALE modems.

This report presents the results of the analytic investigation. Future work will compare the analytic results with simulation results obtained using a Watterson-type HF channel simulator.

The analytic modem performance models were examined to identify benefits available from incorporating rms-delay spread measurements in the ALE frequency selection process.

ANALYTICAL ASSESSMENT

Although modem performance has been the subject of analyses for at least four decades, there still is no expression for performance of even the most common modulation schemes incorporating a set of channel and hardware performance parameters. However, many of these parameters are treated independently. The functional form of the performance for the common modulation types has been derived. We will not attempt to address performance through new mathematical analyses. Rather, we will exploit existing analyses to build a plausible model for empirical performance estimation.

GAUSSIAN NOISE

The simplest analysis concerns binary modem performance in the presence of additive stationary white Gaussian noise (Schwartz, Bennett, and Stein, 1966). There are two forms for bit error probability (P_b) expressions covering modulations of interest to us. The first of these is

$$P_b \approx \frac{1}{2}\exp[-kR] \tag{1}$$

where R is the signal energy-to-noise ratio, $R = E_b/N_0$. Theoretical analyses give k = 1 for binary DPSK and k = 1/2 for noncoherent FSK in this expression. The second form of the error probability expression is

$$P_b \approx \frac{1}{2}\mathrm{erfc}[(kR)^{1/2}] \ . \tag{2}$$

Here erfc is the complementary error function. The theoretical analyses show k = 1/2 for coherent FSK, and k = 1 for coherent PSK in this expression.

For m-ary DPSK, the symbol error probability (P_E) expression is

$$P_E \approx \operatorname{erfc}[(kR)^{1/2}] . \tag{3}$$

An excellent approximation for m-ary DPSK, due to Arthurs and Dym (1962), can be found by using

$$k \approx \sin^2\left(\frac{\pi}{m\sqrt{2}}\right) \tag{4}$$

in equation 3.

In each of these cases, increasing k corresponds to shifting the (log-log) error probability curve to the left (improved performance). Similarly, decreasing k shifts the curve to the right (degraded performance). Such a shift to the right could correspond to hardware implementation losses of various types.

GAUSSIAN NOISE WITH FLAT FADING

Next consider the case of slow flat fading. The fading is termed flat because the entire signal bandwidth fades together. It is termed slow because fading effects are assumed constant over a symbol duration (or two symbols for differential phase-shift keying). In this case, the following expressions are well known (Schwartz, et al., 1966). We have

$$P_E \approx \frac{1}{2(1+k\bar{R})} \tag{5}$$

for noncoherent FSK and binary DPSK, and

$$P_E \approx \frac{1}{2}\left[1 - \frac{1}{(1 + 1/k\bar{R})^{1/2}}\right] \tag{6}$$

for coherent FSK, coherent PSK, and quarternary DPSK. In these formulas, $\bar{R}$ is the average received value of E_b/N_0 ; $\bar{R} = \bar{E}_b/N_0$.

Kam (1991) points out that this traditional analysis applies to the case of zero-fading bandwidth. That is, the fading process results in symbol-to-symbol channel complex gain changes, but is slow enough to be treated as constant over the symbol duration. For DPSK, this, in fact, means slow enough to be treated as constant over symbol pairs. Kam (1991) extends the existing analysis to 2, 4, and 8 DPSK with arbitrary order of diversity. He treats the case where the fading process is constant over the symbol duration but not across symbol pairs. The results of particular interest to us are the case of no diversity with binary or quarternary DPSK.

For binary DPSK,

$$P_b = \frac{1}{2}\left[\frac{1 + (1-\rho)R}{(1 + R)}\right], \tag{7}$$

where ρ is the normalized autocorrelation function of the fading process. For a Gaussian fading process, ρ is given by

$$\rho = \exp[-(\pi BT)^2], \tag{8}$$

where B is the fading bandwidth ($B = \sqrt{2\sigma^2}$ for a Gaussian process with variance σ^2) and T is the modulation symbol duration. When $\rho = 1$ ($B = 0$) this expression reduces to the expression for P_E of DPSK given in equation 5.

For quaternary DPSK, we have

$$P_b = \frac{1}{2}\left\{1 - \frac{1}{\left(\frac{2}{\rho^2}\left(1+\frac{1}{R}\right)^2 - 1\right)^{1/2}}\right\}. \tag{9}$$

This result applies to the slow fading channel, for which $\rho = 1 - \epsilon$, with $0 \le \epsilon << 1$. We can rewrite this expression as

$$P_b = \frac{1}{2}\left\{1 - \frac{1}{\left[1 + 4\epsilon + 6\epsilon^2 \cdots + \frac{2}{\rho^2 R}\left(2 + \frac{1}{R}\right)\right]^{1/2}}\right\}. \tag{10}$$

We can now find the irreducible error bound as the limit of this expression as $R \rightarrow \infty$. The result is

$$P_b \approx \frac{1}{2}\left\{1 - \frac{1}{[1 + 4\epsilon + 6\epsilon^2 \cdots]^{1/2}}\right\} \approx \frac{1}{2}\{1 - (1 - 2\epsilon)\} = \epsilon . \tag{11}$$

We can also evaluate the expression for P_b in the case of zero-fading bandwidth ($\rho = 1$).

$$P_b = \frac{1}{2}\left\{1 - \frac{1}{\left[1 + \frac{2}{R}\left(2 + \frac{1}{R}\right)\right]^{1/2}}\right\} . \tag{12}$$

This expression can be compared to the expression given in equation 6. In particular, for R >> 1/2, equation 12 becomes

$$P_b \approx \frac{1}{2}\left\{1 - \frac{1}{\left[1 + \frac{4}{R}\right]^{1/2}}\right\} \tag{13}$$

which corresponds to equation 6 with k = 1/4. This value of k is quite close to the Arthurs and Dym (1962) value of

$$k = \sin^2\left(\frac{\pi}{4\sqrt{2}}\right) \approx 0.278 . \tag{14}$$

EXTENSIONS FOR DQPSK

At HF, transmitted signals may be refracted from various physical locations in the ionosphere. This results in time dispersion of the received signal. This time dispersion is termed multipath. It produces the phenomenon of frequency selective fading.

Garber and Pursley (1988) have analyzed the case of frequency selective fading for binary DPSK under the particular circumstances where most (>90 percent) of the delayed energy affects only the next symbol, where the dispersion takes on one of four continuous-time distributions, and for three types of DPSK waveforms. Garber and Pursley (1988) show the bit-error performance under these conditions is bounded below by the performance for slow flat fading for low SNR. At high SNR there is an irreducible error bound due to the intersymbol interference. The principal factors influencing P_I (the irreducible bound) are (a) the normalized multipath spread, and (b) the shape of the communication signal pulse or symbol. There is little performance difference among Gaussian, exponential, triangular, or rectangular continuous-delay distributions. No analysis has been identified indicating the effects of discrete-delay distributions or of multiple, disjointed continuous distributions. The following discussion summarizes the Garber and Pursley (1988) results.

Given the normalized multipath spread μ_ν, we can estimate P_I. Bounds for this estimate are plotted in Garber and Pursley's (1988) figure 6. A straight line approximation to this figure (based on their table IV) is

$$P_I = 0.475\mu_\nu^2. \tag{15}$$

Having found P_I, we can find an expression for P_b. This expression defines a curve that transitions between the P_b of the slow flat fading case at low SNR and the constant value P_I at high SNR. Garber and Pursley (1988) do not give a closed-form expression for this performance curve. However, the performance curve is similar to that given by Kam (1991) for the nonselective fading case. We will proceed to construct a model *for empirical use* of the form derived by Kam (1991). This is **not** a mathematical derivation.

Let $g(\xi)$ be the power density spectrum of the channel delay. Then the rms delay is

$$M = \left[\int_{-\infty}^{\infty} \xi^2 g(\xi) d\xi \right]^{1/2} \tag{16}$$

and the rms-multipath spread is defined as

$$\mu = M/T , \tag{17}$$

where T is the communications pulse duration. The normalized-multipath spread μ_ν (that is, normalized to an "effective" communications pulse duration) depends upon the shape of the transmitted communications signal pulses. For rectangular DPSK pulses, $\mu_r = \mu = M/T$. For sinusoid-shaped DPSK pulses, μ_s can be found from the power density spectrum of the channel delay. An approximation to the tabulated values given by Garber and Pursley (1988) is

$$\mu_s \approx 15.1(M/T)^{3.09} . \tag{18}$$

In total, a three-step process is needed to estimate P_I. First, the channel-rms delay must be determined. From this the appropriate normalized-multipath spread must be estimated, incorporating both the shape and duration of the transmitter symbol modulation. Finally, the estimated P_I is calculated using the normalized-multipath spread.

We can now consider modifications to the expression for P_b in the flat-fading case to include multipath frequency-selective fading effects. In particular, multipath results in an irreducible error rate similar to that observed for fading. If in the expression for P_b we insert a term ζ multiplying ρ, and similarly constrain ζ to the range $0 \le \zeta \le 1$, then ζ will have the same effect as ρ. That is, ζ will introduce an irreducible error bound as $R \rightarrow \infty$.

The resulting expression for binary DPSK is

$$P_b = \frac{1}{2} \left\{ \frac{1 + (1 - \rho\zeta)R}{1 + R} \right\} . \tag{19}$$

Now for $R \rightarrow \infty$ and $\rho = 1$ (zero fading bandwidth on individual paths) we have

$$P_b = \frac{1}{2}(1 - \zeta) = P_I . \quad (20)$$

From this expression we find the appropriate value for ζ to be $\zeta = 1 - 2P_I$.

Without justification, but with analogy to the flat-fading case, we model quaternary DPSK performance as

$$P_b = \frac{1}{2}\left\{1 - \frac{1}{\left(\frac{2}{(\zeta\rho)^2}\left(1 + \frac{1}{R}\right)^2 - 1\right)^{1/2}}\right\} \quad (21)$$

Let the multipath have Gaussian distribution with density

$$g(\xi) = \frac{1}{\sigma\sqrt{2\pi}} \exp[-\xi^2 / 2\sigma^2] . \quad (22)$$

The "rms-multipath delay" is just the standard deviation of $g(\xi)$. The symbol M is usually used to denote the rms-multipath delay, so in the Gaussian case $M = \sigma$. The rms-multipath spread measures the time spread of the received signal relative to the symbol duration in "σ-units." These σ-units are multiples of the standard deviation of the multipath spread treated as a density function. For the Gaussian density given above, we then know that: for $\mu = 1$, about 68 percent of the spread signal is contained within ± 1 symbol; for $\mu = 1/2$, about 95 percent of the spread is contained in ± 1 symbol; and, for $\mu = 1/3$, about 99.74 percent of the spread is contained in ± 1 symbol.

EXTENSIONS FOR FSK

For FSK signaling, we can apply another analysis of Garber and Pursley (1989). In general, the multipath interference results in an irreducible error bound. This bound is a bit-error probability that cannot be improved by increasing the signal-to-noise ratio. At high SNR, the bit-error rate is a constant value, P_I. At low SNR, the bit-error rate is dominated by the error-rate curve for nonselective Rayleigh fading. A smooth transition occurs between these regions.

An estimate for the value of P_I can be obtained using the Garber and Pursley (1989) results. This involves estimating the normalized rms-multipath spread

$$\mu^2(h) = 2\int_0^1 g'(x)x^2 [1 + \mathrm{sinc}^2(hx) - 2\cos(\pi hx)\,\mathrm{sinc}(hx)]\,dx , \quad (23)$$

where, for a power density spectrum of the Gaussian delay of the fading process, we have

$$g'(x) = \frac{1}{(M/T)\sqrt{2\pi}} \exp[-x^2/2(M/T)^2] . \tag{24}$$

In the expression for $\mu^2(h)$, the sinc function is just $\mathrm{sinc}(y) = \sin(\pi y)/\pi y$. We see that $\mu(h)$ depends upon the FSK modulation index $h = 2 f_d T$, where T is the baud duration and $2 f_d$ is the tone spacing, and upon the rms-multipath spread.

This result applies to phase-continuous FSK. Phase discontinuities result in irreducible error bounds that are orders of magnitude higher. Since the ALE standard modem is phase continuous, we will not consider discontinuous phase FSK.

The integral may be evaluated numerically. For the Gaussian delay spectrum case, Garber and Pursley (1989) have tabulated values of $\mu(h)$ for h = 1, 2, and 10. This is a rather complicated integral to evaluate since the integrand is close to zero and involves differences of values that are close to 1. However, we can approximate the Garber and Pursley (1989) tabulated data with the expression

$$\mu(h) \approx \min[2.6h(M/T)^{1.86},\ M/T] . \tag{25}$$

This expression is plotted in figure 1 against M/T, with h as a parameter. The tabulated values from Garber and Pursley (1989) are also shown on the figure. In this approximation, the exponent may be changed from 1.86 to 2 (and the multiplier increased from 2.6 to 5.4) with some decrease in accuracy (errors as large as 50 percent). This only makes sense where the simpler expression is required for ease in computation.

Given $\mu(h)$, Garber and Pursley (1989) give bounds on P_I in their figure 4. We can approximate P_I by

$$P_I \approx 0.29\mu^2(h) . \tag{26}$$

The functional form of the bit-error rate curve for FSK should be similar to that given by Kam (1991) for DPSK, since the flat-fading performance form for FSK is similar to that for DPSK. (For flat fading, the form for FSK differs from DPSK by a factor of 2 dividing the average SNR.) We then can model the bit-error rate performance as

$$P_b = \frac{1}{2}\left[\frac{1 + (1-\rho)\bar{R}/2}{1 + \bar{R}/2}\right]. \tag{27}$$

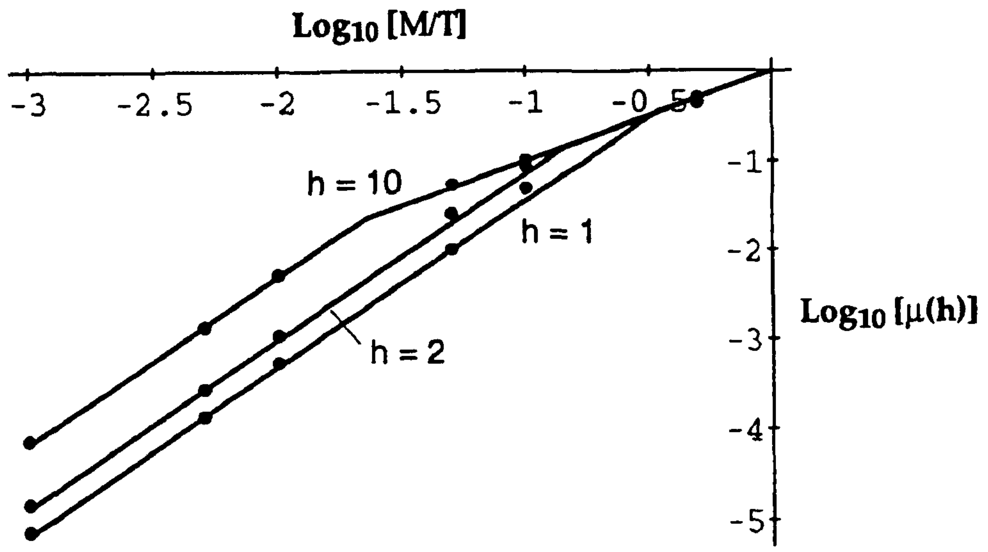

Figure 1. Normalized rms-multipath spread approximation.

If $\rho = 1 - \epsilon$, we have

$$\lim_{R \to \infty} [P_b] = \epsilon/2 = P_I . \quad (28)$$

Using the approximations based on Garber and Pursley (1989), we have

$$\epsilon = 0.58 \min^2[2.6h(M/T)^{1.86}, M/T] . \quad (29)$$

We note, for small ρ (large ϵ), P_b is close to 1/2 for all $\bar{R}$. This does not correspond to a useful channel. If we now consider the crossover point between the two expressions in equation 25, we can solve to find the crossover to be

$$M/T = \frac{1}{(2.6h)^{1.16}} . \quad (30)$$

The largest possible M/T crossover occurs for h = 2 in the ALE case, which implies that $\mu(h) \approx M/T$ (independent of h) whenever

$$M/T \geq \frac{1}{5.2^{1.16}} \approx 0.15 . \quad (31)$$

We can extend the binary FSK analysis to 8-ary ALE FSK. Let the eight frequencies used by ALE be labeled f_1 through f_8, as shown in table 1. Next, assume the

Garber and Pursley (1989) analysis holds for any pair of these frequencies, f_i and f_j. In particular, if f_i is transmitted, then the probability that the receiver will detect f_j will follow the functional form for binary FSK. However, since there are eight tones rather than two, the coefficient should be 1/8. Then in the limiting case $\bar{R} \rightarrow 0$, all tones become equally likely at the receiver output, and P(f_j detected | f_i transmitted) $\rightarrow$1/8. We can write

$$P(f_j \text{ detected} \mid f_i \text{ transmitted}) \approx \frac{1}{8}\left[\frac{1 + (1 - \rho)3\bar{R}/2}{1 + 3\bar{R}/2}\right]. \tag{32}$$

Table 1. ALE frequencies and bit representations.

Frequency	Hz	Bit Pattern
f_1	750	000
f_2	1000	001
f_3	1250	011
f_4	1500	010
f_5	1750	110
f_6	2000	111
f_7	2250	101
f_8	2500	100

Here we assume that $\bar{R} = \sigma^2 E_b/N_0$ is the average received energy per bit divided by the noise-spectral density, and $3\bar{R}$ is the average received energy per symbol (tri-bit) divided by the noise-spectral density. We also have

$$\rho = 1 - 0.58 \min^2[2.6h(M/T)^{1.86}, M/T] \tag{33}$$

and $h = 2f_d T = |f_i - f_j| T$. Since for ALE we have T = 1/125 second, we can write $h = 2|i - j|$.

The number of bit errors given an incorrect frequency decision at the receiver depends on the specific values of i and j. Table 2 shows the number of bit errors as a function of $|i - j|$. If we assume all bit patterns are equally likely at the source, we can write the received bit-error probability as

$$P_b = \frac{1}{8} \sum_{i=1}^{8} \sum_{j \neq i} P(f_j \text{ detected} \mid f_i \text{ transmitted}) \frac{N_{ij}}{3} .$$

Here N_{ij} is the number of bits of the tri-bit in error when f_j is detected, given that f_i is transmitted.

Table 2. Possible error occurrences.

| i - j |	bit errors
1	1
2	2
3	3 single, 2 triple
4	2
5	1 single, 2 triple
6	2
7	1

For small values of ρ (ϵ close to 1), meaning for large M/T, ρ is independent of h and hence independent of i and j. This independence occurs for M/T satisfying $(M/T)^{0.86} \geq 1/(2.6h)$. For ALE, ρ is independent of i and j whenever $M/T \geq 0.15$, since the minimum possible h is 2. This occurs whenever $\rho \leq 0.987$. In this case we have

$$\begin{aligned} P_b &= \frac{P(f_J \text{ detected} \mid f_i \text{ transmitted})}{8} \sum_{i=1}^{8} \sum_{j \neq i} \frac{N_{ij}}{3} \\ &= \frac{1}{64}\left[\frac{1 + (1-\rho)3\overline{R}/2}{1 + 3\overline{R}/2}\right] \sum_{i=1}^{8} \sum_{j \neq i} \frac{N_{ij}}{3} \\ &= \frac{1}{2}\left[\frac{1 + (1-\rho)3\overline{R}/2}{1 + 3\overline{R}/2}\right] \\ &= \frac{1}{2}\left[\frac{1 + 0.58(M/T)^2 3\overline{R}/2}{1 + 3\overline{R}/2}\right]. \end{aligned} \tag{34}$$

In the case of $\overline{R} \rightarrow \infty$, the irreducible error bound is given by

$$P_I = 0.29(M/T)^2 , \tag{35}$$

which applies whenever $M/T \geq 0.15$. The value of P_I for $M/T = 0.15$ is $P_I = 0.0065$.

We can also find the bit-error probability expression for large values of ρ (ϵ close to zero). This corresponds to $(M/T)^{0.86} \leq 1/(2.6h)$. For ALE the maximum value of h is 14, so "large ρ" corresponds to $M/T \leq 0.015$. In this case we have

$$P_b = \frac{1}{64} \sum_{i=1}^{8} \sum_{j \neq i} \left[\frac{1 + (1 - \rho)3\overline{R}/2}{1 + 3\overline{R}/2} \right] \frac{N_{ij}}{3}$$

$$= \frac{1}{64(1 + 3\overline{R}/2)} \left[32 + \frac{\overline{R}}{2} \sum_{i=1}^{8} \sum_{j \neq i} \epsilon_{ij} N_{ij} \right]$$

$$= \frac{1}{64(1 + 3\overline{R}/2)} \left[32 + 7.84\overline{R}(M/T)^{3.72} \sum_{i=1}^{8} \sum_{j \neq i} |i - j|^2 N_{ij} \right] \tag{36}$$

$$= \frac{1}{2(1 + 3\overline{R}/2)} [1 + 310\overline{R}(M/T)^{3.72}] .$$

Here we have explicitly shown the dependence of ϵ on h, and hence on i and j, by writing ϵ_{ij}. As $\overline{R} \rightarrow \infty$, the corresponding irreducible error bound is

$$P_I = 103(M/T)^{3.72} . \tag{37}$$

The value of P_I for $M/T = 0.015$ is $P_I = 1.7 \times 10^{-5}$.

The use of the "large ϵ" approximation provides an upper bound to P_I whenever ϵ is not "large" (smaller M/T). In particular, when $M/T = 0.015$ the "large ϵ" approximation gives $P_I = 6.5 \times 10^{-5}$, about four times the value determined from the "small ϵ" approximation. Since even this larger error rate is quite small for practical HF systems, we may safely use the "large ϵ" approximation in all cases.

ANALYTIC ASSESSMENT CONCLUSIONS

Suppose an HF channel has delay spread but no Doppler spread. We can then use the FSK and DQPSK expressions and compare modem performance for this channel. Figures 2 and 3 show the bit-error performance for DQPSK and FSK respectively. We assumed 2400 bit/second data rate via a 16-tone modem with no error correction coding for the DQPSK case of figure 2. In figure 3 we used the parameters of an ALE modem in estimating FSK performance. The analysis leading to figure 2 applies only for multipath confined to one symbol, or $M < 125$ ($\log M < 2.09$). Figure 3, therefore, shows an error ($P_b > 1/2$) for large M. The true bit-error probability approaches (does not exceed) 1/2 for large M.

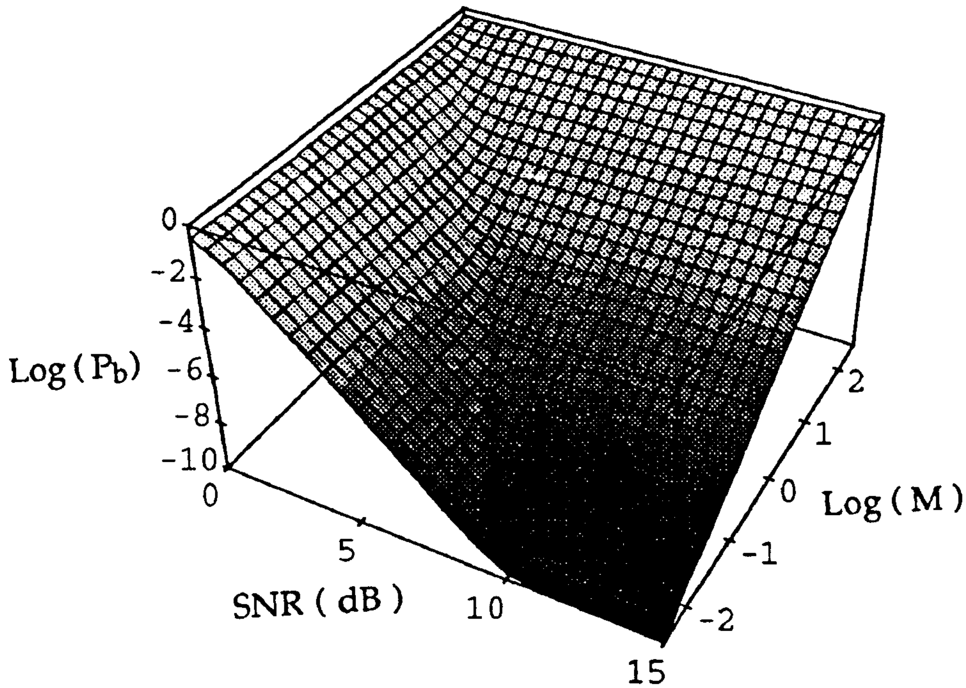

Figure 2. Bit-error probability for 16-tone DQPSK.

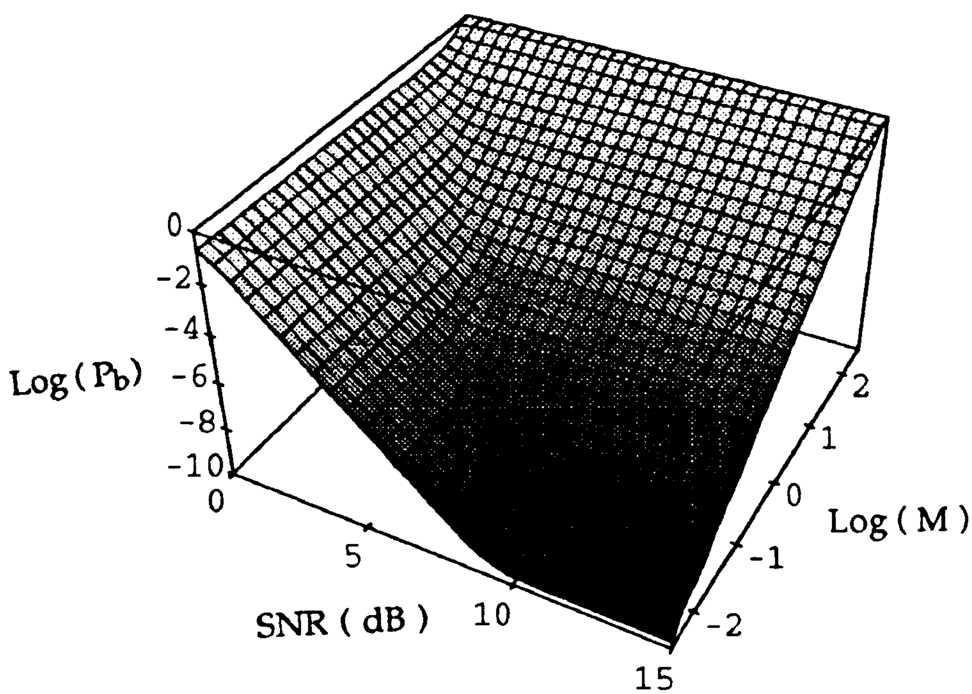

Figure 3. Bit-error probability for 8-ary FSK.

The first point to note about these figures is that they are very similar. This similarity implies that performance of the ALE modem should be a useful guide for projecting DQPSK modem performance on this type of channel.

The second point to note is that modem performance is strongly dependent upon the rms-multipath spread, M. SNR alone is not a reliable predictor of bit-error performance. However, since SNR and P_b together may be used to estimate M for this channel, SNR could be used with either P_b or M to select a best frequency. In practice, this is not the case since an accurate estimate of P_b is difficult to obtain, especially since only a relatively few bits are exchanged in the ALE algorithm. Furthermore, Doppler spread may also significantly affect modem performance—then all three parameters (SNR, Doppler spread, and delay spread) are needed, and SNR plus bit-error rate are insufficient for predicting performance.

The DQPSK analysis shows Doppler spread and delay spread are both significant parameters for predicting modem performance. The implication of the analysis is a technique for measuring delay and Doppler spreads should be incorporated into ALE handshaking, and these values together with SNR should be used in the frequency-selection algorithm. The frequency-selection algorithm would then take modem characteristics into account when prioritizing the available frequencies.

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13. ABSTRACT *(Maximum 200 words)*

The objective is to develop a simple model for HF (2–30 MHz) modem performance as a function of various channel parameters, and to develop techniques for predicting those parameters.

Modem performance was modeled based on three parameters—signal-to-noise ratio (SNR), delay spread, and Doppler spread. Each of these parameters is measurable using known techniques. The model demonstrates the inadequacy of using only SNR and bit error-rate estimates for frequency selection in automatic link establishment (ALE) algorithms.

The benefits from incorporating delay and Doppler spreads in the ALE frequency selection algorithm are sufficiently clear to justify extension of the algorithms to include these factors. Further research is required to verify the sufficiency of SNR, delay spread, and Doppler spread for frequency selection.

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