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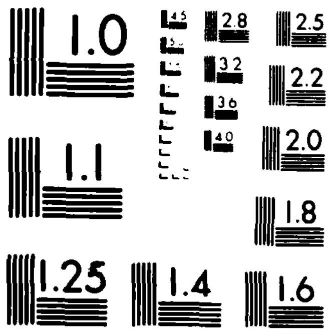

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SPREAD SPECTRUM MOBILE RADIO COMMUNICATIONS

under contract

AFOSR-82-0309

with

Air Force Office of Scientific Research
Bolling Air Force Base
Washington, D.C.

Electrical Engineering Department

School of Engineering and Applied Science

Southern Methodist University

Dallas, Texas 75275

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INTERIM SCIENTIFIC REPORT

September 1, 1982 -

August 31, 1984

S.C GUPTA, PRINCIPAL INVESTIGATOR

Electrical Engineering Department

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In this report, we present some additional results on the performance of Frequency Hopped Multilevel FSK (FH-MFSK) systems under mobile environment. This is followed by a review on packet radio networks and some results on the performance of a packet transmission technique in mobile radio channels.

Simulation studies confirm the theoretical performance estimate of a FH-MFSK hard-limited receiver operating under adjacent cell interference conditions. Both simulation and approximate theoretical considerations testify the usefulness

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of a nonparametric receiver based on rank-sums. This receiver also possesses a robustness against changing probability models. It is believed that this receiver is a competing alternative to the parametric receivers, such as the hard-limited receiver.

We reviewed the signal processing aspect of packet radio in computer communication networks. The feasibility of spread-spectrum technique in PRNET is discussed, with an emphasis on the repeater portion. The alternative implementation technologies, viz Digital, SAW and CCD, required for the processing needs are compared. The performance of a protocol for the mobile PRNET for computer communications over degraded channels is discussed. The analytical results show that the packet error rate and message delay performance degrade rapidly with vehicle speed up to about 80 mph, but above this speed the degradation is slow for reasonably good SNR.

Summary

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Introduction

This report summarizes new results on the performance of FH-MFSK systems in mobile environment. Transmission of data in mobile radio channel by means of packets is also explored. In this chapter, we briefly review FH-MFSK system. Subsequent chapters present our analysis and the results obtained from them.

1.1 Frequency-Hopping Multi-level Frequency-Shift Keyed System (FH-MFSK) [1]

A block diagram of the m^{th} transmitter of FH-MFSK system is shown in figure 1-1. Figure 1-2 shows the block diagram of the receiver. The operation of the system can be understood by referring to these figures. Every T seconds K message bits are loaded serially in a buffer and transferred out as a K-bit word X_m. Assuming the modulo-2^K adder does nothing, for the moment, X_m will select one of the 2^K possible different frequencies from the tone generator. At the receiver, the spectrum of each T second transmission is analyzed to determine which frequency, and hence, which K-bit word, X_n is sent. Of course, the system as such is useless for multiple-user operation. If a second transmitter were to generate X_n, neither the receiver m nor the receiver n would know whether to detect X_n or X_m. To avoid this, we add the address generator as shown in figure 1-1 and assign a unique address to each user.

The basic interval T is divided into L intervals of duration τ each. Over T seconds, the address generator of m^{th} user generates a sequence of L numbers:

$$a_m = (a_{m1}, a_{m2}, \ldots a_{mL}) \tag{1-1}$$

$$\text{Each } a_{mi} \in \{0, 1, 2, \ldots 2^K-1\} \tag{1-2}$$

Here, each a_{mi} is selected at random from the set specified by 1-2 [called 'random address assignment'].

Now, each $a_{m\ell}$ is added modulo-2^K to X_m to produce a new K-bit number

$$Y_{m,\ell} = X_m + a_{m\ell}$$

or

$$\left.\begin{aligned} &\underline{Y} = (Y_{m1}, Y_{m2}, \dots Y_{mL} \\ &\underline{X_m} = (X_m, X_m, \dots X_m) \\ &\underline{Y} = \underline{X_m} + \underline{a_m} \end{aligned}\right\} \qquad 1\text{-}3$$

Each τ seconds, $Y_{m\ell}$ selects the corresponding transmitter frequency. At the receiver, demodulation and modulo 2^K subtraction by the same number $a_{m\ell}$ are performed every τ seconds, yielding

$$Z_{m\ell} = Y_{m\ell} - a_{m\ell} = X_m$$

The sequence of operations is illustrated by the matrices of figures 1-3a and 1-3b. Each matrix is either a sequence of K-bit numbers (code word, address, detection matrix) or a frequency-time spectrogram (transmit spectrum, receive spectrum). The matrices pertain to one link in a multi-user system. Crosses show numbers and frequencies generated in that link. Circles show the contributions of another link. As said earlier, the transmit spectrum is generated by modulating the address with code word using modulo-2^K addition. Equivalently, when each entry in the address matrix is shifted cyclically by the row number specified by the code word matrix, we get the transmit spectrum 1-3a

Because of multi-users, extraneous entries are created in the detection matrix. For example, a word X_n transmitted over the n^{th} link will be decoded by the receiver m as

$$Z'_{m\ell} = X_n + a_{n\ell} - a_{m\ell}.$$

The $Z'_{m\ell}$ are scattered over different rows. The desired transmission, on the other hand, is readily identified because it produces a complete row of entries in the detection matrix. Normally, the fading of the tones and the receiver noise can cause a tone to be detected when none has been transmitted (false alarm) and/or can cause a transmitted tone to be not detected (miss). Even without these impairments, many user entries can combine to produce a complete row other than X_m and hence, can cause errors in the identification of correct row (in the word X_m). Hence, a majority logic rule is attempted: choose the code word associated with the row containing the greatest number of entries. Under this decision rule, an error will occur when insertions (detected tones due to other users and false alarms) combine to form a row with more entries than the row corresponding to the transmitted code word. An error can occur when insertions combine to form a row containing the same number of entries as the row corresponding to the transmitted code word. We view the transmission to each square in the tone detection matrix as an example of non-coherent on-off keying. Because of fading of amplitude and the random change of phase, it is not possible to employ coherent detection in mobile environment. (Recall that we used differential phase detection in FH-DPSK scheme, since the phase is not likely to change significantly from bit to bit).

From the text book formulas we have

$$P_F = \exp(-\beta^2/2) \qquad 1\text{-}4$$

$$P_D = 1 - \exp(-\beta^2/2(1+\bar{\rho})) \qquad 1\text{-}5$$

where P_F denotes false alarm probability, P_D the probability of deletion (miss), β the normalized threshold set in the receiver and $\bar{\rho}$ the average signal to noise ratio. The above scheme, where the presence or absence of energy in each square of detection matrix is decided, together with majority logic decision is called "Hard-limited Combining".

1.2 Receiver Structure for FH-MFSK Modulation

Figure 1-4 shows a section of the non-coherent envelope analyzer. As in [1], let τ be the chip duration, K be the number of bits of information transmitted every $L\tau$ seconds, $W = 20$ MHz be the one way bandwidth and R be the bit rate. Then, we have 2^K such sections in operation corresponding to different orthogonal tones. Let ε_{ij} denote the envelope squared output at the i^{th} envelope analyzer after the j^{th} chip. Corresponding to either signal plus noise or noise only case, we have ε_{ij} to be either exponentially distributed with mean value $(1/\lambda_1)$ or exponentially distributed with mean value $(1/\lambda_0)$, respectively.

A mobile user u receives the signals from the base and creates a decoded matrix every $L\tau$ seconds. The values ε_{kj} become the entries X_{ij} in the decoded matrix (the decoding is done on the received matrix with the address of user u). In general, a receiver chooses a row as the row corresponding to the transmitted word, based on some decision criterion. In [1], where hard-limited combining is employed, corresponding to each entry (i,j) in the matrix, a number n_{ij} is assigned such that

$$n_{ij} = \begin{cases} 1 & \text{iff } X_{ij} \geq T \\ 0 & \text{otherwise} \end{cases}$$

A row k is declared as the correct row if

$$\sum_{j=1}^{L} n_{kj} > \sum_{j=1}^{L} n_{ij} \qquad i \neq k$$

In case two or more rows have the same maximum sum, $\sum$, then any row among these rows is chosen at random as the correct row. In [2] , a linear combiner based on choosing the k^{th} row as the correct row such that

$\max_i \left\{ \sum_{j=1}^{L} X_{ij} \right\} = \sum_{j=1}^{L} X_{kj}$ was analyzed, for mobile to base transmission, using some approximate techniques.

1.3 <u>Likelihood Receiver</u>: We shall assume that the minimum frequency spacing between the hops in the transmitted waves is larger than the coherent bandwidth of the Rayleigh fading channel. This, then, implies that X_{ij} are independent and exponentially distributed. Among the 2^K rows in the decoded matrix, only one row is the correct row, wherein all the X_{ij}'s have a mean value $(1/\lambda_1)$. In each of the rest of the (2^K-1) spurious rows, some elements have a mean value of $(1/\lambda_0)$ and the rest have a mean value of $(1/\lambda_1)$. A spurious row has contributions partly from the interfering users plus noise and partly from the receiver noise. On an average, each spurious row will have a proportion, p, of X_{ij}'s created due to interference, where p is given by

$$p = 1 - (1 - 2^{-K})^{M-1} \qquad \text{1-6}$$

and M equals the number of users operating in the cell.

Since each row can be a spurious row (hypothesis H_0) or not (hypothesis H_1), we have the following testing problem applied to a ℓ^{th} row:

$$H_0: \quad X_{\ell j} \sim p\, \lambda_1\, e^{-\lambda_1 x} + (1-p)\, \lambda_0\, e^{-\lambda_0 x}$$

(vs)

$$H_1: \quad X_{\ell j} \sim \lambda_1\, e^{-\lambda_1 x} \qquad 1\text{-}7$$

where, $j = 1, 2, \ldots L$

$\ell = 1, \ldots 2^K$

It can be noticed that the proportion, p, is known once the number of users operating in the cell is known.

Normalizing $X_{\ell j}$'s with respect to the received signal, we have

$$Y_{\ell j} = \lambda_1\, X_{\ell j} \qquad 1\text{-}8$$

Therefore, 1-7 gets modified as

$$H_0: \quad Y_{\ell j} \sim pe^{-y} + (1-p)\, be^{-by}$$

vs.

$$H_1: \quad Y_{\ell j} \sim e^{-y} \qquad 1\text{-}9$$

Where $b = \frac{\lambda_0}{\lambda_1}$, signal plus noise to noise power ratio (SNNR)

Forming the likelihood ratio, we have

$$S_\ell = -\sum_{j=1}^{L} \ln(p + (1-p)\, be^{-(b-1)y_{\ell j}}) \qquad 1\text{-}10$$

Then, the likelihood receiver chooses the row having $\underset{\ell}{\mathrm{Max}}\{S_\ell\}$ as the correct row.

In the next chapter, the results on the performance of various receivers in a mobile station, obtained by simulation study are presented. In chapter III some nonparametric receivers are analyzed and found to be useful for mobile radio applications. Chapters IV and V discuss the application of packet radio techniques in mobile communication.

II. Simulation Study of Various Receivers

Here, we present a simulation analysis which is based on some analytical methods employed earlier [2], to arrive at a histogram estimate of the distribution of the samples at the output of a non-coherent envelope detector, for detecting the FH-MFSK tones. For simplicity sake, we consider only a 3-cell system and assume a perfect synchronization in the arrival of tones at the mobiles. Once the estimate of the distribution of the samples is known, it is possible to arrive at a performance estimate of any detector which acts on the basis of these samples. A simple study on the performance of a hard-limited receiver shows a good agreement with the fully analytical result obtained earlier. This also testifies the validity of neglecting the non-orthogonal interference, as done in the analytical study. Also, the performance of some non-parametric receivers are tested using the simulated samples.

In section 2.1 we present an algorithm to generate the samples of the envelope detector. In section 2.2 we analyze the performance of a hard-limited receiver. Whereas in section 2.2, results on the U detector is presented. In section 2.4 we evaluate the performance of a Maximum Rank Sum Receiver (MRSR). We borrow some of the notations and the concepts used in [2].

2.1 Algorithm to Generate the Samples of Envelope Detector Output

Consider a 3-cell system and a user u moving along the line AB as shown in figure 2-1. The algorithm presented here generates the samples of the elements in the correct and the spurious rows of the decoded matrix of user u. Instead of considering the envelope detector output, we consider the envelope squared output (or the energy) in each of the elements.

A. Spurious Row Samples

It can be noticed that the spurious row elements are due to three causes:

(i) Interference from the other users in the same cell (cell 1)

(ii) Interference from the users in cell 2 (and cell 3)

(iii) Noise at the receiver input.

Actually, the case (ii) has two contributions; one due to the 'legitimate' interference caused by a user in 2 transmitting tones in a specific slot and the other due to the non-orthogonal interference of adjacent tones entering the output of the tone detector under investigation. Whereas the second contribution was neglected in the analytical study [2], we shall include the significant portion of its effect in this simultation. The interference due to (i) can be included by considering the average interference power IN_0 due to users in the same cell.

By referring to figure 1-4, we can see that the sample y_1 is a Gaussian random variable with zero mean and variance as determined by contribution from various interfering sources mentioned earlier. Since each source is independent of others, the variances add together. Hence, x is exponentially distributed with a mean value which depends on the variance of y_1 (or y_2).

In order to evaluate (ii) consider i^{th} tone detector at n^{th} time slot. It can be established [2] that the variance of the in-phase (or quadrature phase) component at the envelope detector due to a j^{th} tone at the input is given by

$$\text{Variance} = \frac{IN}{\pi^2 (i-j)^2} \sin^2 (\pi(i-j)\,a) \qquad 2\text{-}1$$

Here $a\tau$, a real quantity, accounts for the difference in the propagation delay difference (in mod τ) between the arrival of tones from base 2 (or 3) and base 1 at receiver u. Since we are only interested in an average performance estimate, a can be treated as a random variable with uniform density over (0, 1). As can be seen from (1), only the tones adjacent to i will have significant contribution. When j equals i, we have the case of 'legitimate interference', mentioned earlier. It is easy to see that the significant contribution occurs only from the situations corresponding to j = (i+1): i and (i-1).

In the following table 2-1, we present various possible configurations and the corresponding probability of occurence and the corresponding variance.

Probability	Configuration					Variance
	Rest	(i-1)	i^{th}	(i+1)	Rest	(Approximate)
P_3	--	x	x	x	--	$2V_1+V_0$
P_2	--	A	x	x	--	V_1+V_0
P_2	--	x	x	A	--	V_1+V_0
P_2	--	x	A	x	--	$2V_1$
P_1	--	A	A	x	--	V_1
P_1	--	x	A	A	--	V_1
P_1	--	A	x	A	--	V_0
P_0	--	A	A	A	--	0

-- Don't care x Present A absent

TABLE 2-1

It is easy to observe that the probabilities are given by

$$P_0 = (1 - 3/2^K)^M \tag{2-2}$$

$$P_1 = (1 - 2/2^K)^M - P_0 \tag{2-3}$$

$$P_2 = (1 - 1/2^K)^M - P_0 - 2P_1 \tag{2-4}$$

$$P_3 = 1 - P_0 - 3P_2 - 3P_1 \tag{2-5}$$

Since the non-orthogonal interference spills over two adjacent slots, on an average, the non-orthogonal interference variance is given by equation 2-6 and not by equation 2-1:

For (i-j) = 1, Average interference variance

$$= \frac{IN}{\pi^2} (\sin^2 \pi a + \sin^2 \pi(1-a))$$

$$= \frac{2IN}{\pi^2} \sin^2(\pi a) \tag{2-6}$$

where IN is the variance of the interference due to i^{th} tone from base 2 (or 3) had the tone been completely synchronized with the receiver u. Therefore,

$$V_1 = \frac{2IN}{\pi^2} \sin^2 (\pi a) \tag{2-7}$$

$$V_0 = IN\ (a^2 + (1-a)^2) \tag{2-8}$$

We observe that $E(V_0)$ is only 2/3 IN and hence the approximation, assumed in [2], that this value equals IN is in error. However, by over estimating $E(V_0)$, we could assume that some portion of the non-orthogonal interference effect is included, in an indirect way. Then by normalizing the envelope squared x, with noise variance, we can generate the normalized samples Z. The flow diagram in figure 2-2 illustrates the generation of these samples.

B. <u>Correct Row Samples</u>:

The procedure is very similar to the one described in (A) except now the variance of y_1 (or y_2) has contributions from the following:

(i) Interference from other users in cell 2 (and 3)

(ii) Intended signal power from base 1

(iii) Noise at the receiver input.

Figure 2-2 also illustrates the generation of the correct row samples.

2.2 The Hard-Limited Receiver Performance

For an assumed position of user u (k assumed to be known) and a controlled power of 25 dB (S = 2.5), we arrive at the histogram of the samples. Then, it is possible to calculate the average probability of bit error P_b of a hard limited receiver, once we know the threshold employed in the receiver. Using the optimum threshold value of β^2 [2], we could find out (i) probability of an entry in the spurious row of the decoded matrix of user u and (ii) the probability of an entry in the correct row. Therefore, P_b can be computed as explained in [2]. Table II shows P_b values against k values. By comparing this table with figure 4 of [2], we observe the close agreement between the two. This also testifies that, on an average, the non-orthogonal interference does not have any noticeable effect on the performance.

k	β^2	P_b
0.3	49	1.05×10^{-6}
0.5	49	1.2×10^{-4}
0.7	100	6.2×10^{-5}

M = 100, S = 2.5

TABLE 2-2

2.3 Mann-Whitney U Detector Performance

A Mann-Whitney U test (or the equivalent Wilcoxon rank-sum test) is a useful two sample test for the testing whether among any two samples, one sample is stochastically larger than the other [3]. Also, such a test belongs to the class of consistent tests. If we consider a single isolated cell and a parametric model, we know that the samples in the rows of the decoded matrix have the following distribution [4]:

$$\begin{aligned} &\text{spurious} \quad Z \sim p\,\lambda_1 e^{-\lambda_1 z} + (1-p)\,\lambda_0 e^{-\lambda_0 z} \\ &\text{correct} \quad Z \sim \lambda_1 e^{-\lambda_1 z} \\ &\qquad\qquad \text{with } \frac{1}{\lambda_1} > \frac{1}{\lambda_0} \end{aligned} \tag{2-9}$$

It is clear that in this situation, the correct row samples are stochastically larger than the spurious row samples. There will be deviations from this model due to several reasons like the effect of adjacent cell interference, the departure from the 'idealness' assumed in arriving at the model, the presence of impulsive noise due to vehicle ignition and so on. However, though the exact distribution is unknown, under these conditions, the correct row samples would still be stochastically larger than the spurious row samples.

Given the two samples (X_i, i=1, ..m), (Y_j, j=1, ..n) the u test computes the statistic

$$U = \sum_{j=1}^{n} \sum_{i=1}^{m} u(X_i - Y_j) \tag{2-10}$$

to test the hypotheses

$$H_0: \ X_i, Y_j \text{ are both from the same continuous distribution}$$

vs.

$$K: \ X_i \text{ is stochastically larger than } Y_j$$

It can be observed that the range of U is any integer between 0 and mn. In general, U will have larger value under the alternative K and a lower value under the null hypothesis H_0.

<u>The Algorithm</u>

Since a decoded matrix has 2^K-1 (K equals number of bits in a word) spurious rows and only one correct row, the algorithm of finding the correct row proceeds by considering two rows at a time and by applying the U test. Since, on a test, both the rows can be spurious or one is the correct row, the test is continued successively on a pair of rows, till the U test identifies the correct row. It can be shown that, on an average, the number of pairs to be tested equals 2^{K-2}. The correct row is declared to be identified whenever

$$U > \eta \quad \text{(or)} \quad U < L^2 - \eta \qquad 2\text{-}11$$

Where η is a threshold to be fixed and L is the number of elements in each row. It is to be noticed that we have to perform the two-sided test, even though the correct row samples are known to be larger, because of the uncertainty in identifying the correct row in a pair. Then, if α is the type I error (reject H_0 given H_0) and β is the type II error (reject K given K) of the U test, we can calculate the probability of correct word decision as

η	α	δ	P_b
240	7.75×10^{-2}	2.95×10^{-3}	0.45
260	1.7×10^{-2}	1.75×10^{-2}	0.30
275	4.3×10^{-3}	5.8×10^{-2}	0.13
280	2.5×10^{-3}	8.2×10^{-2}	0.10
285	$1.4 \quad 10^{-3}$	0.115	9.5×10^{-2}
290	5.2×10^{-4}	0.155	9.16×10^{-2}
295	3.2×10^{-4}	0.2	0.11

$M = 100$, $K = 8$, $L = 19$, $S = 2.5$, $k = 0.5$

TABLE 2-3

η	α	ε	P_b
[illegible]	[illegible] $\times 10^{-2}$	5.8[illegible] $\times 10^{-2}$	0.3
270	[illegible] $\times 10^{-3}$	0.1[illegible]7	0.21
280	2.93×10^{-3}	0.183	0.16
290	8.6×10^{-4}	0.259	0.15

$M = 100$, [illegible]

TABLE 2-4

$$P_W = \sum_{i=1}^{J} (1-\alpha)^{i-1} (1-\beta)/J \qquad 2\text{-}12$$

$$\text{where} \quad J = 2^{K-1}$$

Then the probability of bit error becomes

$$P_b = \frac{2^{K-1}}{2^K-1} \left[1 - \frac{(1-\beta)(1-(1-\alpha)^J)}{\alpha J}\right] \qquad 2\text{-}13$$

with $K = 8$, $L = 19$ the requirement of $P_b \simeq 10^{-3}$ dictates that α has to be very low ($\simeq 10^{-4}$). However, at such low α, the power of the test, namely $(1-\beta)$, seems to be too low. From standard tables, it is possible to read off the threshold η for a given α value. However, to find β we need to resort to simulation.

Using the samples generated, as explained in I, we form the U test and the results are shown in table 2-3. The results of U test corresponding to a parametric model on isolated cell samples is shown in table 2-4. In either situation, the power of the U test is so low to be of any concern to FH-MFSK mobile radio.

2-4 Maximum Rank Sum Test (MRST)

As seen in the previous section, a sequential detection scheme in conjunction with U detector performs poorly. The equivalent of a maximum likelihood test [4] in the non-parametric domain would be a test which picks the row having the maximum rank sum. Therefore, the idea behind a maximum rank sum test (MRST) is to rank order the sum in the decoded matrix by considering the entire ($2^K \times L$) samples. Then, by summing these rank orders across each row, we decide the row with the largest sum as the correct row. Picking the maximum rank sum has received attention in other applications. [5, 6].

Using the samples generated as explained in section 2-1, we performed a MRST. Because of a large simulation time required on the computer, we were able to estimate the probability of word errors exceeding 5×10^{-3}. With adjacent cell interference and with parameters $M = 140$, $S = 2.5$, $k = 0.5$ and 2300 iterations, the bit error rate P_b was found to be $\leq 2 \times 10^{-3}$. With $M = 200$, we find $P_b \leq 10^{-1}$. With $M = 140$, SNR of 25 dB, the bit error rate remains practically the same, when the parametric model on an isolated cell is considered. Thus, the MRSR is also found to possess some 'robustness' against the changes in the probability model. Moreover, the above performance with adjacent cell interference, is close to what is obtainable in a hard-limited receiver with an adaptive threshold. (In an isolated cell, the hard-limiter does outperform MRSR). It is clear that the MRST shows good performance and could well be an alternative to the more complex adaptive parametric receivers.

We continue the analysis of MRSR and a reduced rank sum receiver (RRR) in the next chapter.

III. Nonparametric Receivers

As pointed out in the previous chapter, the Maximum Rank Sum Receiver (MRSR) is expected to perform well under mobile conditions. It should be mentioned that the ties in the rank-sums can be broken by randomization. Intuitively, the MRSR seems to be the best [5]. Here, we apply the asymptotic theory to get an approximate estimate of the probability of bit error encountered in a MRSR and a reduced rank-sum receiver. In figure 3-1, we show the operation of a MRSR by means of matrices.

3-1 Reduced Rank-Sum Receiver

With the values of K=8, L=19 (which are optimum for the parametric receivers), it can be observed that over each $L\tau$ (=T) seconds, ($2^8 \cdot 19$) samples will have to be ranked. This amounts to ranking about 5000 samples in 250 μ sec. Since this may imply considerable complexity, we consider a reduced ranking method. In this method, the ranking will be done by considering the samples in each column only [Figure 3-2]. Since L columns of samples arrive sequentially in time, ranking of 250 samples will be done in τ($\approx$ 12 μ sec.) duration.

(i) Simulation Results

By generating the samples based on the model (2-[illegible]), using IMSL routine, it is straightforward to simulate the receiver performance. The tables 3-1, and 3-2 show the performance of MRSR and RRR. As can be [illegible], both the receivers are nearly identical in performance. At SNR of 25 dB, each could accomodate about 135 users at a probability of bit error of $P_b \approx 2 \times 10^{-3}$. By simulating the samples which take into account the effect of adjacent cell interference [7], the MRSR is tested under this condition. The probability of bit error P_b remains practically the same at 2×10^{-3} (with a controlled average SNR of 25 dB and when the user is at about half the way toward the cell corner ($k \simeq 0.5$)). Some robustness in the performance of MRSR against changing probability model is apparent. It should be mentioned that an extensive simulation study could not be carried out because of excessive simulation time requirement.

SNR = 25 dB		
M	P_b	# Simulation Trials
130	2.12×10^{-3}	4000
140	3×10^{-3}	4500
170	2.3×10^{-2}	1000

Table 3-1

Performance of MRSR (Simulation)

SNR = 25 dB		
M	P_b	# Simulation Trials
140	2.12×10^{-3}	8000
160	1.25×10^{-2}	2000

Table 3-2

Performance of RRR (Simulation)

3-2 Error Rate Estimate Based on Asymptotic Theory

It has been shown that the J ($J = 2^K$) rank-sums are asymptotically jointly normal, for large values of L [8, 9]. For values of L, of the order of 20, we expect the asymptotic theory to be only approximately true. However, the error estimates based on the asymptotic theory show reasonable agreement with the simulation results obtained earlier. Actually, the asymptotic estimates of error rate are slightly on the higher side. This approach allows us to estimate the performance of the receiver under different conditions (for example, for different values of M).

(i) Maximum Rank-Sum Receiver (MRSR):

For the maximum rank-sum receiver, the asymptotic procedure to find the probability of correct selection is readily available in the literature [8]. Denoting

$$g(x,y) = \begin{cases} 1 & x \leq y \\ 0 & \text{otherwise} \end{cases} \qquad 3\text{-}1$$

We write the rank-sum for the p^{th} row as

$$S_p = \frac{(L+1)L}{2} + \sum_{s \neq p}^{J} \sum_{\ell,m=1}^{L} g(x_{s\ell}, x_{pm}) \quad p = 1,2,\ldots J \qquad 3\text{-}2$$

Here, x_{ij} denotes the entry in the i^{th} row and j^{th} column of the decoded matrix of the user.

For large L, it is possible to find $E(S_p)$, $Var(S_p)$ and $cov(S_p, S_q)$ and hence characterize the random variables $(S_1, \ldots S_J)$. Without loss of gener ait; , assume j^{th} row as the correct row. Then, the probability of correct selection (or decision) is

$$P_C = \text{Prob}\ [S_j = \underset{i}{\text{Max}}\ (S_i)] \qquad 3\text{-}3$$

$$= \text{Prob}\ [S_j - S_i \geq 0 \quad i = 1, \ldots J \quad i \neq j]$$

The above equation can be shown to reduce to [8]

$$P_C = \int_{-\infty}^{\infty} \Phi^{J-1}((\sqrt{L}\ a + \sqrt{C}\ \ x)(b-C)^{-\frac{1}{2}})\ d\ \Phi(X) + 0(1/\sqrt{L}) \qquad 3\text{-}4$$

Where $\quad a = J(\eta - \frac{1}{2}) \qquad 3\text{-}5$

$$b = (J^2 - 15\ J - 22)/12 + \eta(3J+2) - \eta^2(J^2+J+2)$$
$$+ \theta(J^2 - J + 2) + \psi(J+2) \qquad 3\text{-}6$$

$$c = \eta(1+2J) - \eta^2(1+J+J^2) + \theta(1+J^2) + \psi(1+J)$$

$$- (11 + 13J)/12 \qquad 3\text{-}7$$

$$\eta = \int F_i(X)\, d\, F_j(X) \qquad 3\text{-}8$$

$$\theta = \int F_i^2(X)\, d\, F_j(X) \qquad 3\text{-}9$$

$$\psi = \int F_j^2(X)\, dF_i(X) \qquad 3\text{-}10$$

F_j is the cdf of the samples from the correct row and F_i $(i \neq j)$ is the cdf of the samples from the spurious rows.

If we assume that F and F_i satisfy the model (2-9), it is possible to evaluate P_C from (3-4). The probability of bit error P_b is given by

$$P_b = \frac{2^{K-1}}{(J-1)} (1 - P_C) \qquad 3\text{-}11$$

(ii) Reduced Rank Sum Receiver (RRR):

For this receiver, the rank-sums are given by

$$R_p' = L + \sum_{s \neq p}^{J} \sum_{m=1}^{L} g(x_{sm}, x_{pm}) \quad p = 1, \ldots J \qquad 3\text{-}12$$

Proceeding along similar lines, the probability of correct selection P_C' is given by

$$P_C' = \int_{-\infty}^{\infty} \Phi^{J-1}((\sqrt{L}\, a' + \sqrt{C'}\, x)(b' - c')^{-\frac{1}{2}})\, d\, \Phi(X) + O(1/\sqrt{L}) \qquad 3\text{-}13$$

Where

$$a' = J(\eta - \tfrac{1}{2}) \qquad 3\text{-}14$$

$$b' = \frac{J^2 - J - 2}{12} + \eta(2J) - \eta^2(J^2) + \theta(J^2 - 2J) \qquad 3\text{-}15$$

$$c' = \eta(J) - \eta^2(J^2) + \theta(J^2 - J) - \frac{J}{12} \qquad 3\text{-}16$$

Therefore, the probability of bit error P_b' for the RRR can be computed as

$$P_b' = \frac{2^{K-1}}{(J-1)} (1 - P_C') \qquad 3\text{-}17$$

The error estimates of these two receivers are summarized in Table 3-3

	P_b	
M	MRSR	RRR
100	9.54×10^{-5}	1.0×10^{-4}
120	7.43×10^{-4}	$8. \times 10^{-4}$
140	4.0×10^{-3}	4.36×10^{-3}
160	1.56×10^{-2}	1.64×10^{-2}
170	2.69×10^{-2}	2.82×10^{-2}

Table 3-3

Asymptotic Error Estimates (SNR = 25 dB)

From the table 3-3 we observe that both the receivers have nearly identical performance. This is not surprising when we observe that large J (J = 256) implies that $b' \simeq b$ and $c' \simeq c$ and therefore the multivariates $\{S_j - S_i;\ i \neq j\}$ and $\{S_j' - S_i';\ i \neq j\}$ have nearly identical distribution. From information theoretic point of view, the divergence between the two distribution tends to zero [10]. In other words, the reduced ranking possesses nearly as much information as the full ranking has. Though not exactly related to this problem (because of large SNR) it is useful to recall that the asymptotic relative efficiency (ARE) of MRSR with respect to RRR equals $(1 + \frac{1}{J})$ [5]. This implies that ARE $\simeq$ 1 for large J.

3-3 Choice of K:

It is difficult to arrive at an optimum value of K which would maximize the performance of MRSR (or RRR) under all probability models. It is not easier, even if the parametric model (2-9) is satisfied. However, through some indirect assessment, the value of K = 8 can be justified. Assuming that (2-9) is the underlying probability model, we compute some form of distance measure between two samples that are obtained under the hypotheses of correct and incorrect selection. The value of K which maximizes the distance is found. Another method is to observe the asymptotic error rate (Section 3-2) as a function of K.

Distance Measures:

Consider the received matrix of size $(2^K.L)$. The parameters K and L are related by [1]

$$L = [rK/2^K] \tag{3-18}$$

$$r = \frac{W}{R_b} = 625 \tag{3-19}$$

Here, [ ] denotes the largest integer operation, W the one-way bandwidth, assumed to be 20 MHz and R_b the bit rate.

Assume that the samples from the correct row have the density function f and those from the spurious rows have the density function g. Then, the situation corresponding to the correct and the incorrect row selection can be depicted as follows:

H: Correct selection, L number of f samples identified

N: Incorrect selection, L number of g samples identified.

Equivalently, the above alternatives imply

H: Correct selection, Information conveyed by [(J-1) L number of g samples]

N: Incorrect selection, information conveyed by [(J-2) L number of g samples and L number of f samples].

Therefore, any of the known distance measures [10, 11] can be computed for the density functions under H and N. We present here only the divergence J* and the Bhattacharyya distance B.

Divergence:

The divergence J* can be written as a sum of two components called the directed divergences [10].

$$J^* = I(H, N) + I(N, H) \tag{3-20}$$

where

$$I(H, N) = \int_{\underline{x}} f_H(\underline{x}) \; \ell n\left(\frac{f_H(\underline{x})}{f_N(\underline{x})}\right) \; d\underline{x} \tag{3-21}$$

and I(N, H) is obtained by interchanging H and N in the above equation.

Since all the samples are independent, it is easy to observe that

$$I(H, N) = I_L(f, g) \tag{3-22}$$

Here, $I_L(f,g)$ denotes the directed divergence between L number of f samples and L number of g samples. Once again, due to independence among the samples,

$$I_L(f,g) = L\, I(f,g) \tag{3-23}$$

where $I(f,g)$ is the directed divergence between the densities f and g. That is,

$$I(f,\ g) = \int_0^{\infty} f(x)\ \ln\left(\frac{f(x)}{g(x)}\right)\ dx \tag{3-24}$$

Therefore,

$$J^* = L(I(f,\ g) + I(g,\ f)) \tag{3-25}$$

When f and g satisfy (2-9), we can compute J^* as a function of K. The results are shown in figure 3-3

Bhattacharyya distance:

The Bhattacharyya distance B between the two densities f_H and f_N is given by

$$B = -\ln\left[\int_{\underline{x}} \sqrt{f_H(\underline{x})\ f_N(\underline{x})}\ \ d\underline{x}\right] \tag{3-26}$$

Because of sample independence, this reduces to

$$B = -L\ \ln\left[\int_0^{\infty} \sqrt{f(x)\ g(x)}\ \ dx\right] \tag{3-27}$$

If f and g satisfy (2-9), B can be computed as a function of K. The results are shown in figure 3-4.

As an alternative method, we can observe the effect of K on the asymptotic error rate (see figure 3-5). By observing figures 3-3 through 3-5 it can be seen that K = 8 is nearly optimum under any of these performance measures.

The optimization procedure based on distances are normally employed in parametric situations, when the probability of error can not be easily found

[11]. We assumed that such procedure could also be applied to nonparametric tests operating under a known probability model. This is partially justifiable since the ranking does carry some information contained in the original samples.

Conclusion

Considering the base to mobile transmission, it is found that MRSR or RRR could accomodate about 135 users at $P_b \approx 2.10^{-3}$ and at an average SNR of 25 dB. With the simulated adjacent cell interference, the performance of MRSR remains practically the same (i.e. $P_b \approx 2 \times 10^{-3}$ at a controlled SNR of 25 dB, with receiver at about half the way toward the base station). Thus, MRSR (or RRR) shows some robustness against changing probability model. Moreover, the adaptive parametric hard-limited receiver accomodates only about the same number of users as the MRSR, when adjacent cell interference is taken into consideration. Also, the limited simulation study and asymptotic theory reveal the nearly identical performances of MRSR and RRR. As has been said earlier, it is much simpler to implement the reduced rank sum receiver than to implement MRSR or a parametric receiver. Therefore, one concludes that RRR is a possible competitor to the parametric receivers for FH-MFSK mobile radio.

There are still some problems to be solved in arriving at the error estimates. (i) The goodness of asymptotic estimate is not known, except for the reasonable agreement with limited simulation results, (ii) Some methods are to be found to reduce the excessive simulation time, thereby allowing extensive performance estimation under various conditions.

Finally, (iii) the effect of correlation between the samples on the performances of parametric and non-parametric receivers will have to be estimated and compared.

IV. Signal Processing for Packet Radio in Computer Communication Network

Packet radio has emerged as a viable technology for both fixed and mobile computer communications. Here, first an overview of the whole organization of the computer communication network including packet switch processing and architecture, packet radio network and application of spread spectrum technique for packet radio network is presented. Subsequently, we discuss the signal processing aspect of the packet radio network and alternative technologies available for their implementation.

4-1 Introduction

Why Computer Communication Networks

The ability to communicate data (i.e. bits) over communication lines is relatively old. Modems for a variety of communication lines (especially telephone lines) at a variety of speeds, have existed for many years. However, due to rapidly increasing demands, the traditional voice communication (telephone) networks have failed to provide the required services for data transmission in a computer communication environment. This has given rise to the existence of the computer communication (data) networks. [18, 22]

Why Packet Switching for Computer Communication

The packet-switched communication system has established [23] itself as an attractive option in a wide variety of digital transmission environments including data, voice and in perspective video signals. The switching capability is what makes a collection of communication lines a communication network. This switching allows the interconnection of N users without having to use facilities of N^2 size. The total capabilities of a communication network depend on both the communication capabilities of the individual links and the processing capabilities of the switching nodes.

There are two basic approaches to the switching issue [23], the circuit switching (CS) and the packet switching (PS). It has been established that under the computer communication environment, packet switching offers much better performance than circuit switching. Furthermore, there are two ways of providing communication services on top of either CS or PS systems. One is by virtual circuits (VC) and the other by datagrams (DG). The choice [22] depends on the traffic distribution and the characteristics of the resources in each particular situation. Attempts are made to optimize the most important attributes of computer communication network: reliability, data rate, data integrity and end-to-end delay.

Also, typical computer communication traffic is quite bursty in nature, and requires a low-duty cycle of high-data-rate communication. Therefore, in nearly all circumstances packet switching is the preferred technology for implementing cost-effective computer communication.

What is Packet Switching

- Packet switching is a transmission technique for sending data.
- Data is segmented into small packets
- Small packets find their way through system individually and over the best of multiple transmission paths available between

nodes at the moment of transmission (adaptive routing over distributed network).

- All packets entering and moving through system are under constant flow control - eliminating the need for high intransit storage.

What is Packet Radio

Packet radio is the application of packet switching technique to radio channel. It is a communications network which sends packets of data by radio. In the packet switching aspect of packet radio, the data from each user is bundled into packets, transmitted through a series of store-and-forward (S/F) nodes, and delivered at the destination. The radio aspect of packet radio is its use of a shared broadcast channel to link network nodes.

Although the packet radio system (PRS) is primarily conceived as a data network, voice communication is possible [13]. In fact, it has been established that packet radio (packet-switching computer communication networks) is capable of supporting real-time speech communication [22].

4.2 PACKET FORMAT

In packet communication, the voice or data information stream is divided into small segments, called packets, which are transmitted one-by-one through the network. Each packet has added to it "header" bits used for addressing, routing, error-checking, and other overhead purposes. If message length exceeds the longest packet length allowed by the system, the message is divided into packets at the originating node and reassembled for delivery at the destination node. It is convenient to think of packets as envelopes into which data is placed for delivery to its destination.

In an experimental packet radio [20], a transmitted packet has the structure shown in figure (4-1). It consists of a 48 bit preamble followed by a variable length header (typically 96-144 bits), followed by the text and a 32 bit check sum.

- The packet preamble is used by the radio section of the receiver for several purposes. The first few bits are used to detect the carrier energy and to set the automatic gain control (AGC) to compensate for differing signal strengths of the arriving packets. Correct reception of the packet is totally dependent upon acquisition of the preamble. The next few bits are used to acquire bit timing. Following these, the next set of bits is used to acquire packet timing (identify the end of the preamble and the start of the header).

- Both the header and text are delivered from the radio section to the digital section which knows the header format and can therefore determine the exact start of the text.

- The error control bits consist of a checksum appended by the transmitter and checked by each receiver. After checking, the error control bits are stripped off by the radio section as was the preamble before it. The digital section checks a status register in the interface to determine if the packet is correct.

4.3 PACKET SWITCH PROCESSING

In packet switching processing [20], incoming packets are first stored in memory (capable of storing 4-6 packets) through a storage allocation process. The pointer indicating the address location of each stored packet is then placed in an input queue. Input queue pointers are examined in turn by the switching process, and upon identification of the address of a packet the routing table is examined to determine the preferred outgoing link for each of the possible packet destination addresses. The packet address pointer is then placed in the appropriate queue of the outgoing trunk. After packet pointers are examined at the output queue, packets are located in storage, then dispatched over the outgoing trunk to the next switching (node). Routing of the packets can be virtual or adaptive.

4.4 PACKET SWITCH ARCHITECTURE

The switch performs the functions of reception, storage, switching, processing, error control, routing and retransmission of data packets. This packet switch was originally designed [15] using a microprocessor and other electronic devices. However, the growing need for high speed digital communication demands still faster processing. Fortunately, optical processing is known to be much faster than electrical processing [12]. Recently, an attempt has been made to replace electronic components by optical devices in the packet switch. In this new opto-electronic packet switch most of the electronic components except the microprocessor are replaced by components based on optoelectronic devices like diode lasers, LED's, photodetectors, crossover optical waveguide switches, optical shift registers, multiplexers and optical fibres. Optical fibres used in this ultra-modern switch is only for intercomponent connections and are very short in length. This reduces the possibility of attenuation of optical signals in the fibres. In application, where fast processing and smaller size of the switch are desired (e.g. on board a communication satellite), the optical packet switch seems a very attractive candidate.

4.5 SPREAD SPECTRUM TRANSMISSION IN PACKET RADIO NETWORK [2[illegible], 3[illegible]]

Network aspects of spread-spectrum make spread spectrum transmission an attractive candidate for packet radio network environment. Although a packet radio system need not employ spread spectrum, there are several noteworthy attributes arising from its use.

The use of spread spectrum waveforms in a packet radio system is motivated largely by the desire to achieve good performance in the fading multipath channels resulting from non-sited mobile system users, by the need for coexistence with other systems and for antijamming capabilities in tactical applications. Use of spread spectrum waveform facilitates some additional benefits such as a strong capture capability to enhance access efficiency, the potential for an integrated position location feature, and the ability to operate links, nets, and subnets on pseudo-orthogonal waveforms using 'code-division multiple access' (CDMA).

However, these capabilities are not received free of cost. Use of a fixed spread spectrum waveform adds a modest amount of system complexity, while a time varying spread spectrum waveform requires some degree of network time synchronization and may require added system protocols to effect the distribution of variables to control waveform generating algorithms.

The use of spread spectrum, although desirab : for many applications, is not an apriori requirement for a packet radio system. The most commonly used forms of spread spectrum waveforms are direct sequence pseudo-noise (PN) modulation, frequency hopped (FH_ modulation, and hybrid combinations of the two.

4.6 PACKET RADIO NETWORK [17 - 20]

A typical packet radio network consists of three primary functional elements: terminal, station and repeater. A typical network is shown in figure (4-2). The terminal is the user's interface to the network. The station has the responsibility for over-all management of the network including initialization, routing, flow control, directory, and accounting functions. It also serves as the gateway from the network to other networks. In a network covering a small area, terminals and a station suffice. However, terminals must have limited power to be portable, and this power limits their range. To provide coverage over extended areas, repeaters are needed. The repeater has the function of extending the range of station-terminal links and providing a mechanism for distributing the network management logic. It, therefore, receives and retransmits packets with the additional responsibilities of detecting errors and invoking routing protocols dictated by the station.

A repeater element contains a radio section which provides access to the network radio channel and a digital section which performs the logical functions of error control and packet routing. Figure (4-3) illustrates this organization of the repeater element.

An experimental packet radio (EPR) may operate as a repeater, or may be connected to a user's host computer or terminal, or to a station. The interface between the user equipment and the EPR digital unit is the portal through which packets enter and leave the network.

4.7 PR SIGNAL PROCESSING [20, 21, 24, 26]

The packet radio signal format is as follows:

- Frequency Band: 1710-1850 MHz
- Spread Spectrum Modulation: MSK 12.8 MHz chip rate.
- Data Modulation: DPSK at data rate 100 or 400 Kbps.

A comparative analysis for coherent versus non-coherent processing for the packet radio has shown that

* Demodulation performance (i.e. BER) of the coherent processor is superior to non-coherent processor in the absence of doppler shift.
* Random single errors in coherent PSK while clusters of errors in DPSK. It is easier to implement a design to cope with single errors when forward error correction (FEC) is used.
* Coherent demodulation is adaptive such that improvement in SNR is greater for lower SNR.
* A nearly optimum multipath matched filter results when coherent reference is operated in conjunction with an integrator. This results in a marked improvement in a fading environment.
* Delay of 2.5 μ sec. can be selected for coherent integration. This results in performance enhancement of the 400 Kbps system.

From these findings it is clear that coherent processing will improve the performance of the packet radio.

The repeater element is chosen for discussion since it is an internal network element and invariant to the user interface or specific application. Additionally, it contains most of the needed functions of all network elements. The function of a packet radio repeater is to extend the range of terminal-station links, therefore, it must receive and retransmit packets. It must also invoke routing protocols and error control. For these reasons, the repeater must have access to the radio channel and contain logical decision capability to carry out the routing and error control tasks.

The repeater consists of

- a radio
- a digital processing section
- software for the processing.

The radio portion of the repeater is a transceiver that operates in the 1 to 2 GHz frequency range. The transceiver transmits and receives over a common channel via one omnidirectional antenna.

The signal processing section includes the encoding and modulation processes in the transmitter and the automatic gain control (AGC), demodulation, detection, and synchronization processes in the receiver.

The encoding/modulation process functional diagram is shown in figure (4-4). The data is differentially encoded to avoid the necessity of reconstructing a phase-coherent reference at the receiver. A read only memory (ROM) is used to store a pseudo-random spread spectrum chip code. The differentially encoded data gates the sequence to the impulse generator. The impulse generator impulses a 2-chip long cosine-weighted surface acoustic wave device (SAWD) every chip interval. The resultant output of the SAWD is the transformation of a data bit into a multichip spread spectrum waveform. The chips are minimum phase shift keyed (MSK) which yields a waveform that is constant amplitude and phase continuous.

The signal processes in the receiver are shown in the functional diagram of figure (4-5). The recieved signal, after passing through the RF pre-amplifier, is down converted to the IF frequency. Then, it is amplified by an automatic gain control (AGC) IF amplifier. The output of the AGC amplifier drives a SAWD matched filter which may be viewed as a tapped delay line. The SAWD property allows significantly long delays (up to many micro seconds). By proper design, any causal impulse response can be realized that is within the bandwidth of the device. The SAWD impulse response is reverse-time-matched to the pseudo-random spread spectrum code sequence described in the encoding/modulation process. When the received signal is identical to the SAWD impulse response, the output of the SAWD matched filter is the autocorrelation function of the sequence. The signal passes through two identical SAWDs and the outputs of the two SAWDs are summed and subtracted in the 180° hybrid. This allows the decoding of differentially encoded data by comparing the pseudo-random sequence autocorrelation of a bit to the previous bit's autocorrelation. The sum and difference outputs are each AGC-amplified and envelope detected on separate but identical channels. The outputs of the two channels provide the inputs to the data detector. The data detector consists of a differential comparator that compares the sum and difference inputs and decides the data is a 1 or a 0.

The other processes needed in the receiver signal processing are the bit synchronization circuit and the end-of-preamble detector. The bit synchronization circuit gates the data detector output through a narrow time window to provide time and multipath discrimination. The bit synchronization circuit phase locks to the sum of the difference and sum channels so that the time window is synchronous with the autocorrelation peaks of the SAWD matched filters.

The purpose of the preamble detection circuit is to properly define the beginning of a packet. This is accomplished by attaching a preamble to the front of a packet which consists of a sequence of Barker codes. The Barker codes are selected because of their high peak-to-sidelobe autocorrelation property. The preamble detector is a digital matched filter matched to the Barker bit sequence.

The path loss and multipath environment in repeater-repeater links are less severe than in the terminal-repeater links where the terminal is operating in a mobile environment. In order to take advantage of this and maximize network throughput, the repeater radio uses two data detectors with different data rates. The lower data rate of 100 KB/s is used for terminal-repeater traffic and the higher data rate of 400 KB/s for repeater-repeater traffic. The receiver signal processing is, therefore, duplicated for these two data rates and the pseudo-random spread spectrum codes for the 100 KB/s and 400 KB/s are 128 and 32 chips long, (for 12.8 MHz chip rate and delays of 10 μ sec. and 2.5 μ sec.) respectively. This allows the occupied bandwidth to be held constant for both data rates. Pseudo-random codes could be selected for the two data rates to give low cross-correlation between the two rates and therefore making possible simultaneous reception of high and low data rate packets in the repeater radio.

The packet transmission protocols consisting of repeater initialization, packet routing, packet acknowledgements, and error control demand that the repeater processes significant logical processing power. The digital section of the repeater provides this processing ability. It controls the radio and performs the following functions: packet reception and transmission, error detection, packet routing protocols, and acknowledgement protocols. The hardware of the digital section consists of the CPU, address register decode, direct memory access control (DMA), radio interface and control and memory.

The software portion for the radio repeater system is a multiprogrammed, interrupt driven system. Two independent programs coexist in the system and the state of the system is saved as control is transferred from one program to another. The operating system is interrupt driven with program control being transferred and processing initiated as a result of CPU interrupts. The system is structured into three programs which are defined as executive, background and foreground. The 'executive program' is utilized for operating system initialization, program control, and system test aids. The 'foreground program' contains the radio I/O packet handling process. The 'background program' provides for overlay programs, on-line diagnostics, and performance monitoring.

In the recent developments of packet radio repeaters, VLSI technique for implementing digital section and hybrid thin film circuitry for the radio section is being considered. Better spread spectrum processing gain (about 30 dB or more) can be achieved with convolver (a device which accumulates the cross-product of two signals) surface wave devices.

4.8 SIGNAL PROCESSING ALTERNATIVES [16, 25 - 29, 31]

Packet radio signal processing requires:

- A filter matched to the spread spectrum waveform.
- Some form of non-coherent (post-detection) integration to enhance signal-to-noise ratio (SNR) for successful acquisition.
- Means for synchronization
- DPSK data demodulator.
- Barker Code detector to determine the start of the message
- Error-detection circuit

The Barker code detector and error-detection circuit are implemented using digital technology. The matched filters, non-coherent integrator, synchronization circuit, and data demodulator can all be implemented in a number of ways with differing technologies.

Three competing technologies - DIGITAL, SAW and CCD are applicable to the required processing needs. With present technology, both digital integrated circuits (IC) and CCD implementation can be used up to a bandwidth of approximately 10 MHz whereas SAW devices can be used up to 100 MHz. Also, SAW devices are simpler to fabricate and can be less expensive to produce than CCD's. SAW devices operate at RF whereas CCD's operate at baseband, therefore former is a better choice for multiplication needs.

For high bandwidth real-time applications SAW technology is chosen, whereas for lower bandwidth real-time applications CCD's are used. For non-real-time operation, IC's and μP are probably the best implementation.

* The SAWD is the natural device for implementing an MSK chip filter [16], either transmit or receive, at IF.

* Two matched filter options are available - SAW and CCD. Digital matched filters were rejected due to their inherent performance limitations and high power usage and size disadvantage. The SAW filter uses LSI digital and analog circuits in a hybrid package. The CCD is a baseband device, smaller, more reliable and temperature insensitive. The SAW is a RF device, uses less power and requires no down conversion to base-band.

* The preamble matched filter is used to detect the occurrence of the final 13-bit Barker-coded sequence indicating the start of message. A digital approach is the best due to its simplicity and low-cost.

* The non-coherent integration can be implemented at base-band. Either CCD or digital devices (or a hybrid) are used.

* DPSK demodulation requires either a dual matched filter or a matched filter/delay line implementation. Both the matched filter and delay line use a single technology.

4.9 CONCLUSION

We have focussed on the signal processing of the repeater portion of the packet radio network because it is the critical communications element in a packet radio network. It can be converted to a station by adding a mini-computer, and a terminal by adding appropriate I/O components.

We feel that still there exists sufficient room for further modifications and developments in packet switch architecture, packet radio signal processing and in packet format in order to optimize the overall performance of the communication links.

The recent advances in computer technology, sophisticated signal processing techniques and elegant implementation technologies like digital IC, CCD, SAW are indications of the optimum packet radio networks in the future.

Very recently [14], the low cost packet radio (LPR) which is a new version of a packet radio, has arrived. LPR has high reliability, low weight, low cost and reduced volume while providing forward error correction capability. The LPR may provide affordable means in supporting the future large-scale network. Finally, it is highly probable that packet radio will play a significant future role in computer communications and the local distribution of information.

V. Performance Evaluation of a Protocol for Packet Radio Network in Mobile Computer Communications

The need to provide computer network access to mobile terminals and computer communications in the mobile environment has stimulated and motivated the current developments in this area. Packet radio technology has developed over the past decade in response to the need for real-time, interactive communications among mobile users and shared computer resources. In computer communication systems we have a great need for sharing expensive resources among a collection of high peak-to-average (i.e. bursty) users. Packet radio networks provide an effective way to interconnect fixed and mobile computer resources. This paper presents the results of an attempt to study the performance of the mobile packet radio network for computer communications over degraded channels. We develop a model under fading conditions and derive a protocol for evaluating the performance of the mobile packet radio network (MPRNET) in terms of the packet error rate, packet delay and average number of retransmitted packets per cycle. The analytical results are presented and numerical examples are given to illustrate the behavior of these performance criteria as a function of packet transmission rate, packets transmitted per cycle, packet size, and vehicle speed with the help of appropriate plots.

5.1 Introduction

The need for fixed or mobile computer communications over long-haul or local area networks is widely understood now. Many mobile radio users are interested in data stored in computer files and it appears logical to transmit this data in a digital format. Large computer installations, enormous data banks, and extensive national and international computer networks are now becoming available. They constitute large expensive resources which must be utilized in a cost effective fashion. The constantly growing number of computer applications and their diversity render the fundamental problem of accessing these large resources. Fortunately, radio communication has emerged as a method for providing remote terminal access to computers [45].

Recently, we have witnessed the development of packet radio technology to achieve information distribution and computer communications. This development is directly related to the rapidly increasing demand to provide effective communication services for data distribution. Multiple access and broadcast radio channels have been utilized to form networks which provide packet-switched communication. These packet radio networks are well-suited for computer communications in the ground mobile network environment, due to its rapid and convenient deployment capability, easy configuration possibility and survivability. In addition, packet radio network technology offers a highly efficient way of using a multiple access radio channel with a potentially large number of mobile subscribers to support computer communications and to provide local distribution of information over a wide geographic area and its area coverage and connectivity may be increased easily [41]. A packet radio network can also coexist with other packet radio networks.

Packet radio technology is applicable to ground-based, airborne, seaborne, and space environments and is able to serve users whether on land, at sea, or in the air. Ground-based networks encounter perhaps the most difficult environment in terms of propagation and RF connectivity. Ground radio links, particularly when mobile terminals are involved, are subject to severe variations in received signal strength due to local variations in terrain, man-made structures and foliage. In addition, reflections give rise to multiple signal paths leading to distortion and fading as the differently delayed signals interfere at the receiver. As a result of these phenomena, RF connectivity is difficult to predict and may abruptly change in unexpected ways as mobile terminals move around. However, an important attribute of a packet radio system is its self-organizing, automated network management capability which dynamically discovers RF connectivity as a function of time for use in packet routing [48]. In mobile packet radio networks (MPRNET) the radio connectivity changes frequently because of the mobility of some of the PRU's (Packet Radio Unit). As the PRU's move, they lose and gain radio connectivity with each other at a rate that can be as high as several changes per minute in urban areas. Due to loss of connectivity in MPRNET a severe problem of route failure arises because of the creation of "route loops" or "a dead end". Restoration of the route is speeded up by making use of additional information in the neighborhood of the failure. This is especially important in a mobile environment due to the high frequency of altered connectivity [9].

The ground radio applications of packet communications include such things as communications among moving vehicles (e.g. taxicabs, ambulances, police cars, fire trucks, private fleets, etc.), communications among aircraft, and indeed communications among any mobile units or any widely distributed units in a sparse environment [36]. Packet radio networks should support mobile terminals and computers at normal vehicular ground speeds within the area of coverage with full connectivity. For ground mobile radio, network diameters on the order of 100 miles are appropriate, but the system architecture should allow the geographic area of coverage to be expanded at the expense of increased end-to-end (ETE) delay across the network [48]. One of the difficulties faced in a mobile network is that the number of users in a given RF connected area and the amount of traffic these users generate as a function of time is difficult to predict [39].

A mobile computer communication network can generally be defined by the following features: its host computers and terminals, communication processors, topological layout, communication equipment and transmission media, switching technique, mobile unit and protocol design [44]. These features are chosen to accomplish the function of the network subject to specified performance requirements. The performance measures most commonly quoted include message delay, message throughput, error rate, reliability, and cost. When mobile operations are involved, the measurements indicate temporary degradation in the performance, affecting both throughput and delay. By proper selection of the dominant network protocol parameters, the degradation can be substantially reduced [42]. Improved performance under mobile operations is needed for all traffic types, to reduce the load on the radio channel and improve overall network performance. Several methods for such improvements have been discussed in [42].

The first analysis of packet radio performance assumed that packet collisions were the major cause for loss of a packet and subsequent retransmission. More recently [46], efforts to design packet radio systems to operate over degraded channels have been undertaken. The channel throughput and packet delay, the two primary performance criteria in computer communications, have been extensively studied for basic system concepts such as pure ALOHA, slotted ALOHA, and CSMA [45, 47]. However, we need to consider the effect of link errors due to noise and fading too. In the absence of fading the noiseless assumption is quite good, but on a fading channel the signal-to-noise ratio becomes a critical parameter. The approach here is to model the problem under fading conditions and develop a protocol for evaluating the performance of the mobile packet radio network in terms of packet error rate, packet delay and average number of packets retransmitted per cycle, as a function of packet transmission rate, packet size, the number of packets transmitted per cycle, and vehicle speed.

5.2 Model Description

Experiments in urban areas have shown that noise impulses occur every few milliseconds in both the UHF and L bands, principally due to automobile ignition noise. A packet has a very high probability of encountering one or two impulses and therefore some form of error correction is required [49], if essentially an error-free performance for computer communication is desired. A target objective of no more than one undetected packet error per 10^6 packets assuming 1000 bit packets, a 100 K-bit/sec. data rate and 100 percent occupancy.

In the ground-based mobile packet radio network performance degradation occurs due to transmission errors resulting from packet collisions, noise, fading, and probably shadowing too. The present model assumes that transmission errors are caused only by fading i.e. errors due to other sources of noise and interference are not considered. Fading occurs due to multipath propagation of the signal in which nulling or reinforcement of the direct path signal results [38]. Fading phenomena are often characterized by a specific type of short-term multipath signal reception whose amplitude follows the Rayleigh distribution [33, 38].

Packet radio techniques are used for communications between mobile terminals and computer networks. In these techniques, a message is decomposed into a number of packets which are transmitted individually to one or more destinations where they are assembled to reconstruct the original message. An overhead message is attached to each packet. The overhead message contains information about the addresses of the originating source and the destination, routing information and check-sum bits for error detection [38, 37].

Let,

L = Message length (bits)

B = Packet length (bits)

b = Overhead Message length (or, packet overhead) (bits)

M = Number of Packets in the message

R = Packet transmission rate (bps)

Hence,

Total packet length = $(B + b)$ bits

Packet duration, $T = \{\frac{(B+b)}{R}\}$ seconds

Also,

$$M = \frac{L}{B}$$

The transmission of packets is conducted in cycles. Each cycle consists of the transmission of N consecutive packets plus a short time interval to allow for the reception of the acknowledgement message [37, 38].

5.3 Protocol Description

In any communication system, and in particular, a computer communication system, it is essential to have a set of well-designed basic control procedures to insure efficient, correct, and smooth transfer of information in the system. Traffic management is a set of rules that ensures the smooth and orderly exchange of information among elements of a computer network. Its main functions are protocol, routing, flow control, and monitoring [43]. Protocol is a word borrowed from the terminology of political diplomacy to describe the rules governing orderly exchange of information between different computing equipment in a predetermined fashion [44].

We describe and analyze a protocol derived from the "stop-and-wait" control procedures [37]. According to this protocol, the transmitting unit sends one packet of data at a time and waits for an acknowledgement (ACK) from the receiving unit before proceeding. If the transmitting unit does not receive an ACK after a certain time-out period, the same packet is retransmitted. This operation is repeated for a predefined number of times until the ACK is received. However, when the channel is unreliable, the transmitting unit may be instructed to give up retransmitting the packet [38]. We assume that the acknowledgement traffic is carried by a separate channel and is received reliably. In this protocol, the packet transmission is conducted in cycles. Each cycle represents the transmission of N packets plus an ACK time-out period. The ACK message informs the sending unit of the first packet that was found in error such that in the following cycle this packet and the following ones are to be retransmitted along with some new packets. Based on an estimation of the signal level at the receiving location or on the frequency of packet errors, the number of packets per cycle, N, can be adjusted.

The protocol is conducted as follows [32, 38, 44]:

(1) At the start of each cycle, the transmitting unit starts transmitting N packets before it stops for a short interval, t_a.

(2) The receiving unit sends an ACK signal immediately after the reception of the N packets indicating the address of the first packet that was found in error.

(3) The addressed packet and the packets following it are then rescheduled for the transmission in the following cycle.

An important parameter of the algorithm described above is the number of packets per cycle, N. For a certain set of system parameters, N can be adjusted to provide the minimum delay per message. More important is the fact that N can be made adaptive to the status of the channel to achieve minimum delay in the ever changing environment of the mobile system [34]. We notice that the case when N = 1 represents the well-known stop-and-wait strategy, and the algorithm described above is a generalization of this strategy.

5.- Analysis

Assumptions

In the analysis, we make several simplifying assumptions to reduce the complexity of the problem. These assumptions may not model the real world, however, they lead to some useful and interesting results.

(1) The traffic introduced into the network consists of fixed-length packets generated according to a Poisson Process, i.e. the packets are introduced into the network according to a Poisson Process.

(2) A packet radio node can be either in the receiving mode or in the transmitting mode. If a packet radio node is in the transmitting mode when a packet arrives, it is lost.

(3) The Carrier-Sense Multiple Access (CSMA) is a quite suitable Channel Access Protocol (CAP) for the present problem [45].

(4) Individual user transmissions are independent of one another and that successive user packet transmissions are independent.

(5) The transmission errors will occur only because of the signal fade below the receiver threshold level, which means that thermal noise, ignition noise and different sources of interference and signal distortion have negligible effects in the presence of fade.

(6) The channel is in one of two possible states at any time [38]:

(i) The ON state represents the case when the received signal is above the threshold level, and

(ii) The OFF state represents the case when the signal is below the threshold level.

This representation is illustrated in figure (5-1). This assumption is justified for most digital modulation techniques which usually exhibit a sharp threshold behavior [34].

(7) The packet is received correctly iff the whole packet was contained in a non-fade interval, t_2 (figure 5-1). This means that the packet is assumed to have at least one detected error if it overlaps to any extent with a fade slot.

(8) The non-fade interval, t_2, is exponentially distributed.

(9) The envelope of the fade interval, t_1, is Rayleigh distributed.

(10) Outbound channels considered and all terminals are within range and in line-of-sight of each other.

Let, t_a = Acknowledgement delay

t_1 = Fade interval

t_2 = Non-fade interval

t_3 = Inter-fade interval

τ = Packet delay (or wasted time) per message due to channel error.

η = Average number of retransmitted packets per cycle.

The density function of the variable "t_2" can be written as,

$$f_{t_2}(t) = \frac{1}{T_2} \exp(-\frac{t}{T_2}) \quad , \quad t > 0 \qquad 5\text{-}1$$

where "T_2" is the average value of the variable t_2.

The probability that all N packets are received correctly (i.e. zero retransmission) is given by the probability that the non-fade interval t_2 will last for a period longer than the N-packet's transmission time.

$$P(0) = \frac{T_2}{T_3} P[t_2 > NT]$$

$$= \frac{T_2}{T_3} [1 - P[t_2 \leq NT]]$$

$$= \frac{T_2}{T_3} [\exp -(\frac{NT}{T_2})] \qquad 5\text{-}2$$

This is also the probability of success P_S for N packets of duration T seconds each to be received correctly without any retransmission. Here, "T_2" and "T_3" are the average values of the random variables t_2 and t_3, respectively.

It follows that the packet error rate is given by

$$P_b = 1 - P_S = 1 - P(0)$$

$$= 1 - \frac{T_2}{T_3} [\exp -(\frac{NT}{T_2})] \qquad 5\text{-}3$$

In general, the probability that n out of N packets are retransmitted is:

$$P(n) = \frac{T_2}{T_3} \; P[(N-n)T < t_2 < (N - \overline{n-1})T]$$

$$= \frac{T_2}{T_3} [\exp -(\frac{(N-n)T}{T_2}) - \exp -(\frac{(N-n+1)T}{T_2})] \qquad 5\text{-}4$$

This equation is valid for $n = 1$ to $(N-1)$.

The probability that N packets are to be retransmitted is:

$$P(N) = \frac{T_1}{T_3} + \frac{T_2}{T_3} \; P[t_2 < T]$$

$$= \frac{T_1}{T_3} + \frac{T_2}{T_3} \; [1 - \exp -(\frac{T}{T_2})] \qquad 5\text{-}5$$

where, first term represents the possibility of transmission cycle starting at a fading slot, while the second term represents the case where non-fade interval is less than the duration of a single packet.

Equations (5-2), (5-4), and (5-5) satisfy the obvious condition,

$$\sum_{n=0}^{N} P(n) = P(0) + \sum_{n=1}^{N-1} P(n) + P(N) = 1 \qquad 5\text{-}6$$

The expected value of the retransmitted packet per cycle is given by,

$$\eta = \sum_{n=0}^{N} n\, P(n) \quad = \sum_{n=1}^{N} n\, P(n) \tag{5-7}$$

$$= \sum_{n=1}^{N-1} n\, P(n) + N\,P(N)$$

$$= \frac{T_2}{T_3} \sum_{n=1}^{N-1} n[\exp -(\frac{(N-n)T}{T_2}) - \exp -(\frac{(N-n+1)T}{T_2})] +$$

$$N\left(\frac{T_1}{T_3} + \frac{T_2}{T_3}[1 - \exp -(\frac{T}{T_2})]\right)$$

$$= \frac{T_1}{T_3} N + \frac{T_2}{T_3} N[1 - \exp -(\frac{T}{T_2})] + \frac{T_2}{T_3}[\exp -(\frac{NT}{T_2}) - \exp - \frac{(N+1)T}{T_2}] \cdot$$

$$\sum_{n=1}^{N-1} n \exp(\frac{nT}{T_2})$$

$$= \frac{T_1}{T_3} N + \frac{T_2}{T_3}\left\{N[1 - \exp - (\frac{T}{T_2})] + [\exp -(\frac{NT}{T_2}) - \exp - \frac{(N+1)T}{T_2}] \cdot\right.$$

$$\left.\sum_{n=1}^{N-1} n \exp(\frac{nT}{T_2})\right\}$$

$$= \frac{T_1}{T_3} N + \frac{T_2}{T_3}\left\{\{\exp -(\frac{NT}{T_2}) - \exp - (\frac{(N+1)T}{T_2})\}\{\frac{N[1 - \exp - (\frac{T}{T_2})]}{[\exp - (\frac{NT}{T_2}) - \exp - (\frac{(N+1)T}{T_2})]} +\right.$$

$$\left.\sum_{n=1}^{N-1} n \exp(\frac{nT}{T_2})\}\right\}$$

$$= \frac{T_1}{T_3} N + \frac{T_2}{T_3}\left\{\{\exp - (\frac{NT}{T_2}) - \exp - (\frac{(N+1)T}{T_2})\}\{N \exp(\frac{NT}{2}) + \sum_{n=1}^{N-1} n \exp(\frac{n\,T}{T_2})\}\right\}$$

$$= \frac{T_1}{T_3} N + \frac{T_2}{T_3} [\{\exp - (\frac{NT}{T_2}) - \exp - (\frac{(N+1)T}{T_2})\}\{\sum_{n=1}^{N} n \exp(\frac{nT}{T_2})\}]$$

$$= \frac{T_1}{T_3} N + \frac{T_2}{T_3} [\{\alpha^{-N} - \alpha^{-(N+1)}\}\{\sum_{n=1}^{N} n\,\alpha^n\}] \qquad 5\text{-}8$$

where, $\alpha \triangleq \exp(T_1' T_2)$

We know that

$$\sum_{n=1}^{N-1} \alpha^n = \sum_{n=0}^{N-1} \alpha^n - 1 = \frac{1-\alpha^N}{1-\alpha} - 1$$

$$\sum_{n=1}^{N-1} n\,\alpha^n = \frac{\alpha}{(1-\alpha)^2} [N\alpha^N - N\alpha^{N-1} + 1 - \alpha^N]$$

$$\sum_{n=1}^{N} n\,\alpha^n = \frac{\alpha}{(1-\alpha)^2} [N\alpha^N - N\alpha^{N-1} + 1 - \alpha^N] + N\alpha^N \qquad 5\text{-}9$$

Substituting (5-9) into (5-8), a closed form expression for the expected value of the retransmitted packet per cycle results,

$$\eta = \frac{T_1}{T_3} N + \frac{T_2}{T_3} [\{\alpha^{-N}(1-\alpha^{-1})\}\{\frac{\alpha}{(1-\alpha)^2}[N\alpha^N - N\alpha^{N-1} + 1 - \alpha^N] + N\alpha^N\}]$$

$$= \frac{T_1}{T_3} N + \frac{T_2}{T_3} [\frac{\alpha^{-N}}{\alpha(1-\alpha^{-1})} [N(\alpha^N - \alpha^{N-1}) + (1-\alpha^N)] + N(1-\alpha^{-1})]$$

$$= \frac{T_1}{T_3} N + \frac{T_2}{T_3} [\frac{1}{\alpha(1-\alpha)^{-1}} [N(1-\alpha^{-1}) + (\alpha^{-N} - 1)] + N(1-\alpha^{-1})]$$

$$= \frac{T_1}{T_3} N + \frac{T_2}{T_3} [\frac{N}{\alpha} + \frac{(\alpha^{-N} - 1)}{\alpha(1-\alpha^{-1})} + N - N\alpha^{-1}]$$

$$= \frac{T_1}{T_3} N + \frac{T_2}{T_3} [N + \frac{\alpha^{-N} - 1}{(\alpha - 1)}]$$

$$= \frac{T_1}{T_3} N + \frac{T_2}{T_3} [N - \frac{(1-\alpha^{-N})}{(\alpha - 1)}]$$

$$= \frac{T_1}{T_3} N + \frac{T_2}{T_3} [N - \frac{(\alpha^N - 1)}{\alpha^N(\alpha-1)}] \quad 5\text{-}10$$

In the protocol under consideration, the transmission of M packets is conducted in cycles. Each cycle represents the transmission of N packets each of size (B+b) bits plus a time-out interval, t_a. The time required per cycle,

$$T_c = t_a + N(\frac{B+b}{R}) \quad 5\text{-}11$$

we assume that, acknowledgement delay, $t_a = \frac{b}{R}$.
Due to transmission errors under considerations, on the average "η" packets are to be retransmitted per cycle. The total time required to transmit M packets,

$$T_t = \frac{M}{N - \eta} T_c \quad 5\text{-}12$$

But the minimum time required to transmit L bit message at transmission rate R bps, is $\frac{L}{R}$, therefore, the packet delay per message due to channel error is,

$$\tau = T_t - \frac{L}{R}$$

$$= \frac{M}{N-\eta} T_c - \frac{L}{R}$$

$$= \frac{M}{N-\eta} [\frac{b}{R} + N(\frac{B+b}{R})] - \frac{L}{R}$$

$$= \frac{L}{R} [\frac{(N+1)\ b + NB}{B(N-\eta)} - 1] \quad 5\text{-}13$$

This expression contains the parameters, N, B, b, R, L which are constants and therefore the only unknown is the average number of retransmitted packets per cycle "η", which has been stated before in equation (5-10)

Let,

f_D = Doppler frequency shift,

λ = Carrier frequency wavelength

$$\rho = \frac{P_T}{P_R} = \frac{\text{Threshold Power level}}{\text{RMS Power level}}$$

V = Mobile vehicle speed

Then,

$$f_D = \frac{V}{\lambda}$$

The channel parameters T_1, T_2, and T_3 are related to the Doppler frequency shift f_D and the relative power level ρ, by [32, 34]

$$T_1 = \frac{\exp(\rho) - 1}{f_D \sqrt{2\pi\rho}} \qquad 5\text{-}14$$

$$T_2 = \frac{1}{f_D \sqrt{2\pi\rho}} \qquad 5\text{-}15$$

$$T_3 = \frac{\exp(\rho)}{f_D \sqrt{2\pi\rho}} \qquad 5\text{-}16$$

The average delay per message (τ) or the average delay per packet can be calculated using equations (5-13), (5-10), (5-14), (5-15), and (5-16)

From equation (5-3), the probability of success (P_S) for a packet of length T second is,

$$P_S = \frac{T_2}{T_3} \left[\exp - \left(\frac{T}{T_2}\right)\right]$$

Substituting for T_2 and T_3 we get

$$P_S = \exp[-(\rho + f_D\sqrt{2\pi\rho}\ T)] \qquad 5\text{-}17$$

Therefore, the packet error rate is given by,

$$P_E = 1 - \exp[-(\rho + f_D\sqrt{2\pi\rho}\ T)] \qquad 5\text{-}18$$

This expression gives the probability that a packet transmitted over the channel will have at least one detected error.

5.5 Performance Evaluation

The land-mobile data channels are characterized by a high error rate ($10^{-1} - 10^{-3}$) mainly due to the frequent signal fading and the rapid variation of the received signal level. The average signal-to-noise ratio varies considerably (10 - 30 dB) over the relatively small service area resulting in different error rates and error distributions at different geographical locations within the service area [33, 34, 38]. The message delay due to channel error could be minimized by optimizing the packet length for a certain set of system parameters. However, since the system parameters vary from one location to the other, a variable packet length is required to keep the delay at its minimum value. Because of the complexity associated with variable packet length model analysis, a better alternative will be to transmit more than one packet at a time before the transmitting unit stops and waits for the ACK signal. The number of packets transmitted at a time can then be made adaptive to the status of the channel, to keep the message delay close to its minimum value in all locations [37].

Experimental evidence indicates [49] that a maximum packet size of 1000 bits seems to be a satisfactory choice for the vast majority of computer communications requirements. For portable digital terminals (as with real-time computer speech input) packet sizes of a few hundred bits are more than sufficient. To conduct tests with automobiles in the MPRNET ground computer communications a target speed of 100 mph is considered sufficient in the early stage.

To illustrate the behavior of the performance parameters like packet error rate, packet delay and optimum number of packets transmitted per cycle that will minimize delay we choose carrier frequency of 850 MHz. Figure (5-2) shows vehicle speed versus packet error rate (P_B) for a fixed transmission rate and for two different values of the signal-to-noise ratio. It is observed that higher SNR gives better error rate performance, also the performance degrades rapidly with the increasing vehicle speed and after about [illegible] mph, the degradation is slow. Higher packet size and overheads lead to increased error rates. Figure(5-3) shows vehicle speed versus packet error rate (P_B) for a fixed packet size and overhead and again for two values of the signal-to-noise ratio. It is observed that higher SNR is desirable for better error rate performance. Here, again the performance degrades rapidly with the increasing vehicle speed and after about 80 mph, the degradation is slow or almost constant. Higher rates of packet transmission leads to better performance, particularly if the transmission rate is increased from 100 Kbps to 1000 Kbps, the packet error rate drops down by about 27 percent for 10 dB SNR and by about 10 percent for 30 dB SNR. In figure (5-4) we plot vehicle speed versus message delay due to channel error caused by retransmissions. Here, we have chosen a set of typical values for L, R, B, b to study the performance of the model. Here again, higher SNR leads to better performance. We observe that for a certain packet length the message delay is minimized for an optimum value of N, the number of packets transmitted per cycle. For a SNR of 10 dB, N = 1 minimizes the delay while for SNR of 30 dB, N = 2 is the optimal value. For non-optimal values of N, the delay performance is very poor for low SNR and poor for higher SNR. The degradation of delay performance as a function of vehicle speed is more severe for lower SNR than for higher SNR. For 30 dB SNR, the message delay almost stabilizes after 100 mph. It has been established[38] that for a fixed packet length, N (optimum) is roughly proportional to the square root of the average non-fade duration (T_2) of the received signal, and since "T_2" may vary by an order of magnitude over the service area, then N (optimum) may vary by a ratio of 1:3 for different locations within the service area.

5.6 Conclusions

We have developed a model and derived a protocol to analyze mobile radio packet network performance under degraded channel conditions. We came up with the analytical results for evaluating the performance of the protocol in terms of parameters like packet error rate, message delay and number of packets retransmitted per cycle. The behavior of these performance parameters was studied in relation to packet size, packet overhead, transmission rate, signal-to-noise ratio, packets transmitted per cycle and vehicle speed. We observed that the packet error rate performance and message delay performance degrades rapidly with vehicle speed up to 80 mph, but above this speed the degradation is slow for reasonably good SNR, an interesting result indeed. The analysis presented is good for outbound channels. However, for inbound channels we need to take into account errors caused by the random access policy and particularly errors caused by packet overlaps.

Throughout the paper, it was assumed that all terminals are within range and in line-of-sight of each other. A common situation consists of a population of terminals, all within range and communicating with a single "station" (computer center, gate to a network, etc.) in line-of-sight of all terminals.

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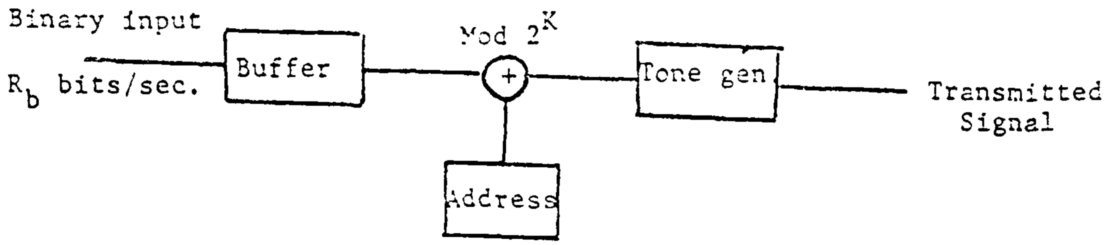


Figure 1-1 Transmitter in FH-MFSK

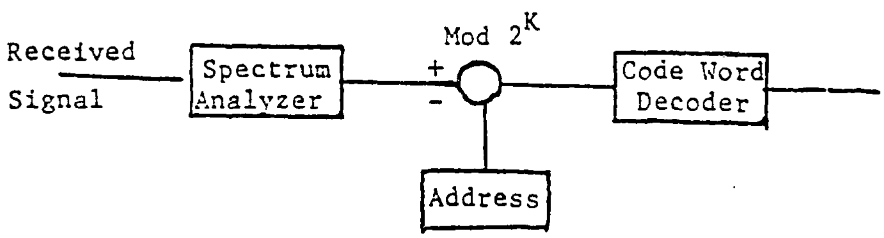


Figure 1-2 Receiver in FH-MFSK

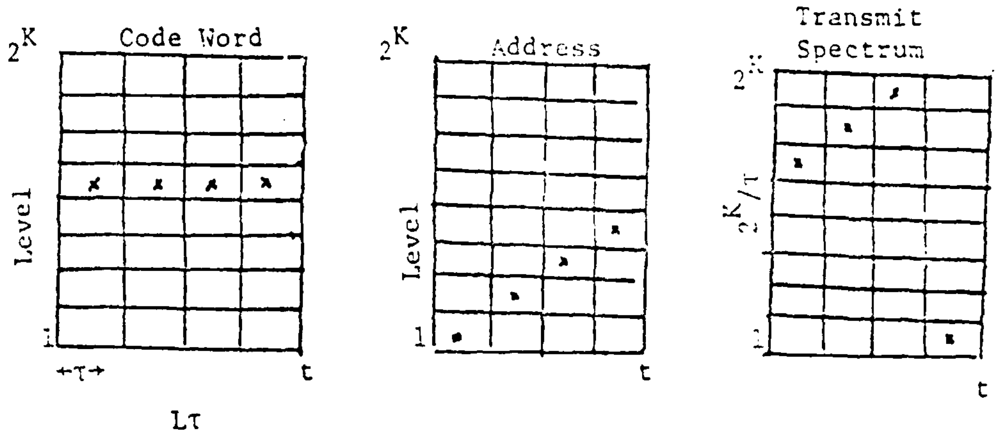


Figure 1-3a   Matrices in Transmit Portion

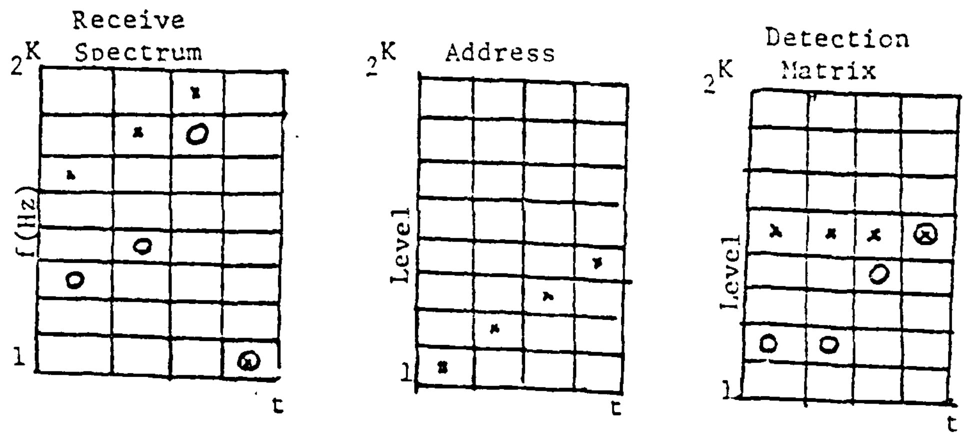


Figure 1-3b   Matrices in Receive portion. The matrices show reception of the frequencies in Figure 1-3a (X) and reception of a set of tones (O) transmitted by one other user.

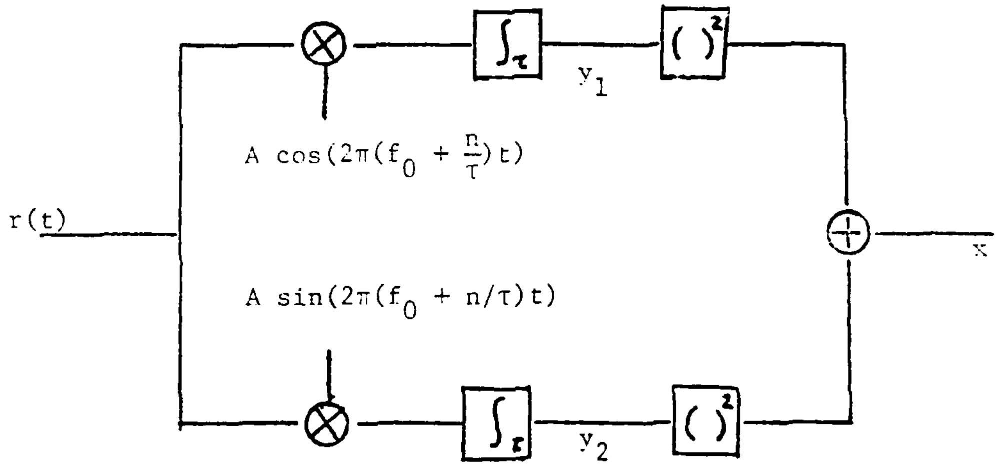


Figure 1-4 Envelope Analyzer

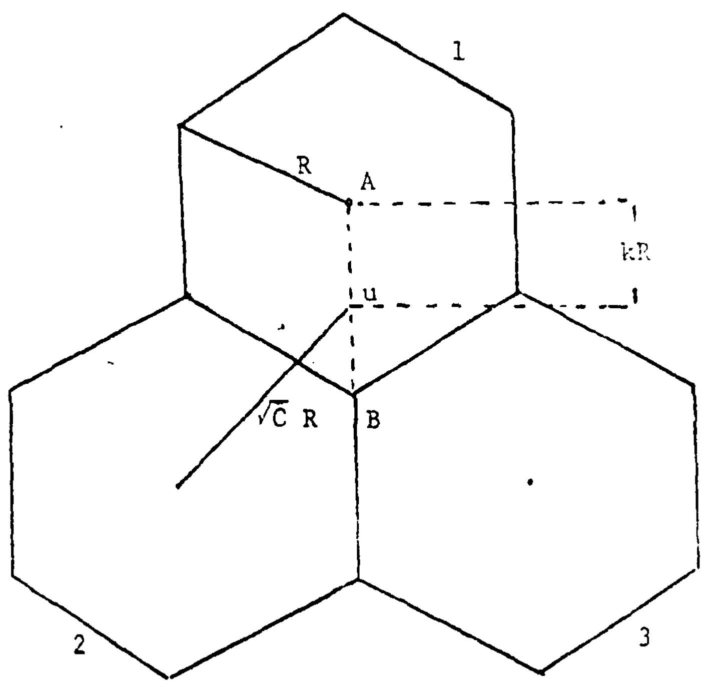


Figure 2-1 Three-cell Geometry

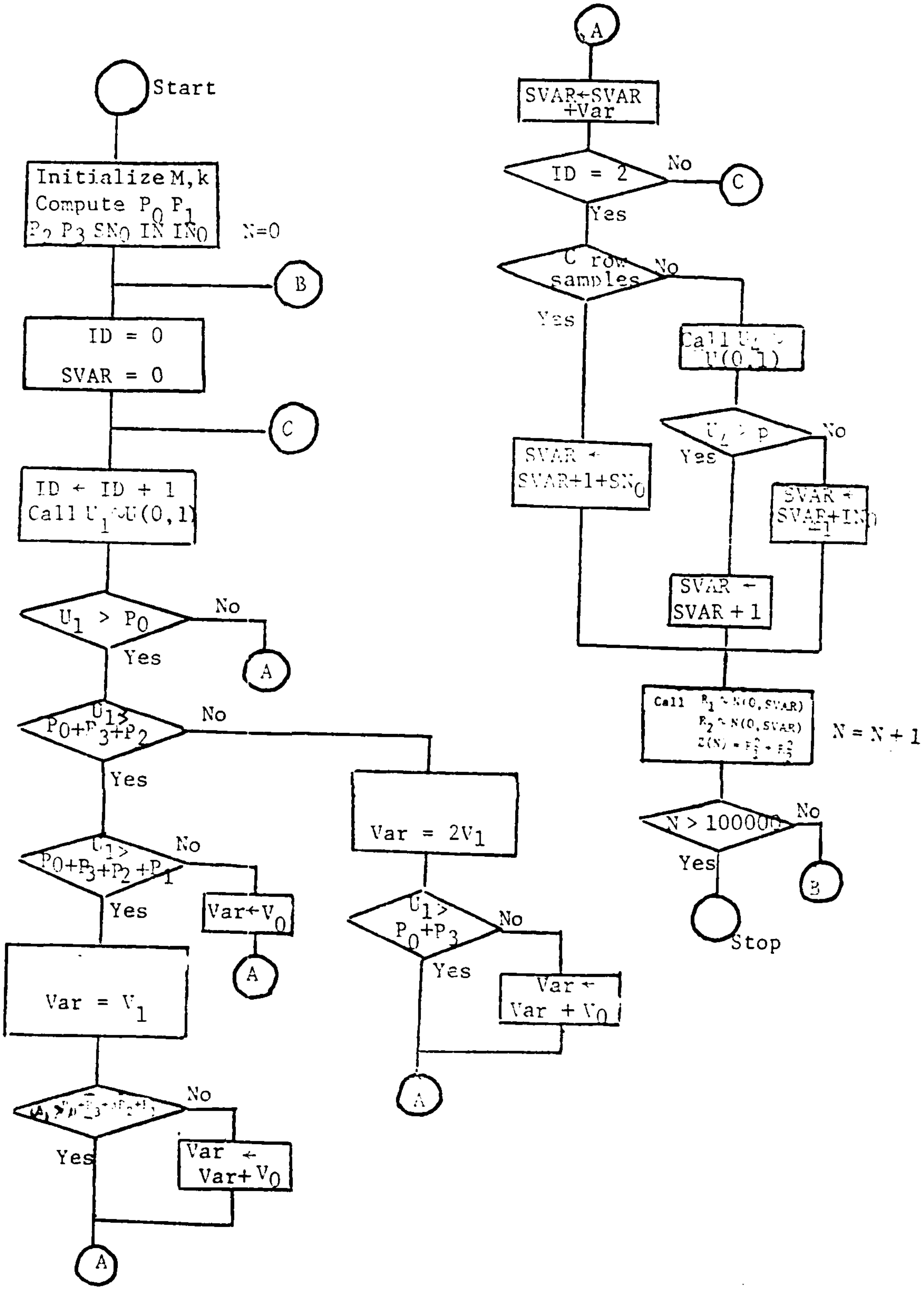
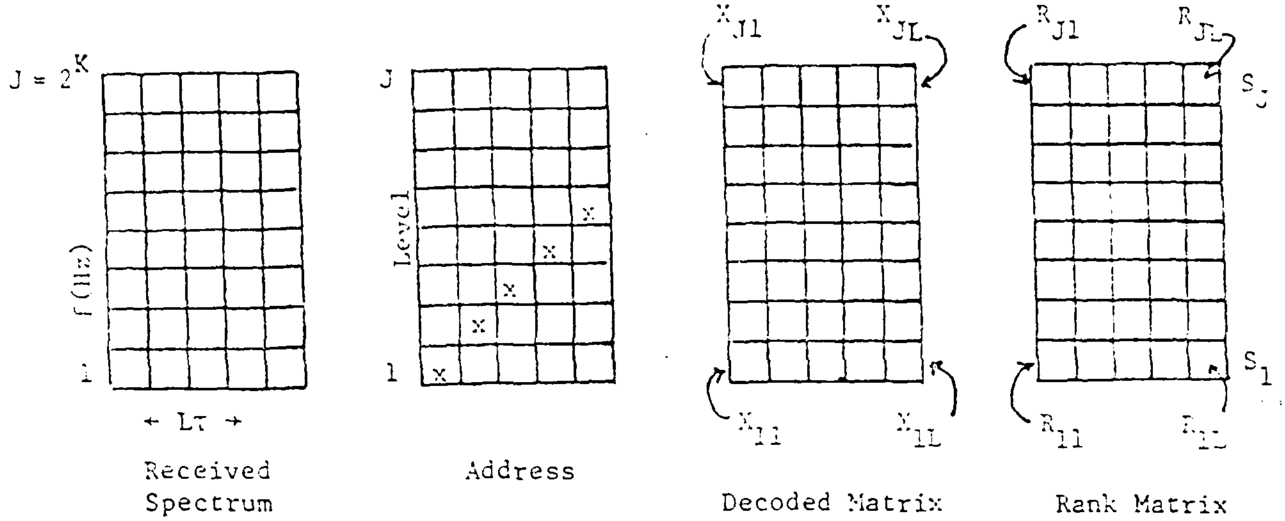


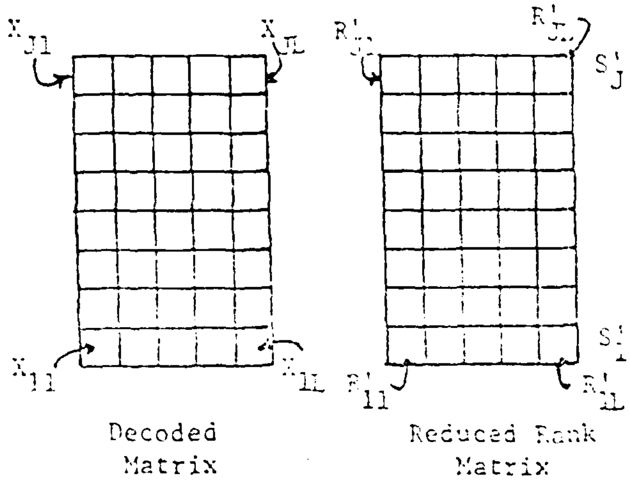
Figure 2-2 Flow diagram to generate the correct and spurious row samples.



$$R_{\ell m} = \text{Rank of } X_{\ell m} \text{ among } \{X_{pq};\ p = 1, 2, \ldots, J;\ q = 1, 2, \ldots, L\}$$

$$S_i = \text{Rank-Sum of } i^{th} \text{ row; } i = 1, 2, \ldots, J$$

Figure 3-1 Maximum Rank Sum Receiver Operation



$$R'_{\ell m} = \text{Rank of } X_{\ell m} \text{ among } \{X_{pm};\ p = 1, 2, \ldots, J\}$$

$$S'_i = \text{Reduced Rank Sum of } i^{th} \text{ row; } i = 1, \ldots, J$$

Figure 3-2 Reduced Rank Sum Receiver Operation

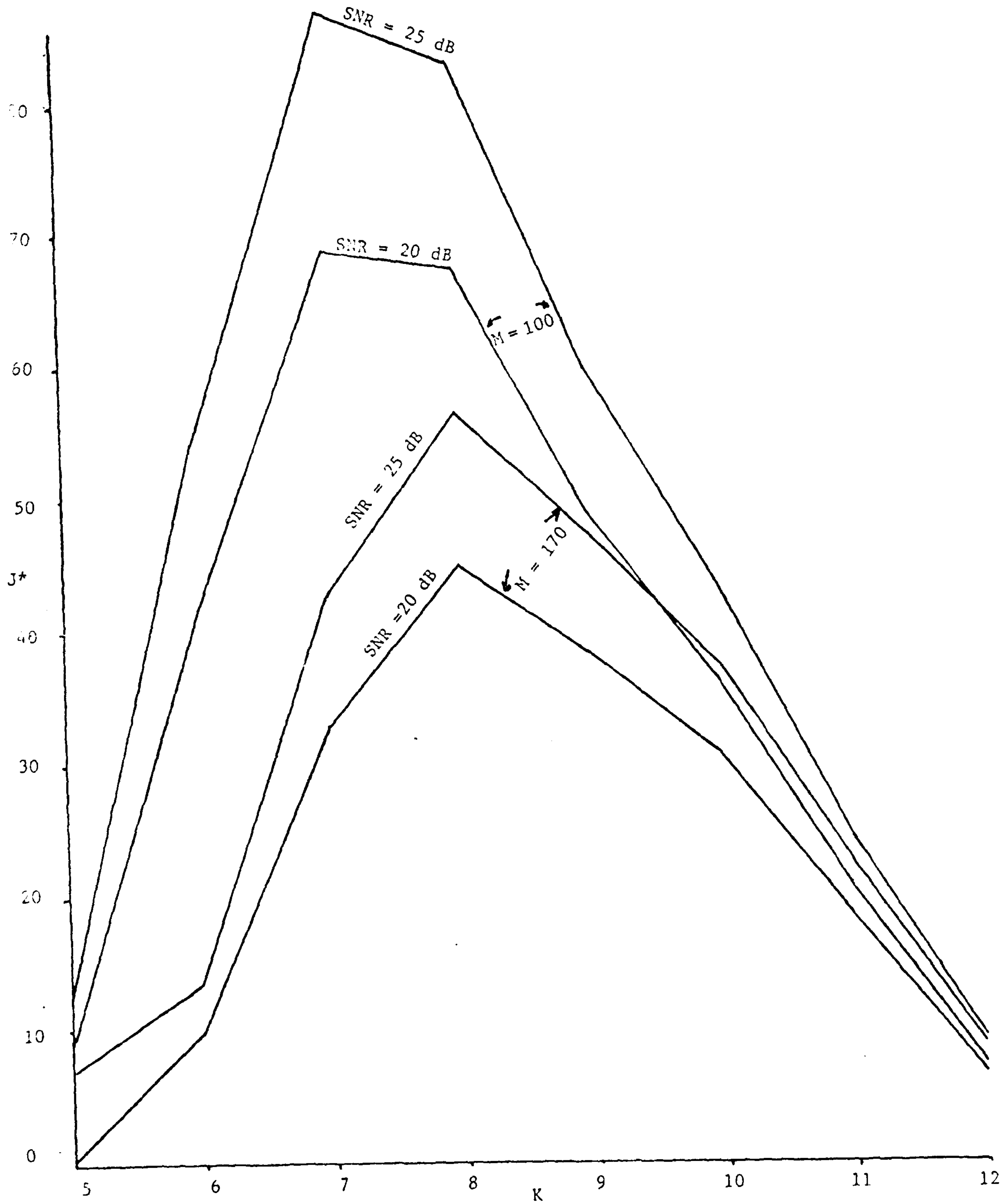


Figure 3-3 Divergence J* vs. K

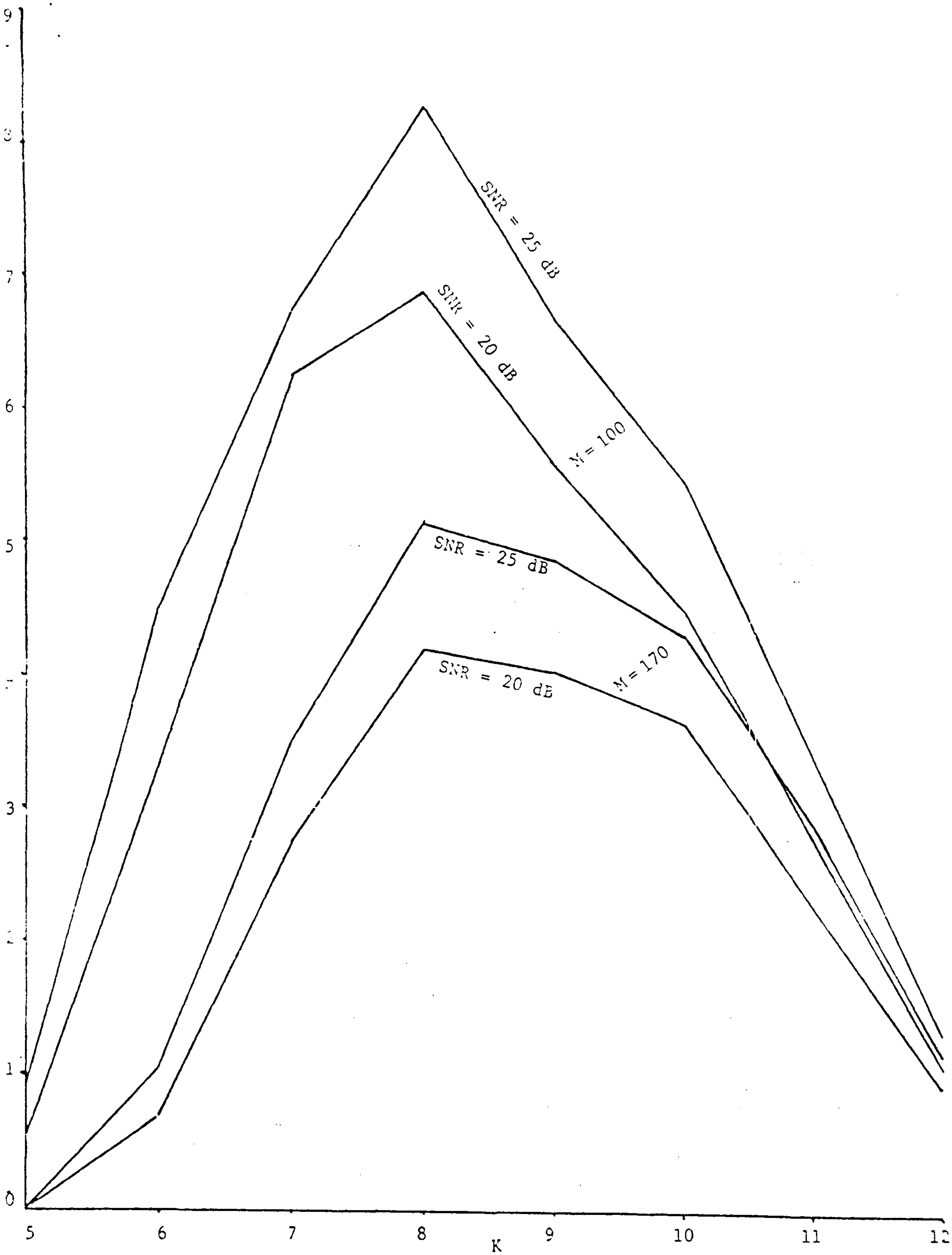


Figure 3-4 Bhattacharyya distance vs. K

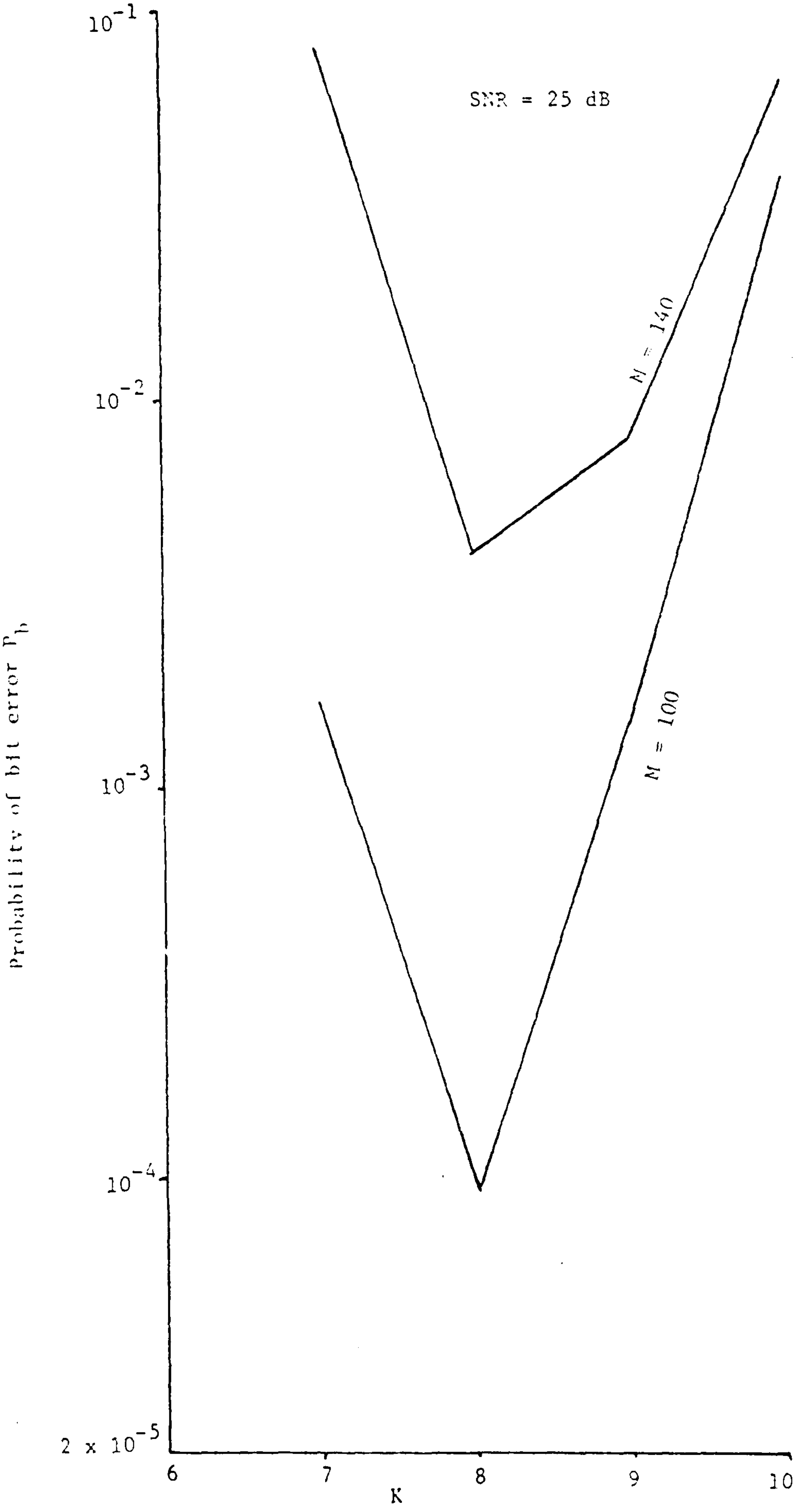


Figure 3-5 Asymptotic Error Estimate vs. K (Maximum Rank Sum Receiver)

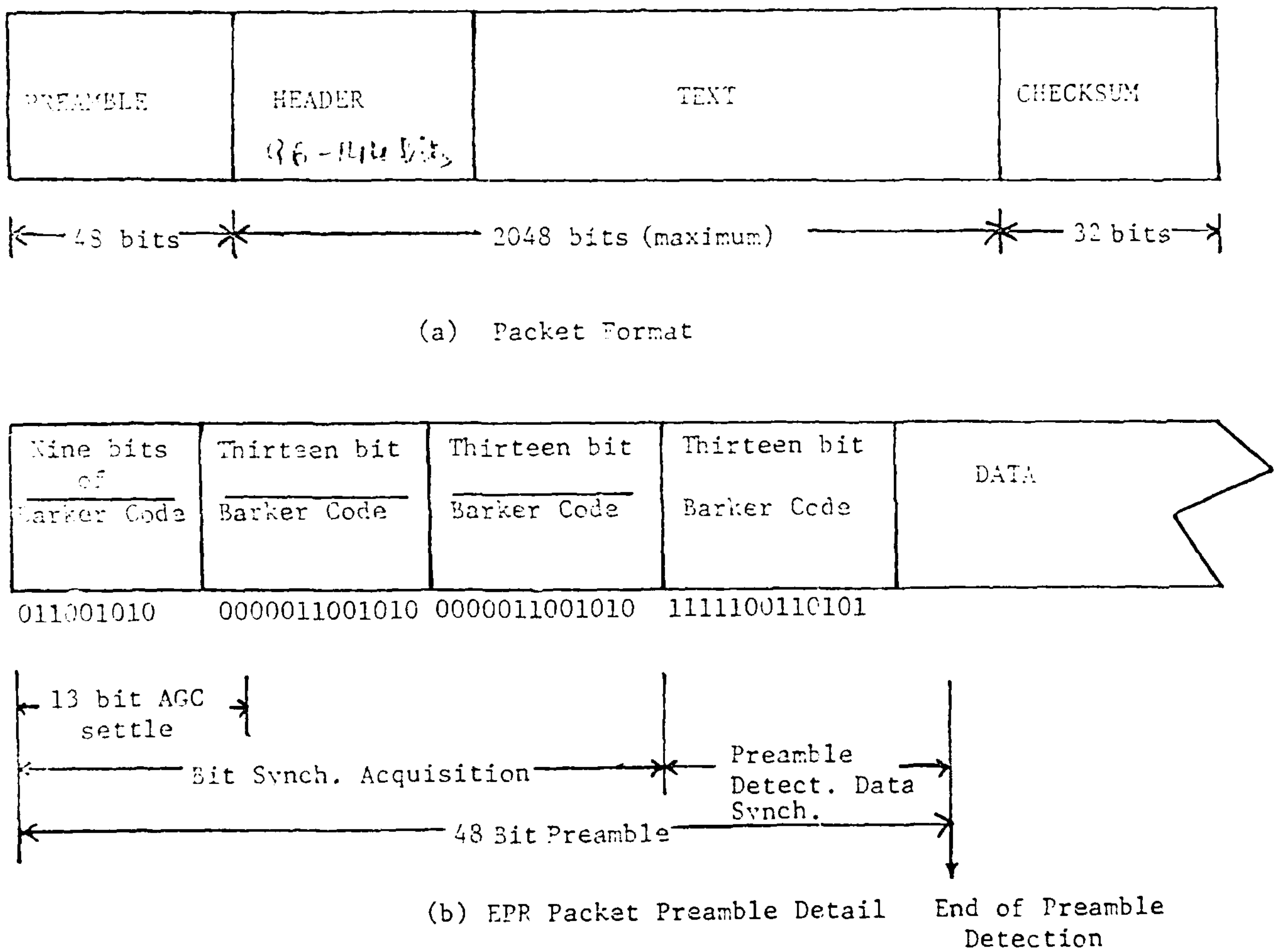


Figure 4-1 Structure of a Transmitted Packet

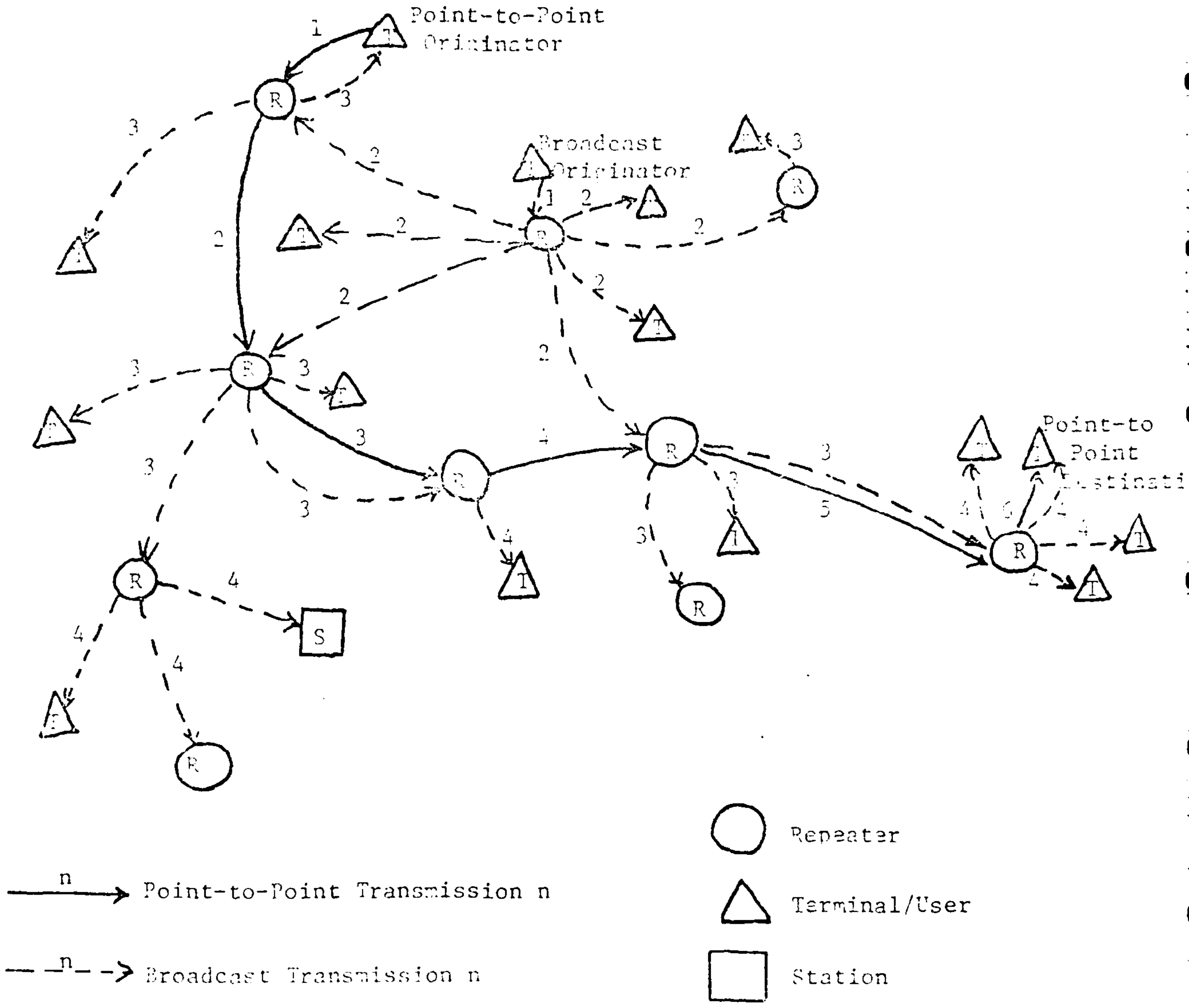


Figure 4-2 Point-to-Point and Broadcast Routing

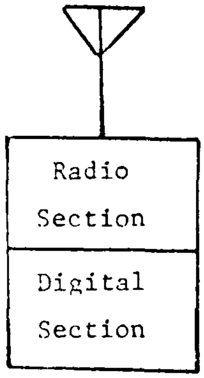


Figure 4-3 Basic organization of Packet Radio Repeater

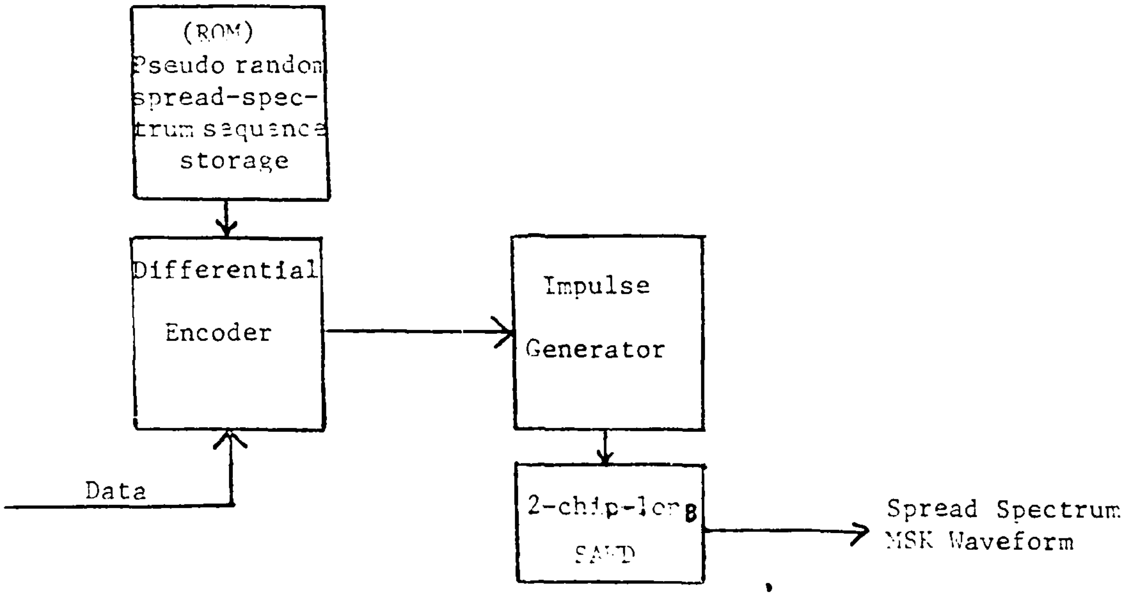


Figure 4-4 Functional Diagram of Encoder/Modulator

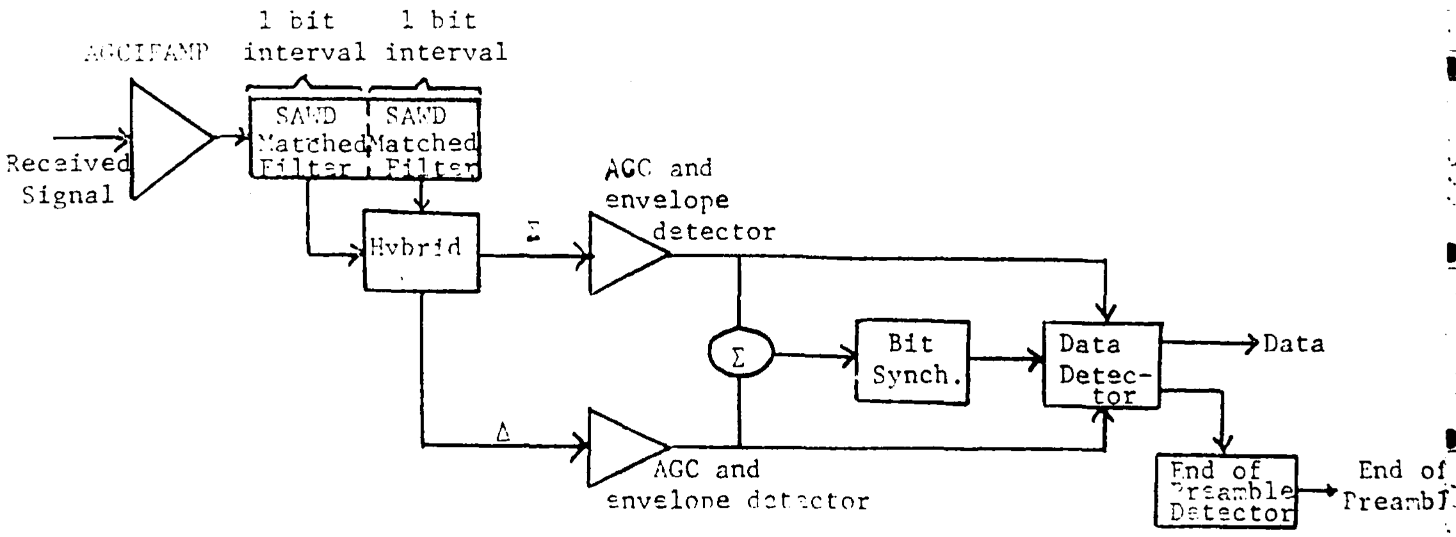


Figure 4-5 Signal Processing Circuits at Receiver

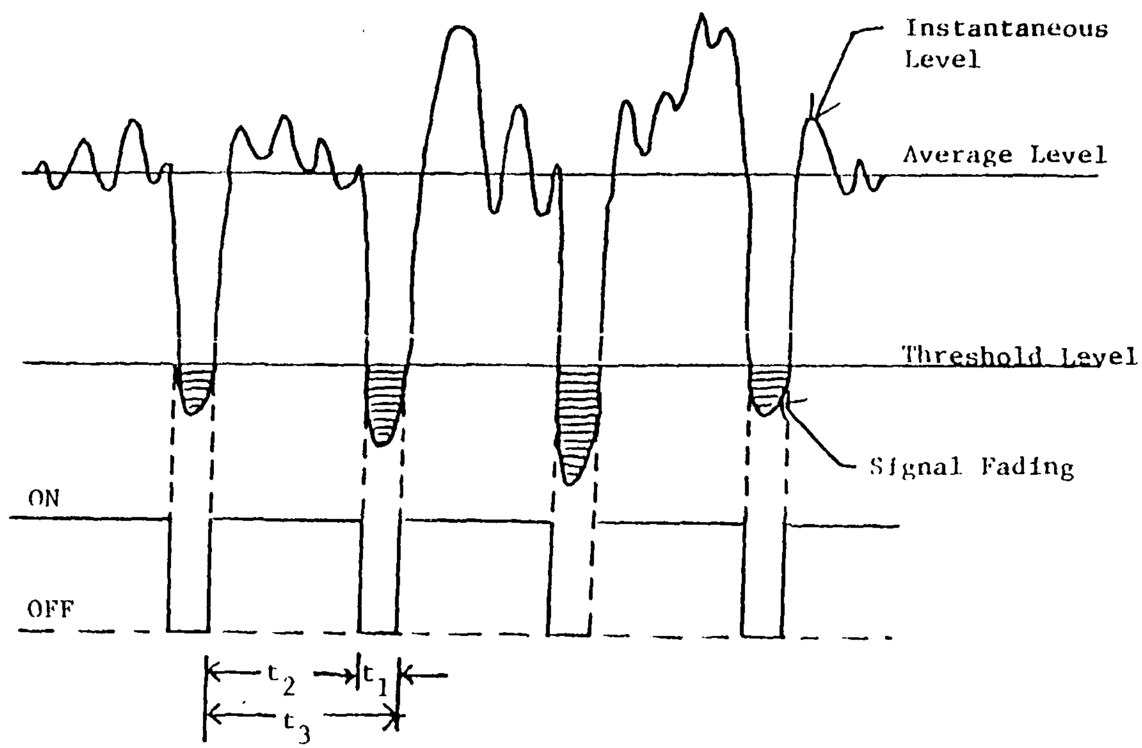


Figure 5-1 ON-OFF characteristics of the land-mobile data channel

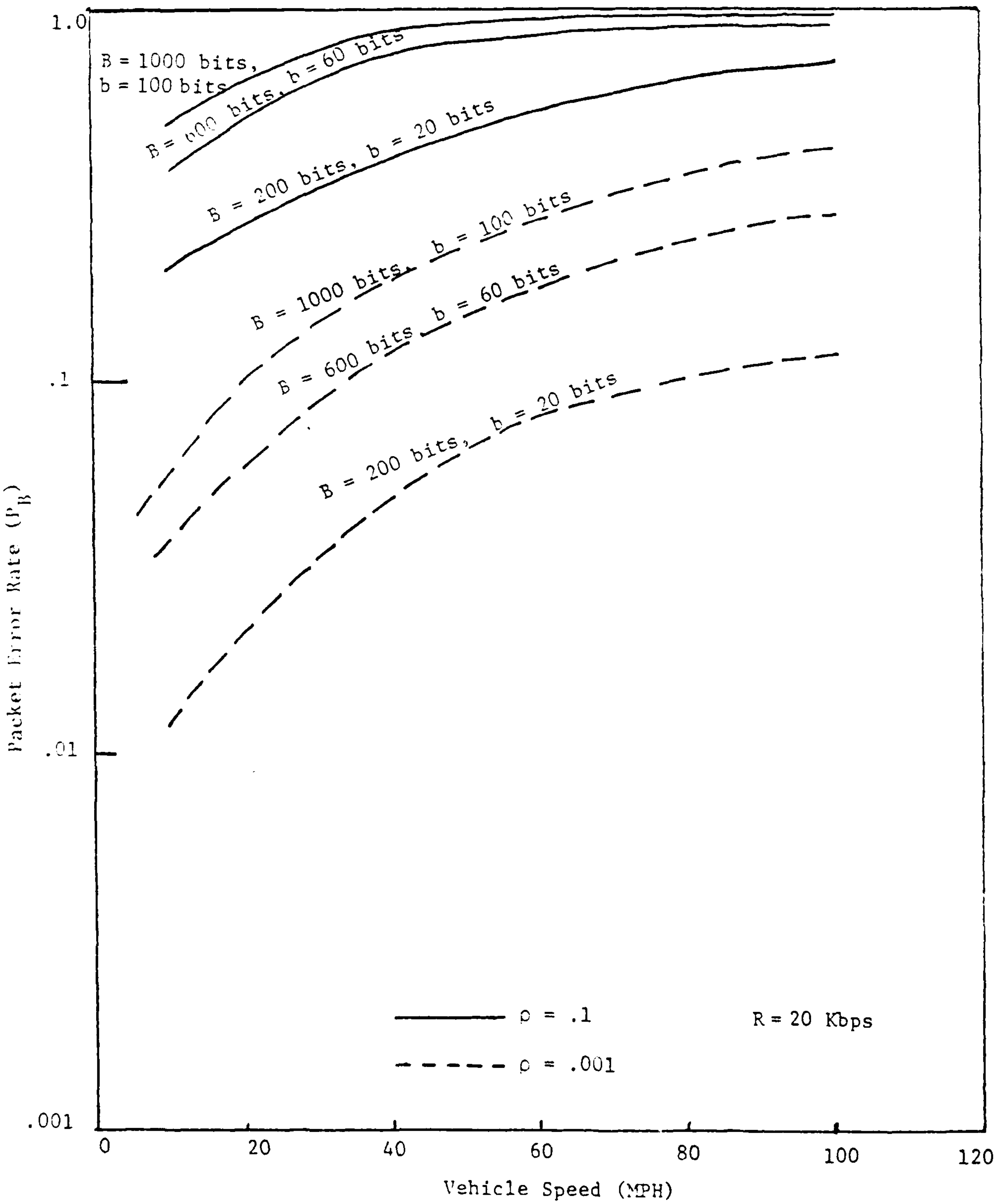


Figure 5-2 Effect of vehicle speed on packet error rate

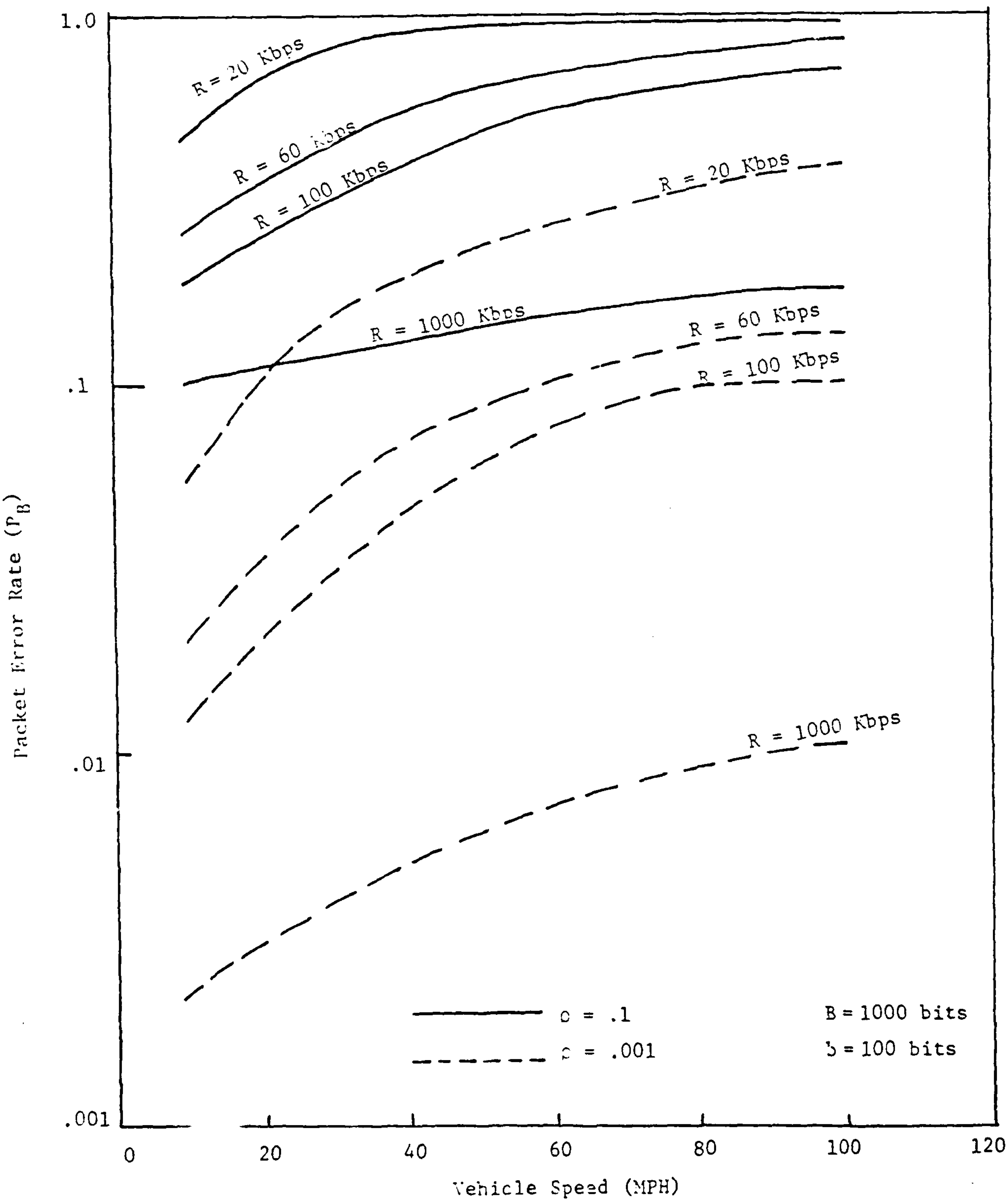


Figure 5-3 Effect of vehicle speed on packet error rate

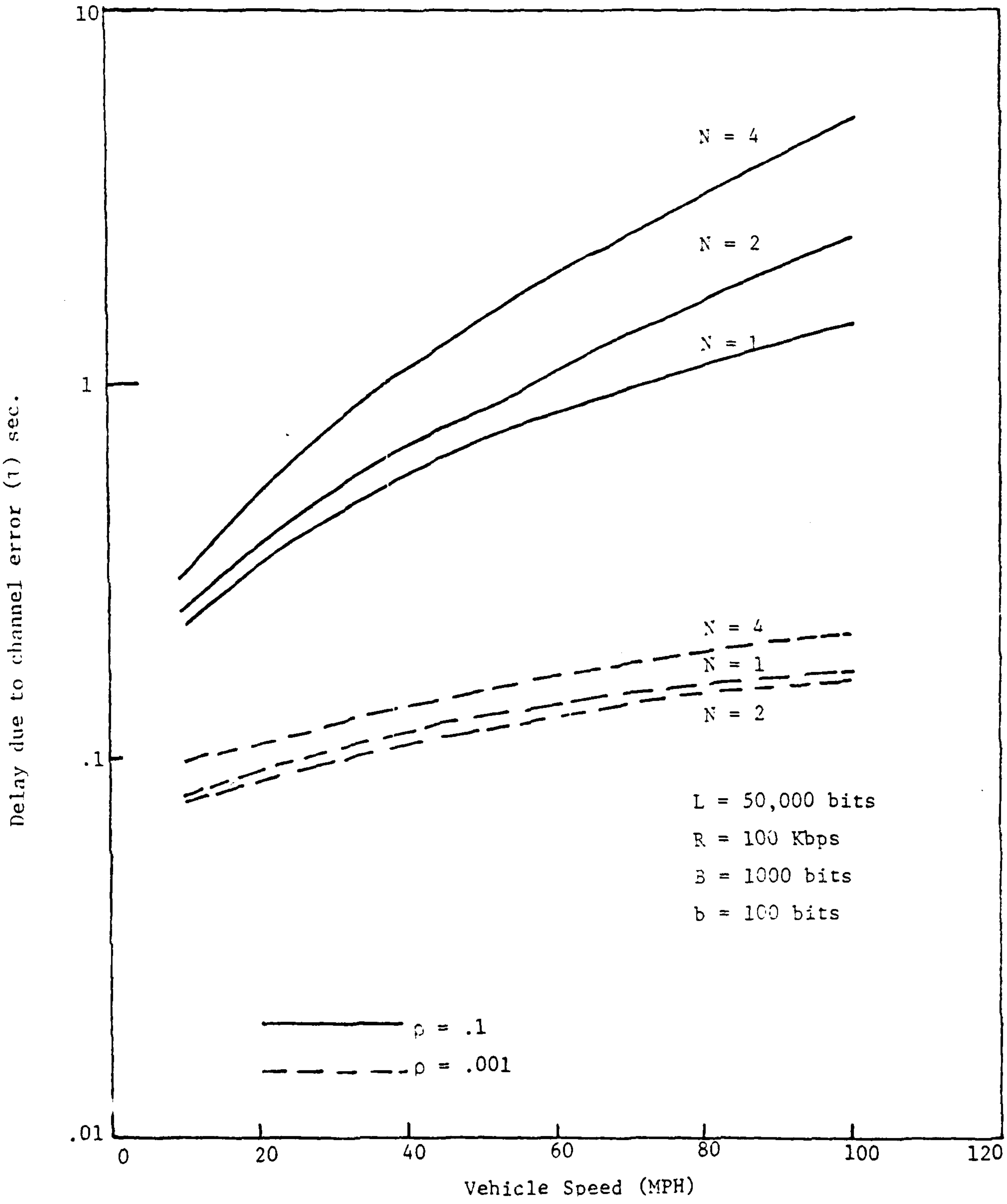


Figure 5-4 Vehicle speed versus delay due to channel error

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