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EXPERIMENTAL PACKET RADIO SYSTEM
DESIGN PLAN

Appendix

13 March 1974

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SYSTEM DESIGN PLAN

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Appendix A.1

Contents: **Routing and Acknowledgment Schemes for the Packet Radio System**;
System Capacity and Data Rate Considerations;
Radiofrequency Channel Capacity Considerations;
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Dynamically Allocated Multiple-Channel Network Concept;
Notes on Radar Test;
Modulation Waveform Types;
Synchronization Preambles;
Impact of Channel Options on Repeater Design;
Power Budget Analysis,
Powers Sources for Repeaters and Terminals;
Criteria of Code Selection in Minimum Shift Keyer Sense; and
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1. INTRODUCTION

In this chapter we discuss routing problems for broadcast oriented packet communication networks. Possible solutions to these problems are described and an approach, tested by simulation, is proposed for system operation.

There are basic differences between the packet radio network and existing point-to-point store and forward networks, such as the ARPANET. For example, the packet radio network serves mobile terminals; devices in the network share a common channel in a random access broadcast mode; and repeaters in this network will have significantly less storage and processing capabilities than the switching nodes in ARPANET like systems. Consequently, many of the routing techniques developed for the point-to-point networks are not directly applicable to the packet radio network.

The objective of the network is to distribute and collect traffic to and from terminals which have high ratios of peak to average traffic requirements. A primary initial goal of the packet radio system is to serve as a local distribution system for traffic destined for the ARPANET. for mobile sources.

The network consists of repeaters to provide area coverage and stations to provide traffic management and interfaces to other nets. Stations serve as a major source and sink for the packet radio net. There are many possible paths via repeaters over which a packet originating at a terminal may flow to reach a station. That is, a packet transmitted from a terminal can be received by several repeaters, and there may be several stages of transmission through repeaters before the packet is received by a station.

Some problems that arise in controlling traffic flow in a large scale broadcast network are:

(1) A packet transmitted can be received by many repeaters or stations or not be received by any.

(2) Many copies of the same packet can circulate in the broadcast network .

(3) Many copies of the same packet can enter the point-to-point network at different stations.

Indications of the consequences of not imposing a suitable flow control mechanism can be observed from combinatorial models analyzed in Chapter 11. In these ideal models, the repeaters are located at corner points of an infinite square grid and time is broken into unit intervals, each slotted into segments. A packet transmitted by a repeater can be received only by its four nearest neighbors. If a packet is correctly received by a repeater, it is retransmitted within the next unit interval of time at a random time slot within the interval. Suppose now that a single packet originates at the origin and that the transmission plus the propagation time falls within one unit interval of time. Then after n intervals of time:

(i) the number of repeaters which receive the packet for the first time, $B(n)$, is:

$$B(n) = 4n, \quad n \geq 1 \quad B(0) = 1$$

(ii) the number or repeaters through which the packet passed,

$A(n)$, is:

$$A(n) = \sum_{j=0}^{n} B(j) = 2n^2 + 2n + 1, \quad n \geq 0$$

(iii) if we assume that a repeater can receive and relay a large number of packets within the same time interval, the number of copies of the same packet received by a repeater at coordinates (d,j) after d + 2k units of time is:

$$N_j^d \; (d + 2k) = \binom{d + 2k}{k + j} \binom{d + 2k}{k} \xrightarrow{\text{for large } k} 2^{4k}$$

where d is the number of units of time that the packet requires to arrive from the origin to the repeater, and j is the horizontal number of units.

Unless adequate steps are taken, the explosive proliferation of redundant packets will severely limit the capacity of the system. One can now recognize two somewhat distinct routing and control problems:

(1) to ensure that a packet originating from a terminal arrives at a station, preferably using the most efficient (shortest) path; and

(2) to suppress copies of the same packet from being indefinitely repeated in the network, either by being propagated in endless cycles of repeaters or by being propagated for a very long distance.

In Section 2, we outline general techniques which can be combined to provide workable routing schemes. Acknowledgement schemes aimed at achieving high throughput and minimum delay are discussed in Section 3. Section 4 gives a detailed description of an efficient routing scheme. In Section 5, a method for repeater labeling to obtain efficient routing is proposed. Finally, some qualitative properties of the routing scheme proposed in Section 4, generated via a detailed simulation are given in Section 6.

2. POSSIBLE ROUTING TECHNIQUES

There are two key objectives in developing a routing procedure for the packet radio system. First, we must assure, with high probability, that a message launched into the net from an arbitrary point will reach its destination. Second, we must guarantee that a large number of messages will be able to be transmitted through the network with a relatively small time delay. The first goal may be thought of as a connectivity or reliability issue, while the second is an efficiency consideration.

A rudimentary, but workable, routing technique to achieve connectivity at low traffic levels can be simply constructed by using a maximum handover number [Boehm & Baron, 1964] and saving unique identifiers of packets at each repeater for specified periods of time. The handover number is used to guarantee that any packet cannot be indefinitely propagated in the net. Each time a packet is transmitted in the net, a handover number in the header is incremented by one. When the handover number reaches an assigned maximum, the packet is no longer repeated and that copy of the packet is dropped from the net. Thus, the packet is "aged" each time it is repeated until it reaches its destination or is dropped because of excessive age.

If the maximum handover number is set large, extensive artificial traffic may be generated in areas where there is a high density of repeaters. On the other hand, if it is set small, packets from remote areas may never arrive at stations. This problem can be resolved as follows: We assume that every repeater can calculate its approximate distance in numbers of hops to stations by observing response packets. (A labeling technique for this calculation is discussed in Section 5). The first repeater which received the packet from a terminal sets the maximum handover number based on its calculated distance from the station. The number is then decremented by one each time it is relayed through any other repeater. The packet is dropped when the number reduces to zero.

When a station transmits a packet, it will set the maximum handover number by "knowing" the approximate radius in "repeaters" in its region.

Even if a packet is dropped after a large number of transmissions, local controls are needed to prevent packets from being successively "bounced" between two or a small number of repeaters which repeat everything they correctly receive. (Such a phenomena is called "cycling" or "looping.") A simple mechanism to prevent this occurrence is for repeaters to store for a fixed period of time entire packets, headers, or even a field within the header that uniquely identifies a packet. A repeater would then compare the identifier of any received packet against the identifiers in storage at the repeater. If a match occurred, the associated packet would not be repeated.

The time allotted for storage of any packet identifier would depend on the amount of available storage at a repeater and the number of bits required to uniquely identify the packet. For example, more than 4K packets could be uniquely identified with 12 bit words. Thus, 4K of storage could contain identifiers for more than 300 packets. With a 500 Kbps repeater to repeater common channel for broadcast and receive and 1,000 bit packets, this would be sufficient storage for over 1.5 seconds of transmission if the channel were used at full rate. Assuming a single hop would require about 20 milliseconds of transmission and retransmission time, a maximum hop number of 20 would guarantee that any packet would be dropped from the system because of an excessive number of retransmissions long before it could return to a previously used repeater not containing the packet identifier.

The combination of loop prevention and packet ageing with otherwise indiscriminate repetition of packets by repeaters will guarantee that a packet travels, on every available path, a maximum distance away from its origin equal to its original handover number.

Thus, if the maximum handover number is larger than the minimum number of hops between the terminal and the nearest station, a packet accepted into the net should reach its destination. Unfortunately, with this scheme, copies of the packet will also reach many other points, with each repetition occupying valuable channel capacity. However, if those packets for which adequate capacity is not available are prevented from entering the net, the network will appear highly reliable to accepted packets.

The above routing scheme is an undirected, completely distributed procedure. Each repeater is in total control of packets sent to it, and the stations play no active part in the system's routing decisions. (They must still play a role in flow control.) In the above procedure, no advantage is taken of the fact that most traffic is destined for a station, either as a terminus or as an intermediate point for communication with the ARPANET. Also, the superior speed and memory space of the station is ignored. For efficiency, one is therefore led to investigate directed (hierarchical) routing procedures.

A directed routing procedure utilizes the stations to periodically structure the network for efficient flow paths. Stations periodically transmit routing packets called labels to repeaters to form, functionally, a hierarchical point-to-point network. Each label includes the following information: (i) a specific address of the repeater for routing purposes, (ii) the minimum number of hops to the nearest station, and (iii) the specific addresses of all repeaters on a shortest path to the station. In particular, the label contains the address of the repeater to which a packet should preferably be transmitted when destined to the station.

When relaying a packet to its destination, the repeater addresses the packet to the next repeater along the preferred path. Only this addressed repeater will repeat the packet and only when this mechanism fails will other repeaters relay the message. A detailed description of the directed routing technique proposed is given in Section 4. However, we first discuss acknowledgement structures for message flow since good acknowledgement schemes are an integral part of an efficient routing procedure.

3. ACKNOWLEDGEMENT CONSIDERATIONS

Acknowledgement procedures are necessary both as a guarantee that packets are not lost within the net and as a flow control mechanism to prevent retransmissions of packets from entering the net. Two types of acknowledgements are common in packet oriented systems:

1. Hop-by-Hop Acknowledgements (HBH Acks) are transmitted whenever a packet is received successfully by the next node on the transmission path.

2. End-to-End Acknowledgements (ETE Acks) are transmitted whenever a packet correctly reaches its final destination within the network.

In a point-to-point oriented network such as the ARPANET, HBH Acks are used to transfer responsibility (and thus open buffer space) for the packet from the transmitting node to the receiving node. This Ack insures prompt retransmission should parity errors or relay IMP buffer congestion occur. The ETE Ack serves as a flow regulator between source and destination and as a signal to the sending node that the final destination node has correctly received the message. Thus, the message may be dropped from storage at its origin.

Both types of Ack's serve to ensure message integrity and reliability. If there is a high probability of error free transmission per hop and the nodes have sufficient storage, the Hop-by-Hop scheme is not needed for the above purpose. Without an HBH Ack scheme, one would retransmit the packet from its origin after a time out period expired. One introduced the HBH Ack to decrease the delay caused by retransmissions at the expense of added overhead for acknowledgements. In the ARPANET, this added overhead is kept small by "piggybacking" acknowledgements whenever possible on information packets flowing in the reverse direction. In the packet radio system, the overhead can be kept small by listening, whenever possible, for the next repetition of the packet on the common channel instead of generating a separate acknowledgement packet.

The value of an End-to-End acknowledgement is sufficiently great that it can be assumed present a priori. However, the additional use of a Hop-by-Hop acknowledgement is not as clear. Therefore, in this section, we examine the question of whether the ETE Ack is sufficient, or whether one needs a Hop-by-Hop (HBH) acknowledgement in addition. The problem is therefore whether an HBH Ack is superior to an ETE Ack with respect to throughput and delay, since the ETE Ack ensures message integrity. It is noted that the routing and flow control by devices in the network depend on the type of acknowledgement scheme used.

We consider a simple case where (n-1) repeaters separate the packet radio terminal from the destination station. Assuming that the terminal is at a distance of "one hop" from the first repeater, one obtains the following n-hop system:

$$T \xrightarrow{\text{hop } 1} R_1 \xrightarrow{\text{hop } 2} R_2 \cdots \rightarrow R_{n-1} \xrightarrow{\text{hop } n} S$$

A simple model is used to evaluate the total average delay that a packet encounters in the n-hop system when using HBH and ETE acknowledgement schemes. When the ETE acknowledgement scheme is used, every repeater transmits the packet a single time. If the packet does not reach the station, retransmission is originated by the terminal. The ETE acknowledgement is sent from the station. In the HBH scheme, repeaters store and retransmit the packet until positively acknowledged from the next repeater stage.

If, after a terminal (or a repeater in the HBH case) transmits the packet, an acknowledgement does not arrive within a specified period of time, it retransmits the packet. The waiting period is composed of the time for the acknowledgement to arrive when no conflicts occur plus a random time for avoiding repeated conflicts.

Two different schemes for ETE acknowledgement and one scheme for HBH acknowledgement are studied. Curves for the total average delay as a function of the number of hops and the probability of successful transmission per hop are obtained. Two cases are considered: One in which the probability of success is constant along the path and another in which the probability of success decreases linearly as the packet approaches the station. Finally, channel utilizations are compared when using ALOHA [Abramson; 1970, 1973] random access modes of operation.

It is demonstrated that the HBH scheme is superior in terms of delay or channel utilization. This conclusion becomes significant when the number of hops increases or when the probability of successful transmission is low. For example, in a five hop system, if the probability of success per hop is 0.7, then the total average delay is 12.5 and 53 packet transmission times for the HBH and ETE acknowledgement schemes, respectively.

The model used is based on [Kleinrock & Lam, 1973; Roberts]. The model is simplified, however, by assuming that the probability that a packet is blocked is the same when the packet is new or has been blocked any number of times before. Although the more general equations could have been written, the numerical solution is rather elaborate [Kleinrock & Lam, 1973] and seems unnecessary for this comparative study. It is further assumed that the probabilities of being blocked on different hops are mutually independent. The "total delay" is defined as the time between the transmission of the first bit by the terminal and the correct reception of the last bit of the packet by the station.

3.2 DELAY CONSIDERATIONS

The delay equations, normalized by the number of packet transmission times, are given by:

$$D(HBH) = (1 + \beta) \cdot n + (1+2\beta+\alpha+\delta)\left(\sum_{i=1}^{n} \frac{1-q_i}{q_i}\right) \tag{1}$$

$$D_1(ETE) = (1+\beta)\cdot n + [(1+2\beta+\alpha)\cdot n+\delta]\left(\frac{1-Q}{Q}\right) \tag{2}$$

$$D_2(ETE) = (1+\beta)\cdot n + (1+2\beta+\alpha+\delta)\left(\frac{1-Q}{Q}\right). \tag{3}$$

In these equations, α is the ratio of the acknowledgement transmission time to the packet transmission time; β is the ratio of the average propagation time per hop to the packet transmission time; and δ is the ratio of the average waiting time (beyond the minimum) for avoiding repeated conflicts, to the packet transmission time. The quantity q_i is the probability of successful transmission on hop i; and $Q = \prod_{i=1}^{n} q_i$.

As indicated before, two different cases for the ETE acknowledgement are considered. D_1(ETE) represents the delay when the terminal waits the expected time for the packet to reach the station and for the ETE acknowledgement to be received by the terminal before retransmitting the packet. D_2(ETE) is for the case in which the terminal retransmits after shorter periods of time because it anticipates a low probability of successful transmission Q. In particular, we examine the case in which the retransmission delay is the same as in the HBH method.

Figures 1, 2, and 3 show delay curves for the three acknowledgement schemes using the parameters $\alpha = 0.5$, $\beta = 0.02$, and $\delta = 2.0$. Figures 1 and 2 are for the case in which q is constant along the path.

The curves show the delay as a function of the probability of successful transmission q rather than the channel utilization. Thus, they can be used for slotted or non-slotted ALOHA, or possibly for other access schemes.

It is evident from Figure 1 that the delays for the ETE acknowledgement schemes grow much more rapidly than delays for the HBH scheme. For example, in the 5-hop system, if the packet transmission

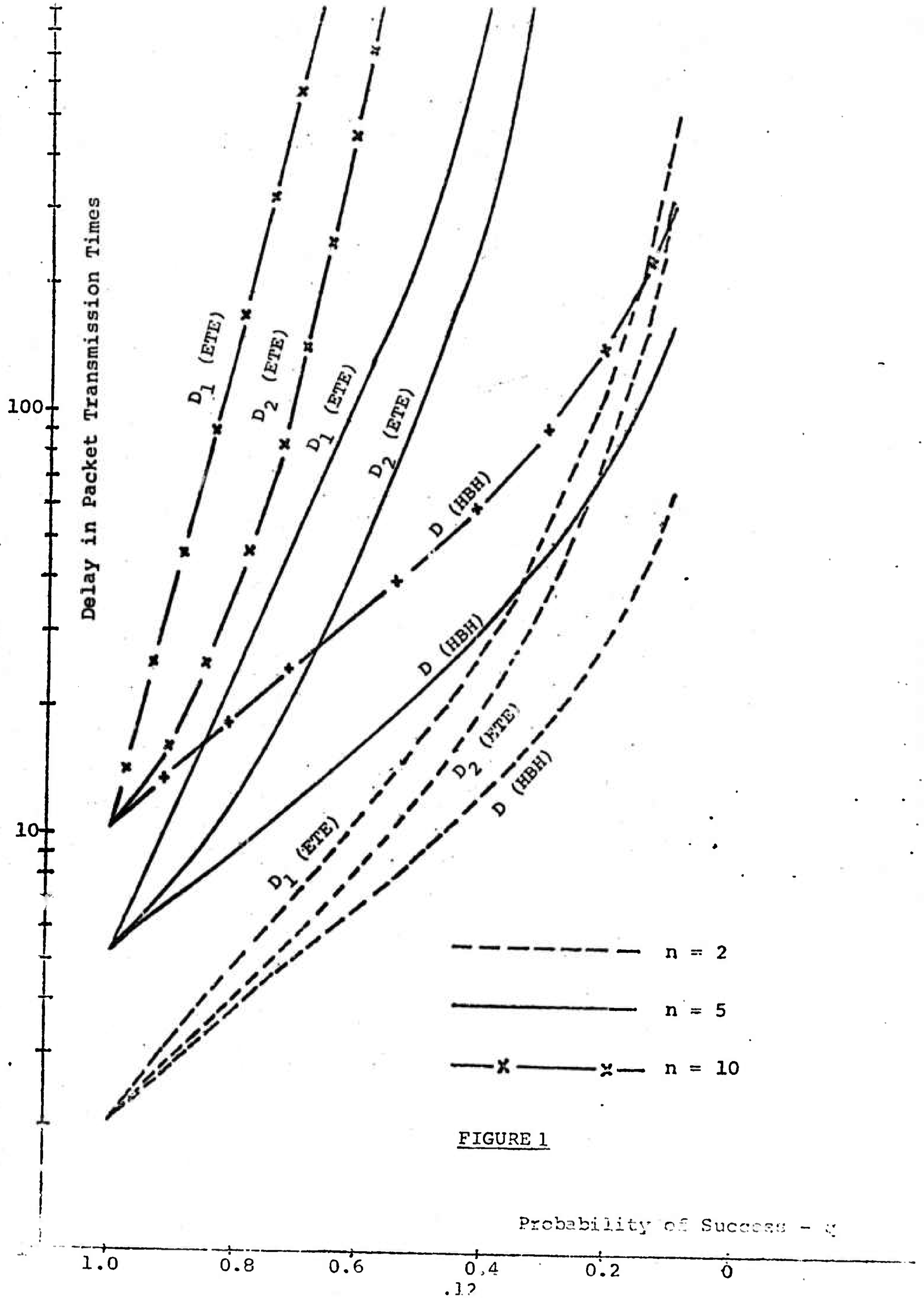
Delay in Packet Transmission Times
100
10
1.0
0.8
0.6
0.4
0.2
0
Probability of Success - q
D1 (ETE)
D2 (ETE)
D (HBH)
n = 2
n = 5
n = 10

FIGURE 1

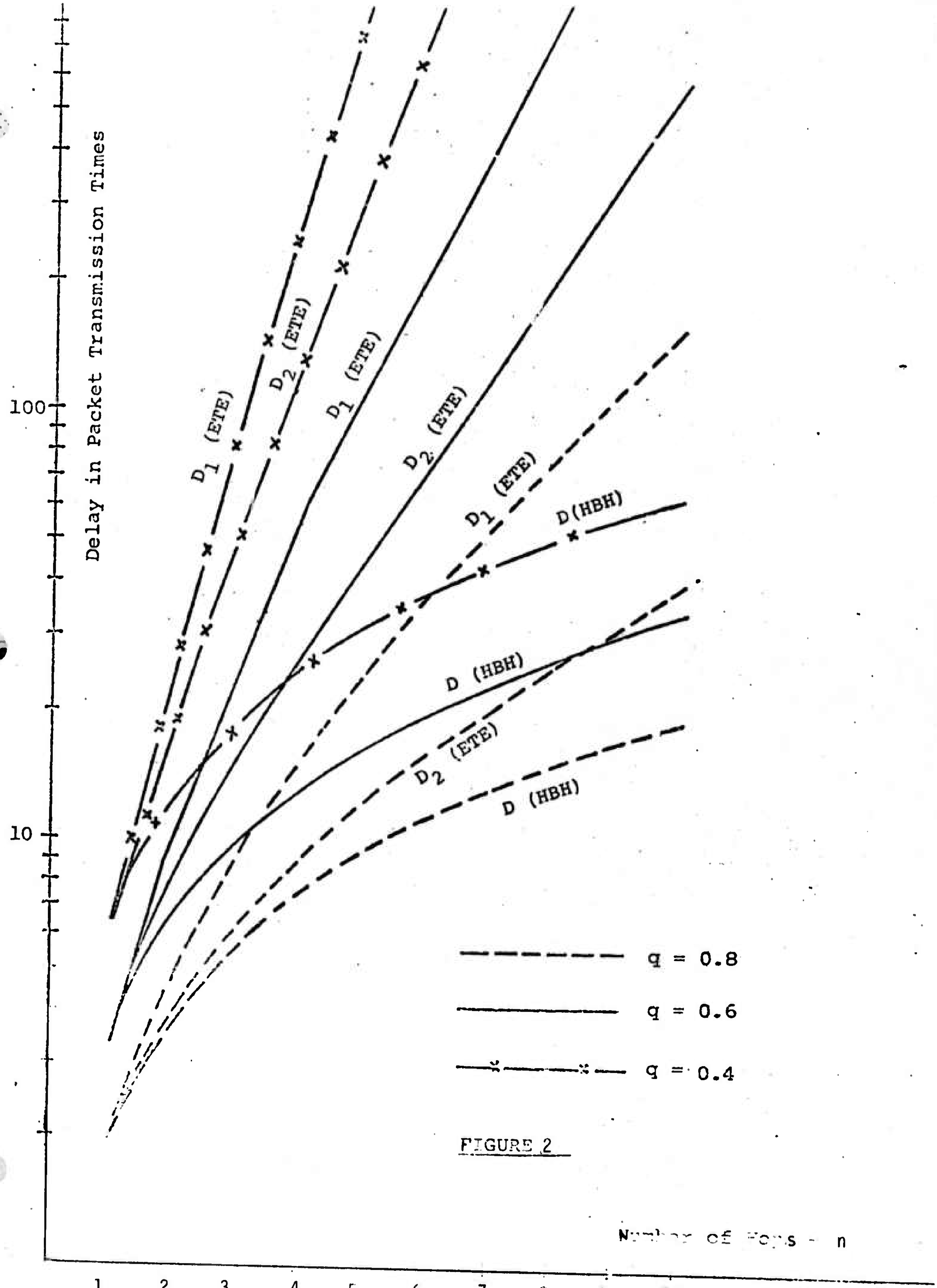
Delay in Packet Transmission Times
100
10
D_1 (ETE)
D_2 (ETE)
D_1 (ETE)
D_2 (ETE)
D_1 (ETE)
D(HBH)
D (HBH)
D_2 (ETE)
D (HBH)
q = 0.8
q = 0.6
q = 0.4
1
2
3
4
5
6
7
8
9
10

FIGURE 2

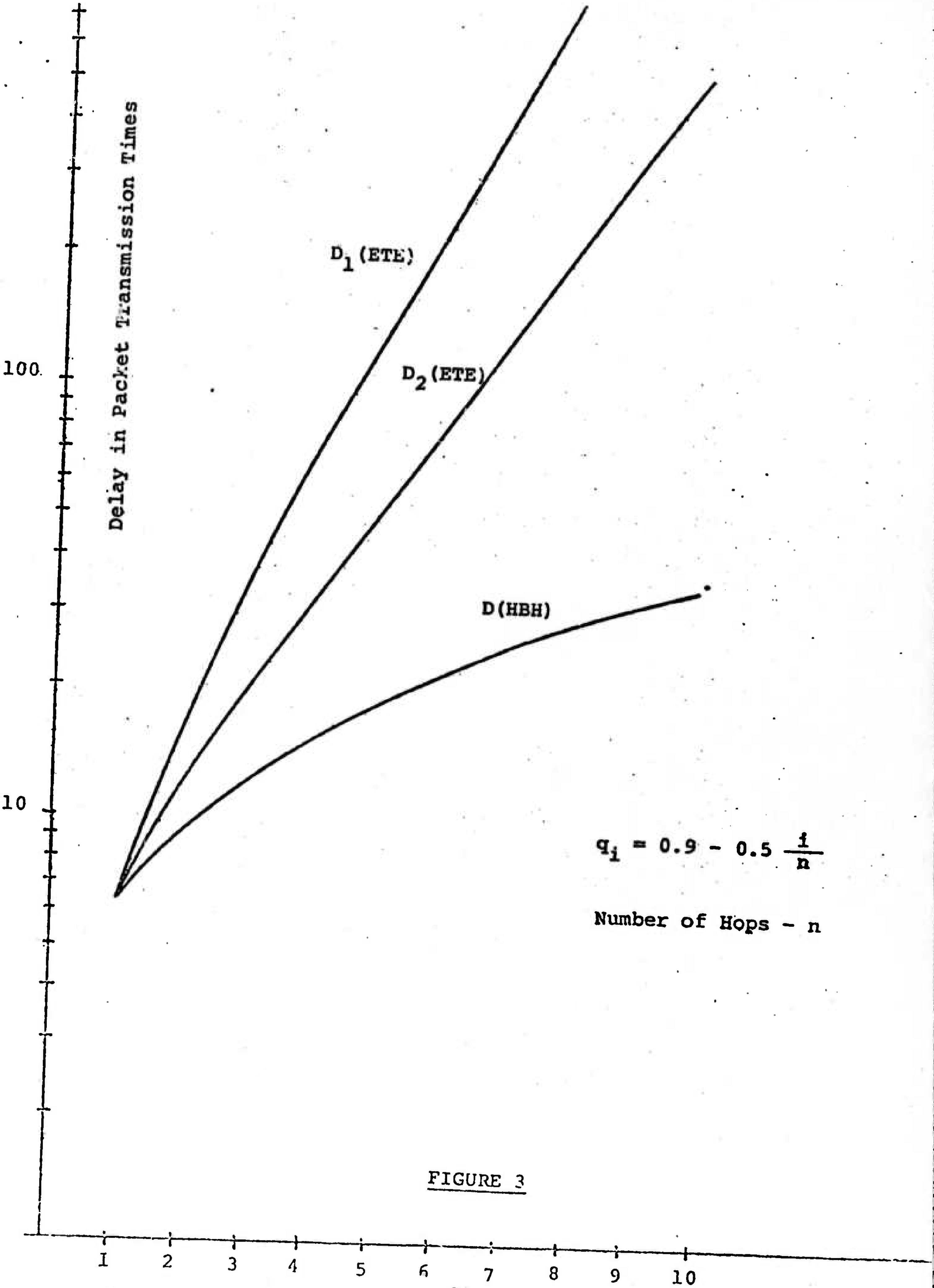
Delay in Packet Transmission Times
100
10
D_1(ETE)
D_2(ETE)
D(HBH)
$q_i = 0.9 - 0.5 \frac{i}{n}$
Number of Hops - n
1
2
3
4
5
6
7
8
9
10

FIGURE 3

time is 10 msec, the average delays are 170 msec, 470 msec, and 1180 msec, for D(HBH), D_2(ETE), and D_1(ETE), respectively.

In practice, q will differ along the path. It is reasonable to assume that the probability of success, q, will decrease when the packet approaches the station. (Simulation results confirm this assumption.) When random access ALOHA systems are used, the practical range for q is from 1/e for which the effective utilization is maximum to 0.9 for which the utilization is 4.7% and 9.4% for the non-slotted and slotted case, respectively. We take a function of the form:

$$q_i = 0.9 - 0.5 \frac{i}{n} \; ; \; i = 1, 2, \ldots n \tag{4}$$

The normalized average delay as a function of n, with q_i, as in Equation (4) is shown in Figure 3.

3.3 EFFECT ON CHANNEL UTILIZATION

We now consider the effect of the acknowledgement scheme on the maximum utilization achievable when using slotted and non-slotted ALOHA random access schemes. To simplify the comparison, we take $\delta = 0$ (this affects the comparison with ETE scheme 1) and assume that q is constant along the path. It is further assumed that the arrival process, to each repeater, of new packets and new packets plus retransmissions are both Poisson with mean rates S and G, respectively and that the packet transmission time is one unit.

Given an n-hop system, suppose that one wants to use an ETE acknowledgement scheme such that the average delay equals that when using a HBH scheme. Equating (1) and (2), and (1) and (3), respectively, one obtains *:

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* We use subscripts 1 and 2 to denote variables for ETE schemes 1 and 2, respectively; variables without a subscript will denote quantities related to the HBH scheme.

$$q_1 = q^{1/n} \; ; \; q_2 = \left(\frac{q}{n-(n-1)q}\right)^{1/n} \tag{5}$$

The relation between the channel traffics G for the acknowledgement schemes, when using slotted ALOHA are:

$$G_1 = \frac{G}{n} \; ; \; G_2 = \frac{Y^{-1}}{n} \tag{6}$$

where

$$Y = \frac{e^{-G}}{n-(n-1)e^{-G}} \tag{7}$$

Consequently, the channel utilizations (or throughputs) S are related as follows:

$$\frac{S_1}{S} = \frac{1}{n} e^{-G(1-\frac{1}{n})} \tag{8}$$

$$\frac{S_2}{S} = \frac{e^{-G}}{nG} Y^{1/n} \ell n Y^{-1} \tag{9}$$

The ratios of utilization (Equations (8) and (9)) as a function of n are shown in Figure 4, for the case G = 0.5 which is equivalent to 30% utilization in the slotted ALOHA random access system.

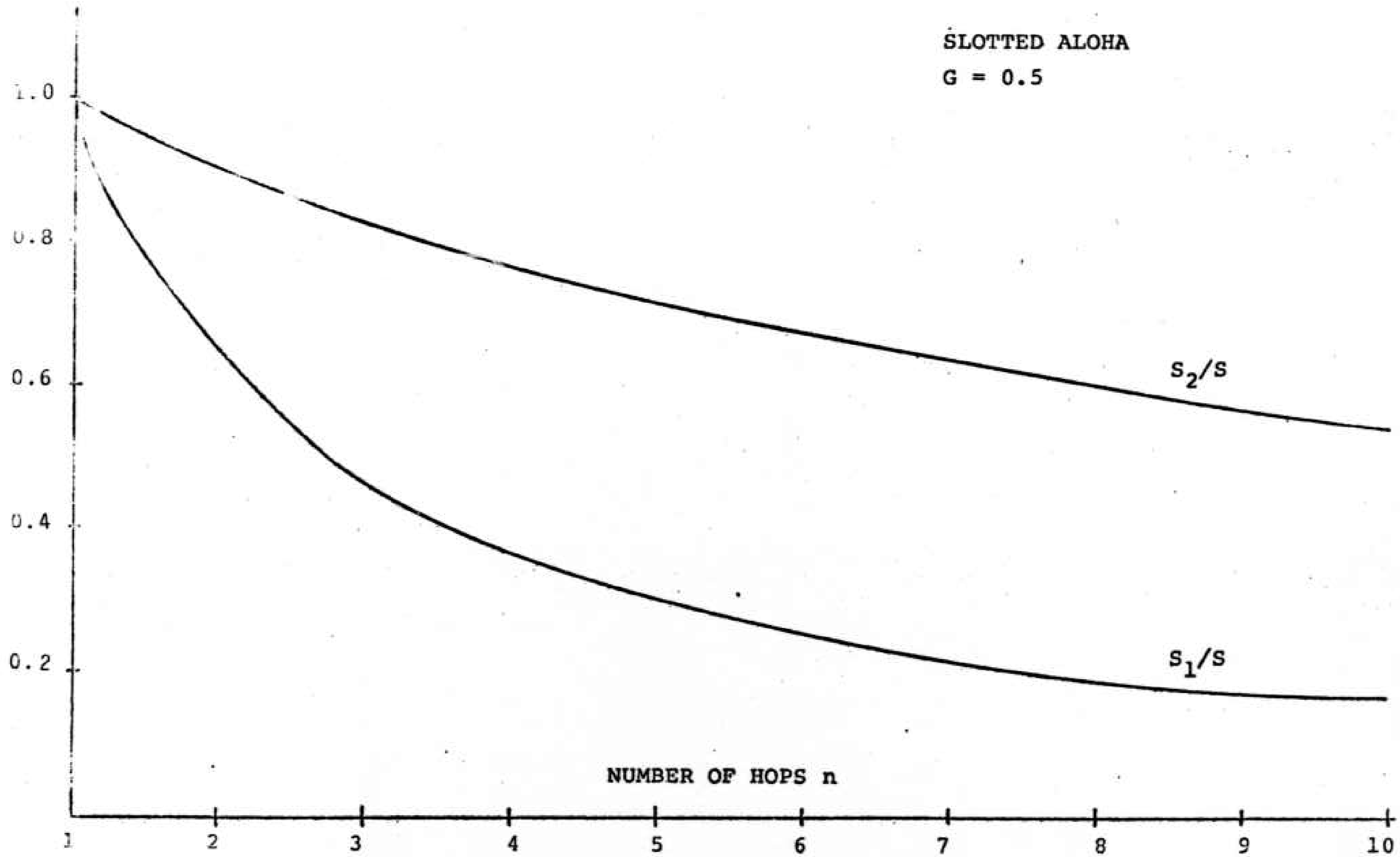
SLOTTED ALOHA
G = 0.5
1.0
0.8
0.6
0.4
0.2
S2/S
S1/S
NUMBER OF HOPS n
1
2
3
4
5
6
7
8
9
10

FIGURE 4

4. A DIRECTED ROUTING PROCEDURE

In this section, a routing scheme is proposed aimed at achieving maximum throughput and minimum delay. This is obtained by using shortest path (minimum hop) routing from terminal to station and from station to terminal, and by preventing, where-ever possible duplicate copies of a packet from being circulated in the network. However, the routing procedure includes sufficient flexibility so that when the first choice shortest path cannot be used, the packet departs from this path and uses a shortest path from its new location. One pays overhead for this efficiency by "carrying" two labels in the packet header.

4.1 LABELING

The shortest path routing is obtained by labeling the repeaters to form, functionally, a hierarchical structure as shown in Figure 5. Each label includes the following information:
(i) a specific address of the repeater for routing purposes,
(ii) the minimum number of hops to the nearest station, and
(iii) the specific address of all repeaters on a shortest path to the station and the address of the repeater to which a packet has to be transmitted when destined to the station.

For simplicity, we describe routing for the case of a one station network. A label of repeater R_i of hierarchy level j will be denoted by L_{ij}; $i,j>1$. The station will have the label L_{11}. L^*_{ij} will denote the label of the repeater which is the "nearest available" to the communicating terminal.

A label is composed of H subfields, where H is the maximum number of hierarchy levels (H-1 is the maximum number of hops on the shortest path between any repeater and the station). Every subfield has three possible entries, blank (BLK), a serial number (SER), or ALL. L_{ij} has j entries SER's and (H-j) BLK's as shown below:

1	2		j-1	j	j+1		H
SER	SER	. . . .	SER	SER	BLK	. . . .	BLK

$$\underbrace{\qquad\qquad\qquad\qquad}_{j\ \text{serial numbers}}\qquad\underbrace{\qquad\qquad\qquad}_{(H-j)\ \text{blanks}}$$

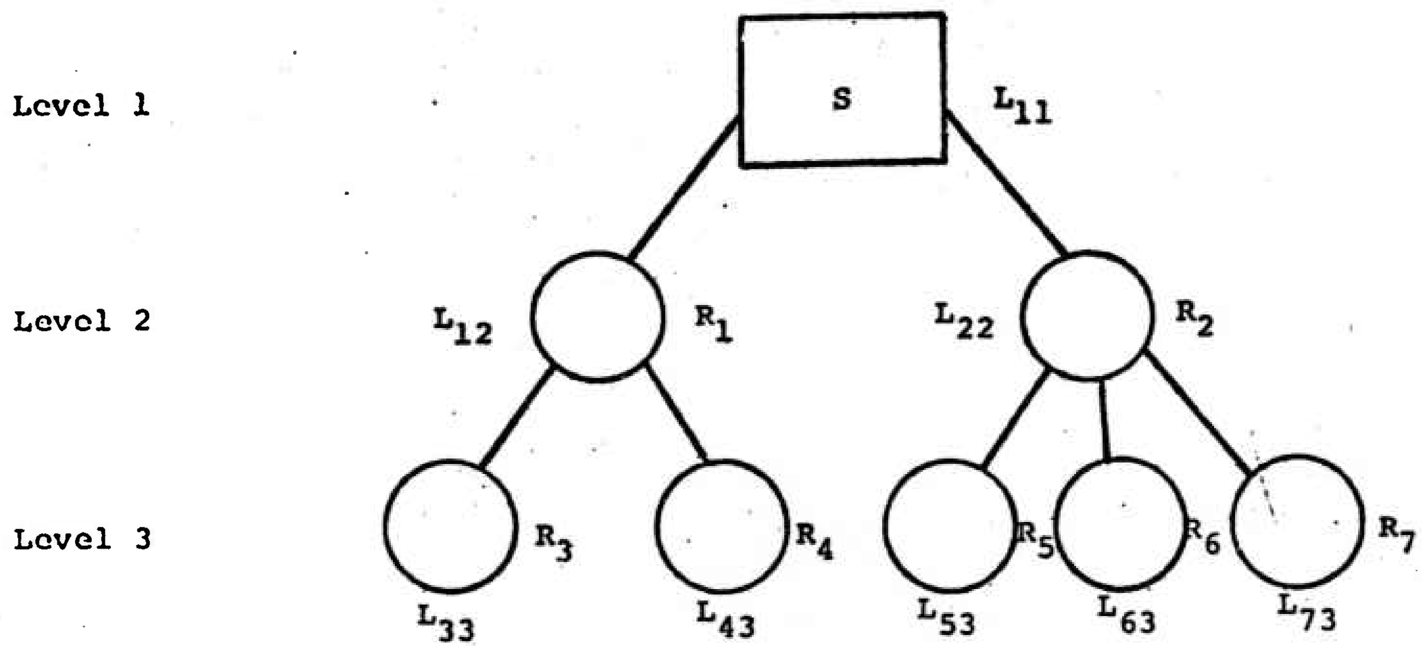
Level 1
Level 2
Level 3
S
L_{11}
L_{12}
R_1
L_{22}
R_2
R_3
R_4
R_5
R_6
R_7
L_{33}
L_{43}
L_{53}
L_{63}
L_{73}

FIGURE 5

We say that L_{ij} "homes" on L_{kp}, $h(L_{ij}) = L_{kp}$, if $p = j-1$ and the first j-1 subfields of both are identical. If two repeaters at level j home on the same repeater, their labels will differ only in the entry to subfield j.

As an example, if we use 3 bits per subfield, the labels of the station and the repeaters of the network shown in Figure 5 are as follows:

	Subfield 1	Subfield 2	Subfield 3
L_{11}	0 0 1	0 0 0	0 0 0
L_{12}	0 0 1	0 0 1	0 0 0
L_{22}	0 0 1	0 1 0	0 0 0
L_{33}	0 0 1	0 0 1	0 0 1
L_{43}	0 0 1	0 0 1	0 1 0
L_{53}	0 0 1	0 1 0	0 0 1
L_{63}	0 0 1	0 1 0	0 1 0
L_{73}	0 0 1	0 1 0	0 1 1

In this example, a subfield in which all bits are "0" is considered "blank." Note that all entries in Subfield 1 are the same since all repeaters home (eventually) on the same station.

4.2 ROUTING

The packet header, in both directions, will include the following routing information.

L_{kn}	L°_{ij}	OTHER HEADERS AND PACKET INFORMATION
TO	LABEL OF NEAREST REPEATER TO THE TERMINAL	

L_{kn} is the label of the repeater to which the packet is currently addressed. The complete packet will always be transmitted to a specific device; other devices which may receive the packet will drop it. The shortest path from a terminal to the station consists of L°_{ij}, $h(L^\circ_{ij})$, $h(h(L^\circ_{ij}))$, up to L_{11}, in the given order, and in the reverse order when routing from station to terminal. When a specific repeater along the shortest path is not known (by the terminal) or not available, then the terminal or repeater (which has the packet) will transmit only the header part of the packet, trying to identify a specific repeater. In that case, the label L_{kn} will include some entries ALL.

A. Routing from Terminal to Station

When a previously silent terminal begins to communicate, it first identifies a repeater or a station in its area. It transmits only the header part of the packet with all entries in L_{kn} set to ALL. The header is addressed to all repeaters and stations that can hear the terminal. A device which correctly receives this header substitutes its label in the space L_{kn} and repeats the header. This particular L_{kn} is also L°_{kn} and will be used by the terminal to transmit all packets during this period of communication. If a terminal is stationary, it can store this label for future transmissions. L°_{kn} begins to transmit the complete packet along the shortest path to the station.

Suppose that L_{ij} along the shortest path is not successful in transmitting the packet to $h(L_{ij})$. Then L_{ij} begins the search stage of trying to identify another repeater. In the first step, it tries to identify a repeater which is in level $p \le j-1$. This is done by using the following label:

1	2	3		j-1	j	j+1		
SER	ALL	ALL	. . .	ALL	BLK	BLK	. . .	BLK

The header is addressed to all repeaters in levels 2 to j-1, which eventually home on L_{11}. If this step is not successful, in the second (last) step, L_{ij} tries to identify any available repeater by using a label in which the first entry is SER and all other entries are ALL. When a specific repeater is identified and receives the packet, it transmits the packet on the shortest path from its location.

Note that if repeaters have sufficient storage, they can save alternative labels and thus reduce the necessity of searching for a specific repeater. Alternative solutions in which repeaters have multiple labels are also possible.

B. Routing from Station to Terminal

L°_{ij} contains sufficient information for shortest path routing to the terminal. Denote by h^{-1} the inverse of h and by $h^{-2} = h^{-1}(h^{-1})$, etc. The shortest path from station to terminal includes $h^{-(j-1)}(L^\circ_{ij})$, $h^{-(j-2)}(L^\circ_{ij})$, ... $h^{-1}(L^\circ_{ij})$, and L°_{ij}. If some L_{kp} is not successful in transmitting the packet along the shortest path, it begins the process of identifying another specific repeater. Note that when routing to the station, the next label is always a function of the label of the repeater that currently stores the packet. When routing to the terminal, the next label is a function of L°_{ij} and the hierarchy level of the repeater that currently stores the packet. Thus, when routing to the terminals, it will be useless to transmit the packet backwards, since it will usually arrive back at the current location; therefore it is more efficient to delay the packet. If, when routing to the terminal, a repeater on the shortest path is temporarily unreachable, the procedure attempts to by-pass this particular repeater and regain the original shortest path route. The labels that will be used by $h^{-3}(L^{\circ}_{ij})$ are shown below:

	1	2		j-3	j-2	j-1	j	j+1		
$h^{-3}(L^{\circ}_{ij})$	SER	SER		SER	BLK	BLK	BLK	BLK		BLK
$h^{-2}(L^{\circ}_{ij})$	SER	SER		SER	SER	BLK	BLK	BLK		BLK
SEARCH	SER	SER		SER	SER	ALL	BLK	BLK		BLK

All the entries SER are taken from L°_{ij}.

4.3 ACKNOWLEDGEMENTS

In Section 3, it has been shown that the use of HBH acknowledgements in the routing scheme in addition to the ETE acknowledgement is desirable. However, one can prevent specific acknowledgement packets from being transmitted by using the passive "echo" acknowledgement. This approach has other advantages as well, which will be described in the next section. Echo acknowledgement will be employed along the path. That is, the device transmitting the packet waits in a receive mode to receive this same packet when it is repeated by the next stage. The reception of the packet when transmitted by the next stage constitutes the acknowledgement, since it indicates that the next repeater stage has correctly received the packet and will store and retransmit it as necessary.

In fact, one has the option of adding parity bits after the header. In this event, it would be sufficient to "hear" the header of the packet, and thus, the header plus parity bits will constitute the acknowledgement. At the end of a path, the terminal or station will repeat the header. Note that the probability of correctly receiving an acknowledgement would be higher than the probability of correctly receiving a packet, due to the difference in transmission time. Furthermore, in many cases the station may correctly receive the header, whereas the entire packet is received in error. The information contained in the header can be used by the station for control purposes.

4.4 TRAFFIC CONTROL

The control procedures to be implemented would use control packets from stations to repeaters, from stations to terminals, and possibly from repeaters to terminals. Some of these may be implemented in the station - repeater protocol and relate to the initialization of repeaters, relabeling of repeaters under various overload conditions, activation and deactivation of repeaters, and so on. In this section, we discuss controls necessary for the routing scheme. These are:

(i) Initial search by the terminal.
(ii) Maximum handover number (MHN).
(iii) Maximum number of transmissions (MNT).

It was demonstrated in [Kleinrock & Lam; 1974] that after channel traffic exceeds a certain value, throughput reduces. If the number of retransmissions is not limited, the offered channel traffic will increase indefinitely and the throughput will reduce to zero. Thus, one problem is to prevent new traffic from entering the system when the system is congested. This control can be obtained by the search procedure which is used by terminals when entering the system. This control is "local" in the sense that it depends on the traffic level in the geographical neighborhood of the terminal. Terminals will nevertheless be able to enter the system when being "far" from stations, and the traffic introduced will propagate towards the stations.

The MHN is a control aimed at suppressing packets from being propagated in endless cycles of repeaters or being propagated over paths containing many hops. This may occur when packets depart from the shortest path. Furthermore, if the routing scheme used is relatively unsophisticated, the MHN will prevent the packet from arriving at remotely located stations. The MHN for a given packet depends on the number of hops between the originating terminal and the station with which it communicates (or vice versa). It will therefore be a function of the hierarchy level of the repeater with the label L°_{ij}.

The MNT is also a local control which reduces the traffic level when the system is congested by discarding a packet after a specified number of retransmissions have been attempted. It also prevents repeaters from indefinite transmission of a packet when surrounding repeaters are temporarily blocked and are unable to accept packets.

4.5 PACKET FORMAT

A possible packet format for performing the routing described is shown below:

HEADER	PACKET INFORMATION	PARITY

The header includes the following items:

T/F	C/I	DID	OID	L_{kn}	L°_{ij}	MHN	E/C

T/F - a bit, indicating whether the packet is addressed To station or From station.

C/I - a bit, indicating whether the packet is a Control packet or an Information packet.

DID - Destination address.

OID - Origination address.

L_{kn} - The label of the repeater to which the packet is currently addressed.

L°_{ij} - The label of the repeater "nearest" to the terminal which originated the packet or to which the packet is transmitted.

MHN - Maximum handover number.

E/C - An error or control message. If it is an information packet, the space may include a sequential number, specification of the packet number in the message, etc. If it is a "header packet," it may include an error message asking for retransmission of a certain number of packets. If it is a control packet from the station, this space may be used for the control message.

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5. A PROCEDURE FOR REPEATER LABELING

In this section, we identify some of the problems of repeater labeling and propose one approach for the initial labeling of a repeater network. Assume that initially every station and repeater has a fixed ID, R_i. This ID will be used for labeling purposes, to identify whether the device is operative or dead, to activate and deactivate a repeater, and for other control purposes.

The station will determine and assign labels to all repeaters in its area in the initial labeling procedure. When more than one station operates in an area, the initial labeling will be done by the stations sequentially, and repeaters may be allowed to choose the home station according to the minimum number of hops.

First, it is necessary to specify two parameters: (i) the maximum number of subfields or hierarchy levels, say H, and (ii) the number of bits per subfield, say B. These parameters are to be the same for the entire broadcast network in order to have the same packet format when transmitting information. If some sections of the broadcast network are disjoint, it is sufficient that B x H be the same for the entire network. As indicated before, H-1 is the maximum number of hops that a packet will travel when using the shortest path route, and 2^B-2 is the maximum number of repeaters that can home on a single repeater or station (one label is needed for ALL and another for BLK). 2 x H x B is the number of bits in the header which will contain the routing information.

The initial labeling procedure is:

STEP I:

The station transmits a control packet to every repeater sequentially. This packet includes an MHN as well as another MHN to be used by the addressed repeater for its response packet. There is no directed routing at this stage; every repeater which correctly receives the control packet decrements its MHN, and stores and retransmits it until echo acknowledged by the next stage. The control packet is dropped when its MHN reduces to zero.

The repeater to which this packet is addressed transmits a response packet to the station using the assigned MHN. Every repeater which receives this packet will decrement the MHN and add its R_i in order.

The station may receive one response packet, several, or none. If no response packet is received, the station can try several more transmissions, each time increasing the MHN's, or conclude that the repeater is dead (this repeater can possibly be reached from another station).

STEP II:

The information acquired from the response packets is sufficient to determine a hierarchical labeling structure. In this step, the station processes the information and determines an "optimized" structure. The processing performed during this step is described in the next section.

STEP III:

In this step, the station tests the shortest path, particularly in the direction from station to repeaters, which was not tested before. The station transmits a control packet to every repeater, using its label. The station uses an MHN equal to the number of hops on the shortest path so that if this path is not possible the repeater will not be able to receive the packet. A repeater which receives this packet transmits a response to the station, which constitutes an ETE positive acknowledgement. If all repeaters have been successfully tested, the procedure ends; otherwise, the program returns to Step II for further processing.

5.1 AN ALGORITHM FOR DETERMINING THE LABELS

We describe a technique for processing the response packets and for determining the hierarchical labels in Step II of the labeling procedure. In general, the repeaters may be distributed at random locations, and the station may not know the geographical locations of repeaters. Furthermore, there may be more repeaters than the number needed for efficient routing. Ideally, one would want to obtain a network of repeaters which has the following properties:

(1) There should be a minimum number of hierarchy levels.

(2) There should be a shortest path from every repeater to the station.

(3) The entire area should be covered with a minimum number of repeaters.

(4) Every repeater should be able to transmit directly to at least j (say 2) other repeaters.

(5) The number of repeaters which home on one single repeater or station should be $\leq 2^B-2$.

A solution which satisfies all the requirements may not exist. For example, if more than 2^B-2 repeaters can directly reach the station and none can be deactivated, requirement (2) will not be satisfied.

Suppose that there are N-1 repeaters and one station denoted by R_1. The station first constructs a connectivity matrix $\underline{C} = (c_{ij})$, where,

$$c_{ij} = \begin{cases} 1 & \text{if device j can hear i directly} \\ 0 & \text{otherwise.} \end{cases} \qquad i,j = 1, 2, \ldots, N$$

$\underline{C}$ is constructed from the response packets in Step I. For example, if a response from R_i contains (in order) R_e, R_m, R_k, ..., then $c_{ie} = 1$, $c_{em} = 1$, $c_{mk} = 1$, ... One can see that the station does not have to transmit the first labeling packet to all repeaters, since it can learn about the functional location of some of the repeaters which were on the return path of previous control packets. Furthermore, the number of response packets to a control packet can be increased when the MHN assigned by the station is increased. Finally, we note that C is not necessarily symmetric.

The entries 1 in row i indicate the repeaters to which R_i can directly transmit, and the entries 1 in column i indicate the repeaters from which R_i can directly receive. The structure of the repeater network will be recorded by the vector $\underline{h}$, where h_j, $j = 1, \ldots, N$, indicates the repeater on which R_j homes.

Let $S(m)$ denote the set of repeaters whose shortest path to the station includes exactly m hops. Assume that all repeaters in $S(1)$, $S(2)$, ..., and $S(m)$, have been labeled (assigned home repeaters). We describe the labeling of (repeaters in) $S(m+1)$ by (repeaters of) $S(m)$. At every state k of labeling $S(m+1)$ by $S(m)$, we characterize repeaters of $S(m)$ by:

$d_i(k)$ = The number in $S(m+1)$ which have been labeled R_i (say, the degree of R_i at state k)

$v_i(k)$ = The number in $S(m+1)$ which still can be labeled R_i

$f_i(k)$ = The potential degree of R_i at state k, i.e., $f_i(k) = d_i(k) + v_i(k)$.

Repeaters of $S(m+1)$ will be characterized by:

u_j = The number of repeaters in $S(m)$ which can label it.

At every state k, we distinguish among three disjoint subsets of $S(m)$ and $S(m+1)$:

$S_m^F(k) = \left\{R_i : v_i(k) = 0\right\}$ = cannot label more*

$S_m^L(k) = \left\{R_i : f_i(k) \leq 2^B-2;\right.$ ordered according to increasing values of $d_i(k)$ and when the same then according to increasing values of $\left.v_i(k)\right\}$

$S_m^R(k) = \left\{R_i : f_i(k) > 2^B-2;\right.$ ordered according to increasing values of $\left.v_i(k)\right\}$

$S_{m+1}^F(k) = \left\{R_i : h_i(k) > 0\right\}$ = already labeled (assume that $h_i(0) = 0$)

$S_{m+1}^L(k) = \left\{R_i : c_{ip} = 1\right.$ for some $R_p \varepsilon S_m^R(k)$; ordered according to decreasing values of the number of repeaters in $S_m^R(k)$ that can label them and when this is the same, then according to decreasing values of $\left.u_i\right\}$

$S_{m+1}^R(k)$ = The remaining repeaters of $S(m+1)$ ordered according to decreasing values of u_i

Note that $S_m^R(k)$ is the set which can potentially violate requirement (5), and $S_{m+1}^L(k)$ is the set to be labeled which may result in this violation. Therefore, one should try to label $S_{m+1}^L(k)$ by $S_m^L(k)$. When such a label is assigned, it decreases the values of $f_i(k)$ in $S_m^R(k)$. Furthermore, the orders of the subsets of $S(m)$ according to $d_i(k)$ are aimed at obtaining a network in which the repeaters of $S(m+1)$ are divided equally among repeaters of $S(m)$ (this was not specified in the requirements). The order of $S_{m+1}^L(k)$ is done so that (if possible) the repeater which can be labeled by the largest number of repeaters of $S_m^R(k)$ is labeled first.

The algorithm proceeds as follows:

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* There may be repeaters in $S(m)$ for which $v_i(0) = 0$. It is convenient to refer to these as "end repeaters of level m" since no repeater will home on these.

STEP A:

Take the first of $S^L_{m+1}(k)$ and label it by one of $S^L_m(k)$ (beginning with the first). If it is labeled, evaluate the subsets of state k+1 and do the same; if not, take the next repeater of $S^L_{m+1}(k)$ and do the same.

STEP B:

(i) If $S^R_m(k)$ is empty, then $S^L_{m+1}(k)$ is also empty; complete the labeling of $S^R_{m+1}(k)$ by $S^L_m(k)$ using the procedure of Step A, then return to Step A.

(ii) If $S^R_m(k)$ is not empty, label one of $S^L_{m+1}(k)$ by $S^R_m(k)$ using the procedure of Step A, then return to Step A.

Note that if all of $S^R_m(k)$ becomes part of $S^L_m(k)$ at some state, the network produced satisfies requirements (1), (2), and (5); since (5) is satisfied by the last statement, (2) implies (1), and (2) is satisfied by the definition of S(m). If one of the above requirements must be violated, modification of (ii) in Step B of the algorithm is required.

The sets S(m), defined at the beginning of this section are constructed recursively as follows:

$$S(1) = \left\{ R_k : c_{k1} = 1 \right\}$$

$$\vdots$$

$$S(m+1) = \left\{ R_k : R_k \not\in \bigcup_{p=1}^{m} S(p),\ c_{kj} = 1 \text{ for some } R_j \varepsilon S(m) \right\}$$

That is, to construct S(m+1) it is necessary to examine in the matrix C only the entries 1 in the columns which corresponds to repeaters of S(m) and choose the ones that have not been identified yet.

5.2 REMARKS

1. The shortest path label assignment does not, in [illegible] physical [illegible]. That is, the label assignment depends on the terrain as well

as possible variations in transmission power and reception sensitivity of devices. Thus, it is a functional rather than a geophysical assignment.

2. In the practical case, it may be necessary to label redundant repeaters and then deactivate them for use as stand-by repeaters. Furthermore, the procedure for this process should satisfy requirements (3) and (4) of the previous section.

6. SOME SIMULATION RESULTS

A computer program which simulates in detail the operation of the packet radio network and the proposed routing and control techniques is currently available and will be described in [NAC; 1974 (b)]. Some qualitative observations of the model's performance as related to routing are described in this section.

Figure 6 shows one network that has been extensively studied. The labels of repeaters in this network were assigned *a priori*, and the lines connecting the devices in Figure 7 signify the hierarchical structure created by the implementation of the directed routing technique. Terminal traffic is introduced into the system at random times, and originates at random locations on the plane. Once a terminal is introduced, it begins the search procedure, and the communication between a terminal and a station proceeds using the routing and control schemes described including the ETE and Echo HBH acknowledgements. Figure 6 shows the connectivity of the network simulated. That is, when a particular repeater transmits, all devices connected to it by line can receive the packet.

Simulation results demonstrate that the critical hop in the packet radio network is between the first level repeaters and the station. Thus, special attention should be given to the flow control design on this hop. In particular, repeater placement in the neighborhood of the station and the control of these repeaters by the station are significant. These repeaters also have higher power duty cycles, since they repeat all packets of repeaters which home on them.

It is also demonstrated that there is a higher probability of end-to-end successful transmission from station to terminal than from terminal to station. This is observed from the higher frequency of repeater searches and dropped packets when routing towards the station. One cause is that the station is the largest user and thus has higher probability of successful transmission over

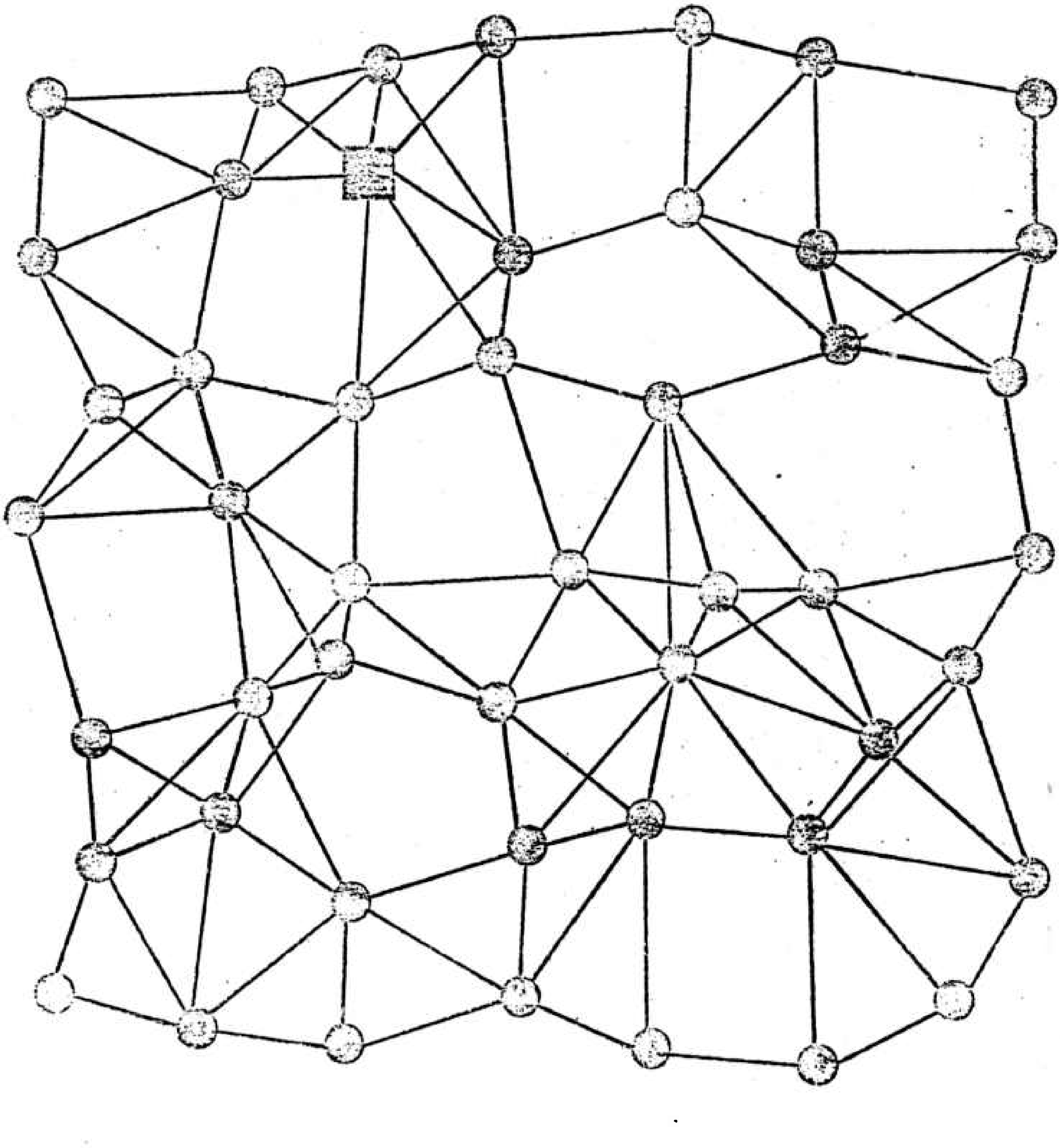


FIGURE 6

the critical hop because it manages its own traffic rather than competing with itself in a broadcast node.

The terminal simulated recognizes a packet addressed to it by checking a portion of the packet header not related to repeater labels. Consequently, the terminal can receive from a different repeater than the one to which it transmits. Since the response to a search packet by repeaters is randomized in time, the terminal frequently identifies a repeater which is not necessarily the nearest to the terminal or the nearest to the station. As a result, the path from the terminal to the station is not necessarily the same as from the station to the terminal. The latter is usually shorter. Such a case is illustrated in Figure 8. Here, the terminal will usually receive its packets from R1 at the time R1 transmits to R2. The echo by the terminal will usually acknowledge R1 and R2 simultaneously. Furthermore, R3 need not handle traffic to the terminal.

Figures 6 and 7 show that while the station has connectivity 7, only 4 of these repeaters are labeled as first level repeaters that home on the station. In particular, note that the station can hear all packets transmitted towards it by R26; however, these packets are addressed to R27. The station finally receives the packets from R27. Consequently, the station is busy a fraction of the time with non-useful traffic. This can be improved by changing the reception and transmission operation of stations. That is, the station can be made to receive from any repeater along the shortest path and to transmit to the repeater nearest to the terminal that it can reach. Another advantage of this type of station operation is that more repeaters can be placed in the neighborhood of the station and labeled arbitrarily. These repeaters may be required for area coverage or reliability considerations.

Other observations of the simulation show that some terminals are blocked when the system in their neighborhood is congested, and that a higher frequency of alternate routing occurs when the traffic offered to the system is increased.

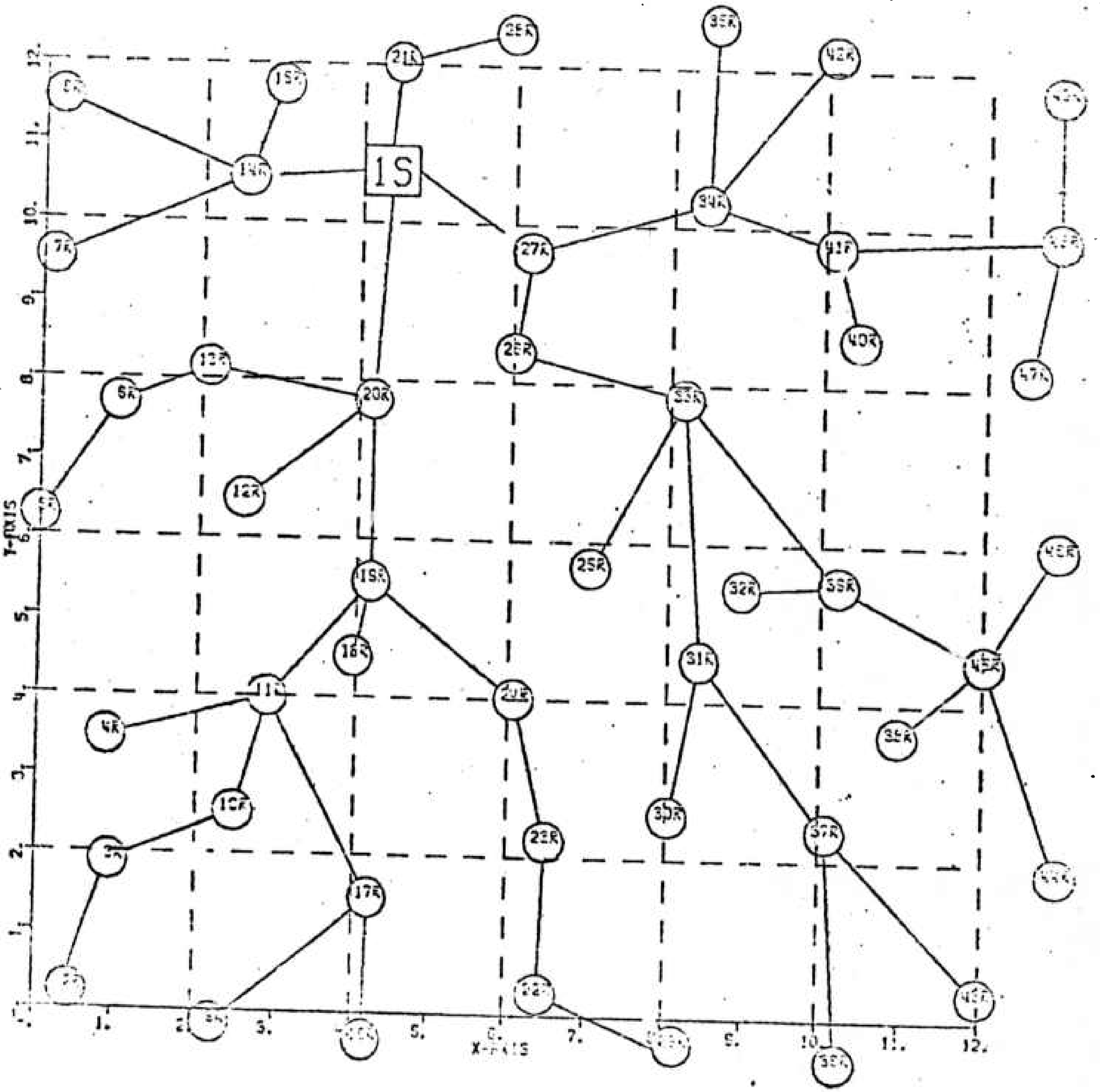


FIGURE 7

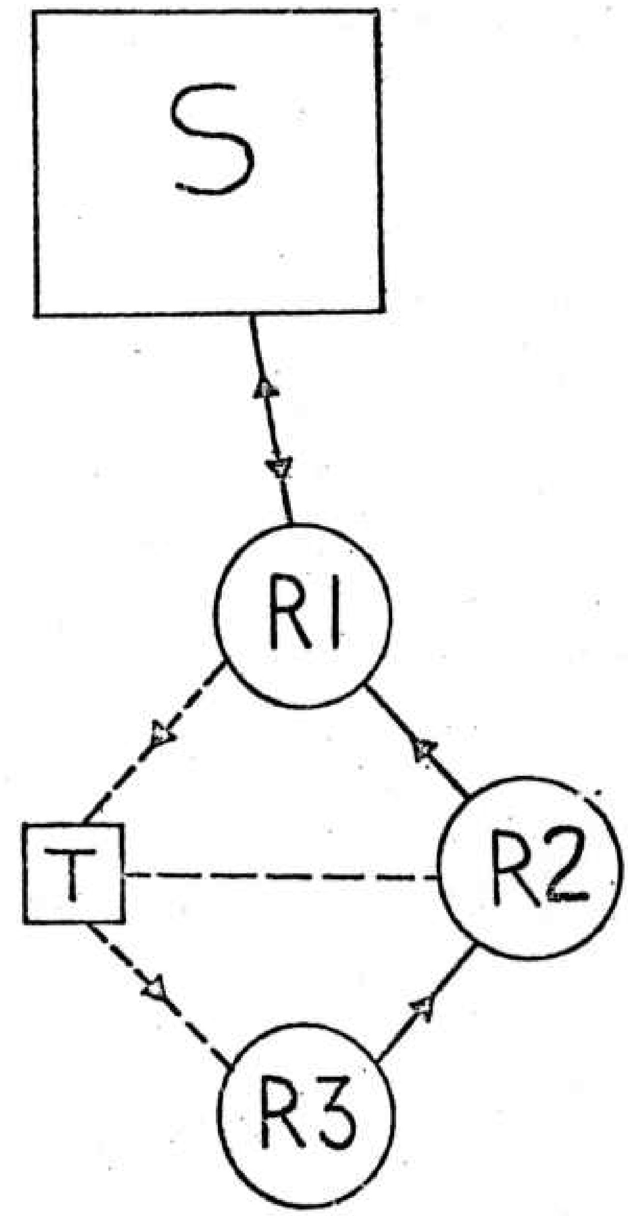


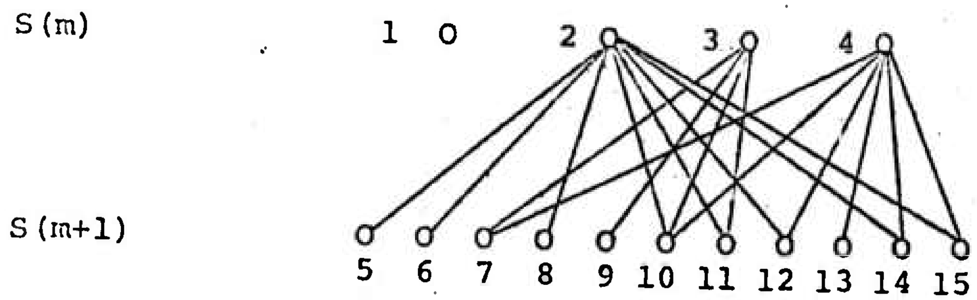
FIGURE 8

An extensive description of the simulation program and experiments is forthcoming [NAC; 1974(b)]. It is expected that the routing procedures described above will be modified as more experience is gathered. Further topics to be investigated include buffer management, flow control, and system initialization. These topics will be the subject of forthcoming reports.

7. APPENDIX: AN EXAMPLE OF REPEATER LABELING

The figure below shows the set S(m+1) to be labeled by S(m). The connection line between $R_i \varepsilon S(m+1)$ and $R_j \varepsilon S(m)$ was drawn to demonstrate that $c_{ij} = 1$, also B = 3 bits.

$f(0) = v(0) =$ 0 8 4 6



u_j = 1 1 2 1 1 3 2 2 1 2 2 .

$S^F_m(0) = \{R_1\}$

$S^L_m(0) = \{R_3, R_4\}$

$S^R_m(0) = \{R_2\}$

$S^F_{m+1}(0) = \{\phi\}$

$S^L_{m+1}(0) = \{R_{10}, R_{11}, R_{12}, R_{14}, R_{15}, R_5, R_6, R_8\}$

$S^R_{m+1}(0) = \{R_7, R_9, R_{13}\}$

First label assigned is $h_{10} = R_3$, then,

$S^F_m(1) = \{R_1\}$

$S^L_m(1) = \{R_4, R_3\}$

$S^R_m(1) = \{R_2\}$

$S^F_{m+1}(1) = \{R_{10}\}$

$S^L_{m+1}(1) = \{R_{11}, R_{12}, R_{14}, R_{15}, R_5, R_6, R_8\}$

$S^R_{m+1}(1) = \{R_7, R_9, R_{13}\}$

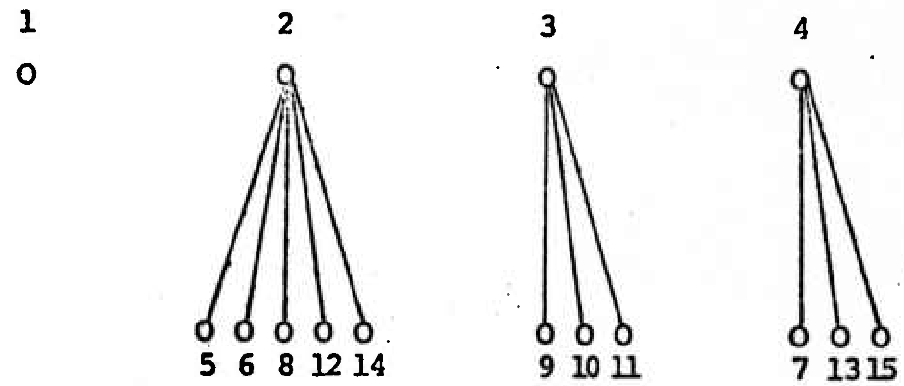
The second label is $h_{11} = R_3$, then,

$$S_m^F(2) = \{R_1\} \qquad S_{m+1}^F(2) = \{R_{10}, R_{11}\}$$

$$S_m^L(2) = \{R_2, R_4, R_3\} \qquad S_{m+1}^L(2) = \{R_{11}, R_{12}, R_{14}, R_{15}, R_7, R_5, R_6, R_8, R_9, R_{13}\}$$

$$S_m^R(2) = \{\phi\} \qquad S_{m+1}^R(0) = \{\phi\}$$

From now on, any labeling will satisfy requirement (5) since $S_m^R(2)$ is empty. The labeling by the algorithm proceeds as follows: $h_{12} = R_2$, $h_{14} = R_2$, $h_{15} = R_4$, $h_7 = R_4$, $h_5 = R_2$, $h_6 = R_2$, $h_8 = R_2$, $h_9 = R_3$, $h_{13} = R_4$. The final network is:



Appendix A.2

System Capacity and Data Rate Considerations

System Capacity and Data Rate
Considerations

1.0 INTRODUCTION

This paper describes some experimental results for the Packet Radio System obtained by simulation. The simulation was developed as a design tool to assist in solving intractable analytic problems. Consequently, we expect to further experiment with the program to answer practical questions pertinent to system design. Two series of experiments are discussed. In the first series, some of the bottlenecks in system performance are identified. In the second series, specific questions related to system design, such as the trade offs between device range and device interference, and the trade offs between single and dual data rate systems are examined. We also estimate the maximum throughput, delays, and blocking in the systems simulated. The aspects of system operation which differ from Appendix A.1 are described in the Appendix to this section.

2.0 PERFORMANCE MEASURES

The packet radio system is a loss and delay system. Thus, throughput and delay measures in the usual sense are not sufficient to characterize system performance. In particular, one cannot obtain a case in which the delay is very high since above a certain delay, blocking will increase faster than the delay. The blocking phenomena is desirable since it can be used as a control to limit the traffic level. The following measurements taken in the simulation are used to evaluate system performance.

Throughput: Throughput is measured per terminal slots. It is measured at the station and evaluated by counting the number of IP's from terminal received at the station plus the number of IP's transmitted to terminals and which have been ETE acknowledged, divided by the number of terminal simulation slots. The actual throughput of the system may be slightly higher than the one measured since some of the IP's may have arrived at terminals and not yet been acknowledged while some ETE ACKS can be received at the station after the station discarded packets. These are not counted.

Delay: The following delays are measured: terminal search delay, the delay of the arrival of the first IP from the terminal to the station, the delay until the terminal receives the ETE ACK from the station, and total time spent in the system (terminal interaction delay). The search delay is not significant, it will usually be below one terminal slot; otherwise, the terminal will most often be blocked. The delay of the arrival of the first IP to the station and the ETE ACK back to the terminal seems to be significant since they are independent of the delays in the PTP network and of the number of packets that the terminal has to receive in response. Furthermore, the ETE ACK delay is an indication that the terminal established communication with the PTP network.

Prob. Station Busy: The station is sampled during the simulation and is assumed busy if it is actively receiving or transmitting. Otherwise, it is considered idle. This measure is an indication of the channel traffic at the station.

Blocking: If a terminal does not identify a specific receiver after transmitting a SP the number of times specified, it departs from the system. We say that such a terminal is blocked. The packets considered blocked in this case are the packets from the terminal to the station and the response packets that the terminal was supposed to receive from the station. We use this count to evaluate the percentage of IP's blocked.

Total Loss: In addition to the terminals blocked, some terminals may depart from the system, after establishing communication with the station, without receiving any or all of the response packets from the station.

Terminals Remaining: This indicates the number of terminals which are still in the system when the simulation ends.

Rate of Station Response: A parameter in the simulation indicates the average number of response packets from station to terminals. When the value measured is much below the average, it indicates that the system is saturated since the offered traffic rate to the system is much higher than the maximum throughput of the system. This will also be reflected in the difference between the total loss and the blocking, as well as in the number of terminals remaining in the system.

Remarks: (1) Some of the measures give "overlapping" information, and are taken in order to obtain more detailed observations of system performance. For example, the total loss indicates the probability of blocking in the system; however, by taking the blocking measurement, one is able to determine whether the loss is due to the inefficiency in routing or because of difficulty in entering the system.

(2) The differences between offered rate, throughput and loss is in the number of terminals remaining.

3.0 PRELIMINARY OBSERVATIONS

The first system simulated was a common channel single data rate system, in which the station is routing traffic as a repeater (Naive Station). We denote the system as CCSDR (NS). The system defined has a single data signalling rate for communications between terminal and repeater (or station) and in the repeater-station network; the channel is used in a half-duplex mode. When the station is routing traffic as a repeater, it cannot receive packets not addressed to it.

In all experiments reported here, the labels of repeaters and station were preassigned. The hierarchical (directed) labelling scheme of the system in this experiment is shown in Figure 1. Figure 2 shows the connectivity of the repeaters and station. That is, when a device transmits, all the devices connected to it by line are within an effective range and "hear" the transmission.

The objective of the first series of experiments was to observe the detailed operation of devices and the efficiency of the system. The following observations were made:

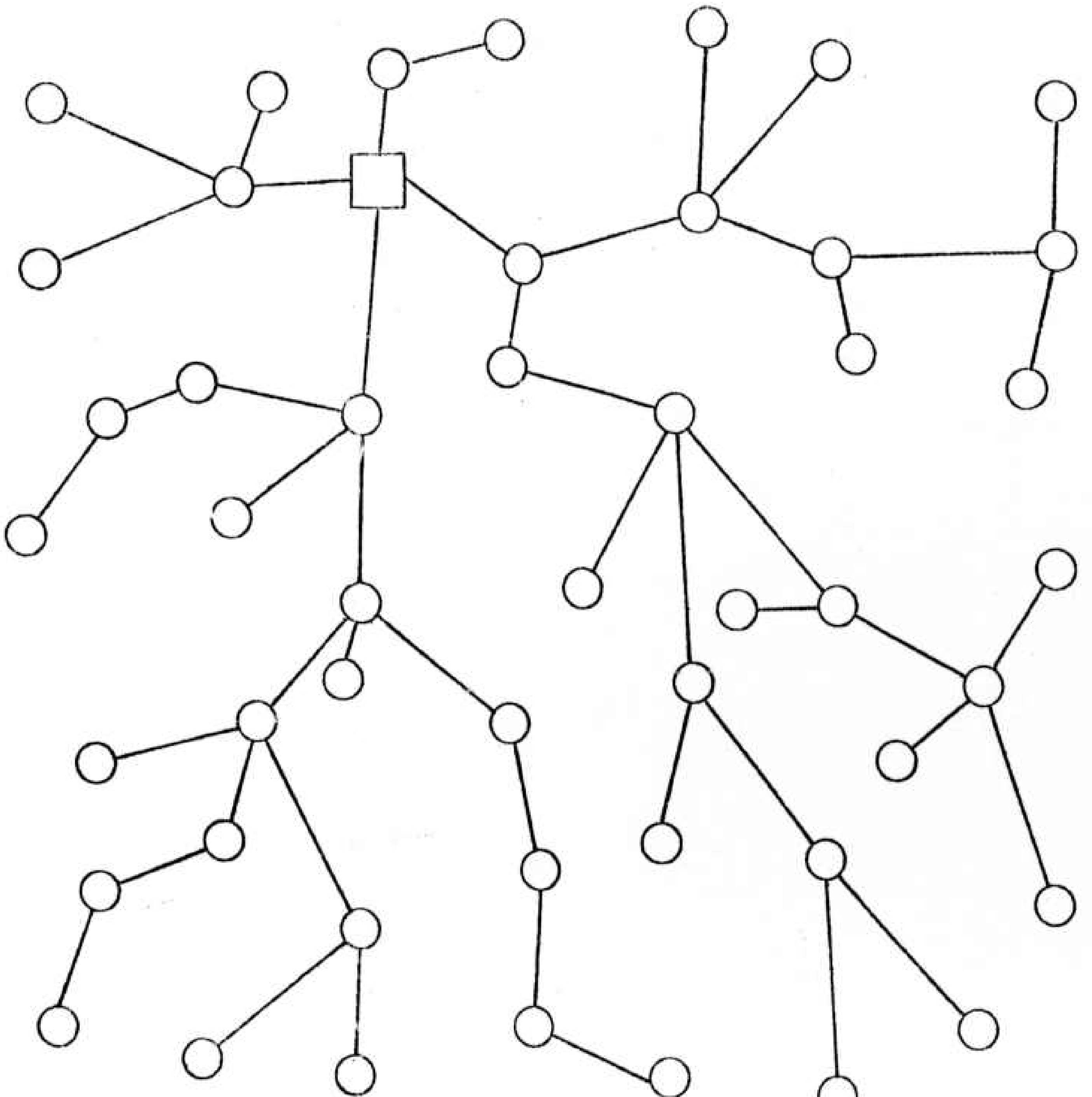


Figure 1.

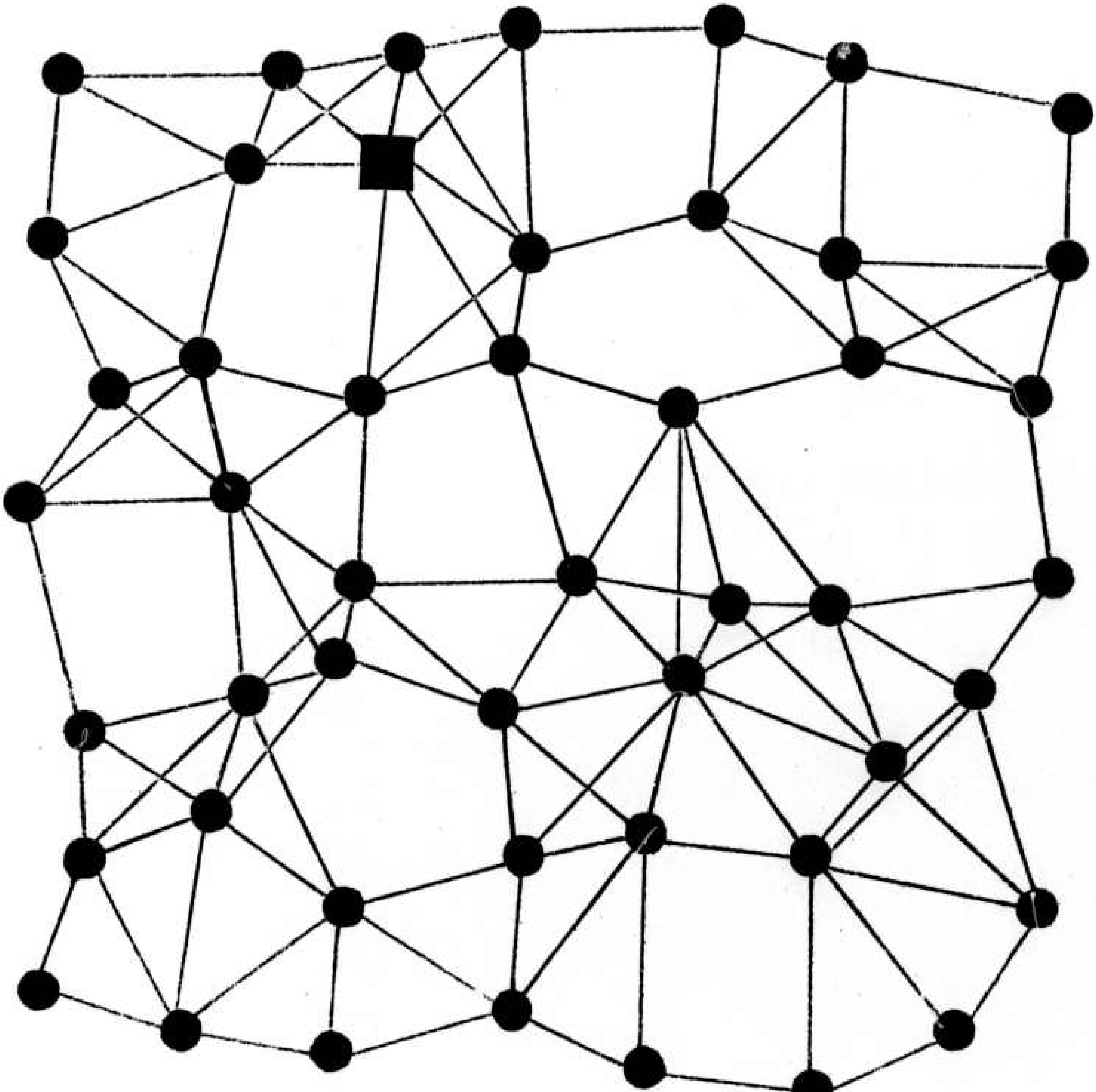


Figure 2.

1. The "critical hop" in the system is that between the first level repeaters and the station. This was concluded by observing the frequency at which repeaters began to search and at which they discarded packets, and from the observation that there is no significant difference in the delay when the number of hops from the station that a packet travels is increased.

2. There is a higher probability of end-to-end successful completion when routing from the station to a terminal than when routing from a terminal to the station. Practically, there is almost "no" difference in time delay between the delay of an information packet from the terminal arriving at the station and the time that the terminal receives an ETE ACK from the station.

3. Many packets associated with terminals that have departed from the station are routed in the network.

The effect of improving the routing capabilities of the station can be readily observed. In particular, one can see in Figures 1 and 2 that while the connectivity of the station is 7, there are only 4 repeaters labelled from the station. Consequently, the station is busy part of the time with non-useful traffic. This situation can be improved by changing the routing of the station so that: (1) it receives any packet that it can hear and which is (eventually) addressed to it; and (2) it transmits response packets to the repeater nearest to the terminal along the routing path that it can reach. This change was implemented for all system studies subsequent to the initial experiments. Apart from the change implemented, the observation suggests that particular attention should be given to the design of the repeater network in the neighborhood of the stations. It is also noted that these repeaters have a higher power duty cycle since they handle all packets collected from other parts of the network. The routing change made at the station enables the allocation of many more repeaters in the neighborhood of the station than are functionally needed, without resulting in an increase in the artificial traffic generated. The exact labelling of these repeaters is also not critical.

One of the reasons leading to observation 2 is that the station has a higher probability, than the first level repeaters, of successful transmission over the critical hop because it is the largest user and does not interfere with its own transmissions. Theoretically, one may expect a similar conclusion when considering transmission in a section of the network in which two repeaters, one of which "homes" on the other, compete. This, however, may not be realized in the system simulated because of the limited storage available in repeaters.

Observations 2 and 3 suggest a change in the terminal-station protocol. The basic question is whether a terminal should release itself from the system or whether it should be released by the station. The former was initially simulated. It was observed that in many cases a terminal departed from the system after receiving an echo to the ETE ACK for the last IP without this ETE ACK arriving at the station. This resulted in the reactivation of IP's by the station for this terminal, the routing of these packets in the net, and then the maximum number of transmissions and search by the repeater nearest to the terminal. The protocol simulated in the systems discussed later is such that the last transmission must always be from the station to the terminal. This transmission may be considered as a terminal release packet. Another change in protocol implemented is that whenever possible the terminal acknowledges a sequence of packets rather than individual ones, to reduce the overhead in the direction towards the station.

4.0 RANGE VERSUS INTERFERENCE

For the experiments discussed in the previous section, it was assumed that repeater-repeater range is the same as the terminal-repeater range. This, however, is not always a realistic assumption since repeaters can be placed on elevated areas and can have more power than terminals (especially hand held terminals). Thus, if repeaters are allocated for area coverage of terminals, the repeater range will be higher than terminal range and higher network connectivity or device interference will result.

The problem which then arises is to determine the impact of this interference on system performance. Alternatively, one may seek to reduce repeater transmission power when transmitting in the repeater-station network. To study this issue, two CCSDR systems were simulated, one with high interference CCSDR(HI) and the other with low interference CCSDR(LI). The routing labels of the two systems were the same and are shown in Figure 1. The interference of the CCSDR(LI) system is shown in Figure 2 and the interference of the CCSDR(HI) system in Figure 3. Figure 3 shows only the connectivity of two devices in the network.

The results are shown in Figure 4 and Table 1. Figure 4 shows the throughput of the two systems as a function of time while Table 1 summarizes all other measures of performance. The third row of Table 1 summarizes performance of the high interference system under an improved set of repeater labels. This experiment is discussed in detail in the next section. It is clear that the high interference system is much better than the low interference system. The only measure of the low interference system which is better is terminal blocking which is a direct result of the low interference feature. In fact, CCSDR(LI) is saturated at the offered

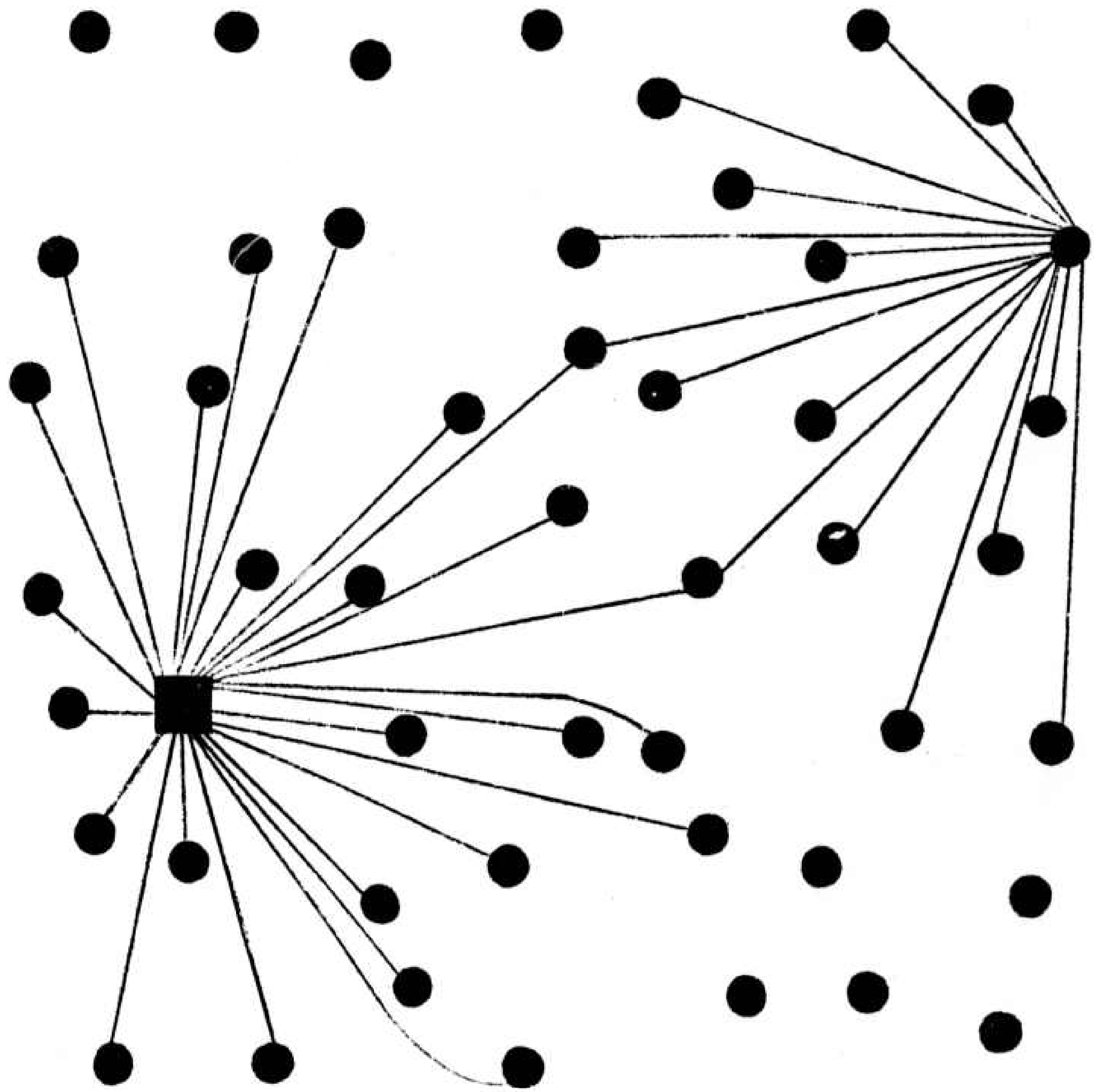


Figure 3.

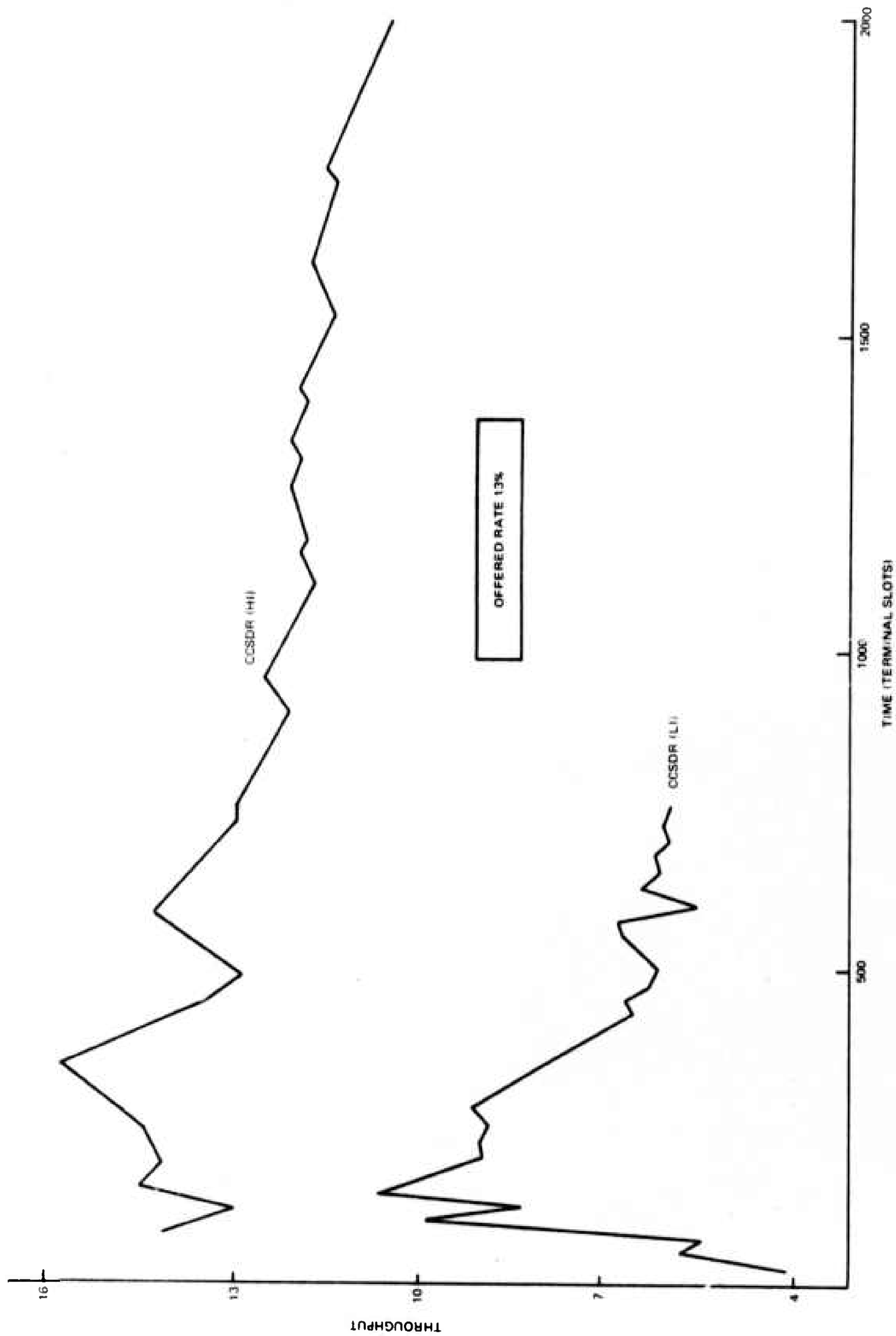

THROUGHPUT
16
13
10
7
4
500
1000
1500
2000
TIME (TERMINAL SLOTS)
CCSDR (H)
CCSDR (L)
OFFERED RATE 13%

Figure 4.

Table 1

	OFFERED RATE (%)	THROUGHPUT (%)	DELAY OF IP TO STATION (Terminal Slots)	RATE OF STATION RESPONSE	PROB STATION BUSY	% OF IP BLOCKED	TOTAL % OF IP LOSS	TERMINALS REMAINING
CSSDR (LI)	13	5.95	40.11	1.14	0.53	2.98	32.83	13
CCSDR (HI)	13	10.55	23.93	1.81	0.43	9.83	9.83	13
CCSDR (HI) (Improved labels)	13	12.14	16.61	2.06	0.50	10.63	11.41	10

traffic rate. This can be seen from the fact that the throughput is decreasing as a function of time, the relatively high total loss, and the low station response.*

---

* The average number of station response packets assumed for these studies is 2.0.

8.0 TWO VS. ONE DATA RATE SYSTEMS

The results of the previous section demonstrate that a better performance system is obtained when repeaters and station use high power to obtain long range despite the interference that results. We now examine the problem of whether repeaters and station should use their fixed power budgets to obtain a long range with a low data rate channel or have a short range with a high data rate channel. The following systems were studied.

- A CCSDR(HI) of the previous section with improved labels which we denote by CCSDR. That is, we take advantage of the high range to improve the routing labels of repeaters and obtain fewer hierarchy levels. The routing labels used are shown in Figure 5, and the connectivity is shown in Figure 3.*

- A Common Channel Two Data Rate (CCTDR) System with the routing labels as in Figure 1 and connectivity as in Figure 2.

In the CCTDR system, the terminal has a low data rate channel (the same rate as in the single data rate system) for communication with a repeater or station. Repeaters and stations have two data rates. The high data rate is used for communication in the repeater-station network. The two data rates use the same carrier frequency so that only one can be used at a time.

The two systems are tested with offered rates of 13% and 25%. The throughput as a function of time for the two runs are shown in Figures 6 and 7 respectively; and the summary of other measures is given in Table 2. The comparison demonstrates that the CCTDR system is superior

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* This system was also compared with CCSDR(HI)

6d

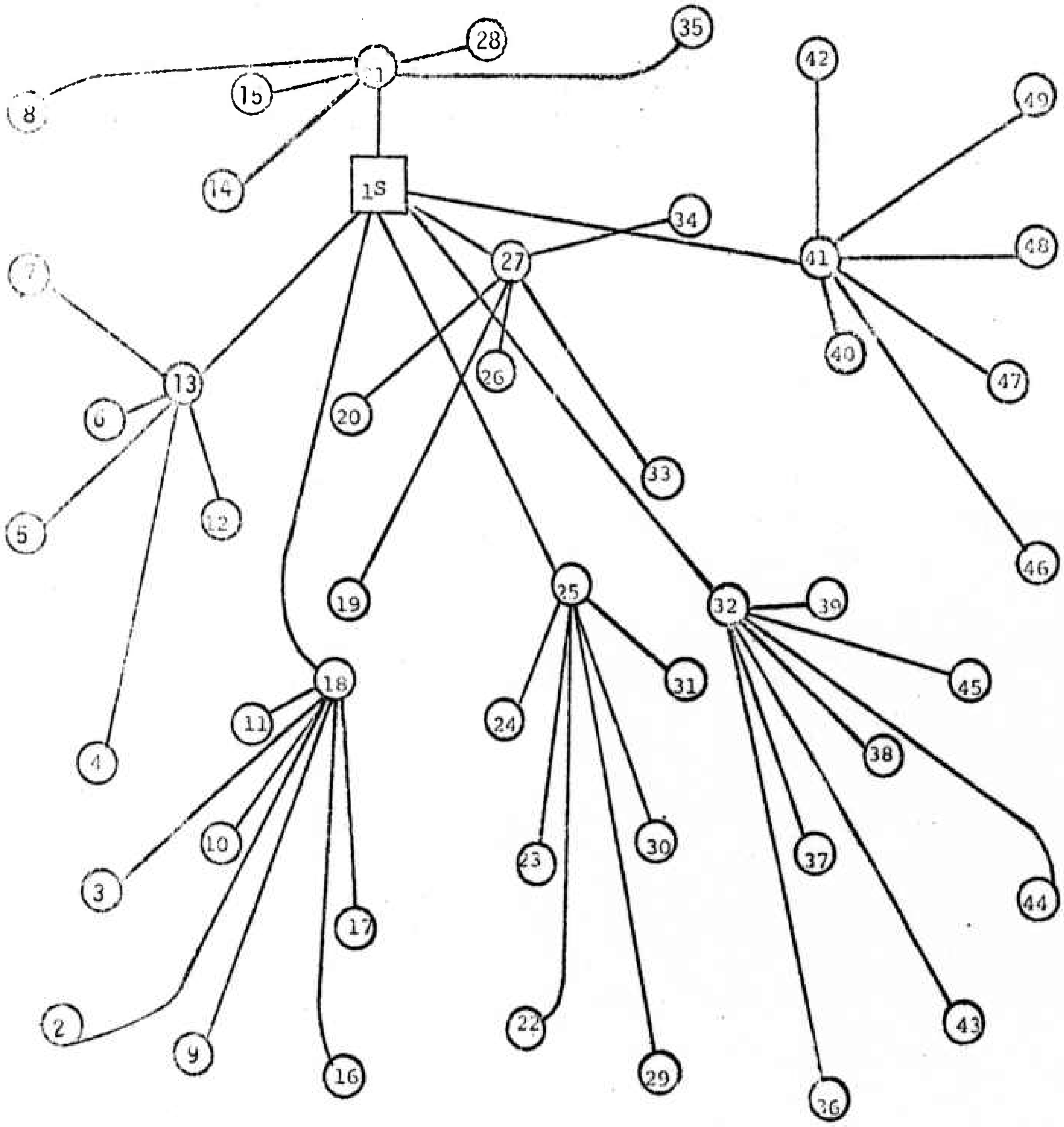

1S
1
8
15
28
35
14
42
49
34
27
41
48
7
13
6
5
12
4
20
26
40
47
33
46
19
25
32
39
45
18
11
24
31
38
10
3
23
30
37
44
17
2
22
9
16
29
43
36

Figure 5.

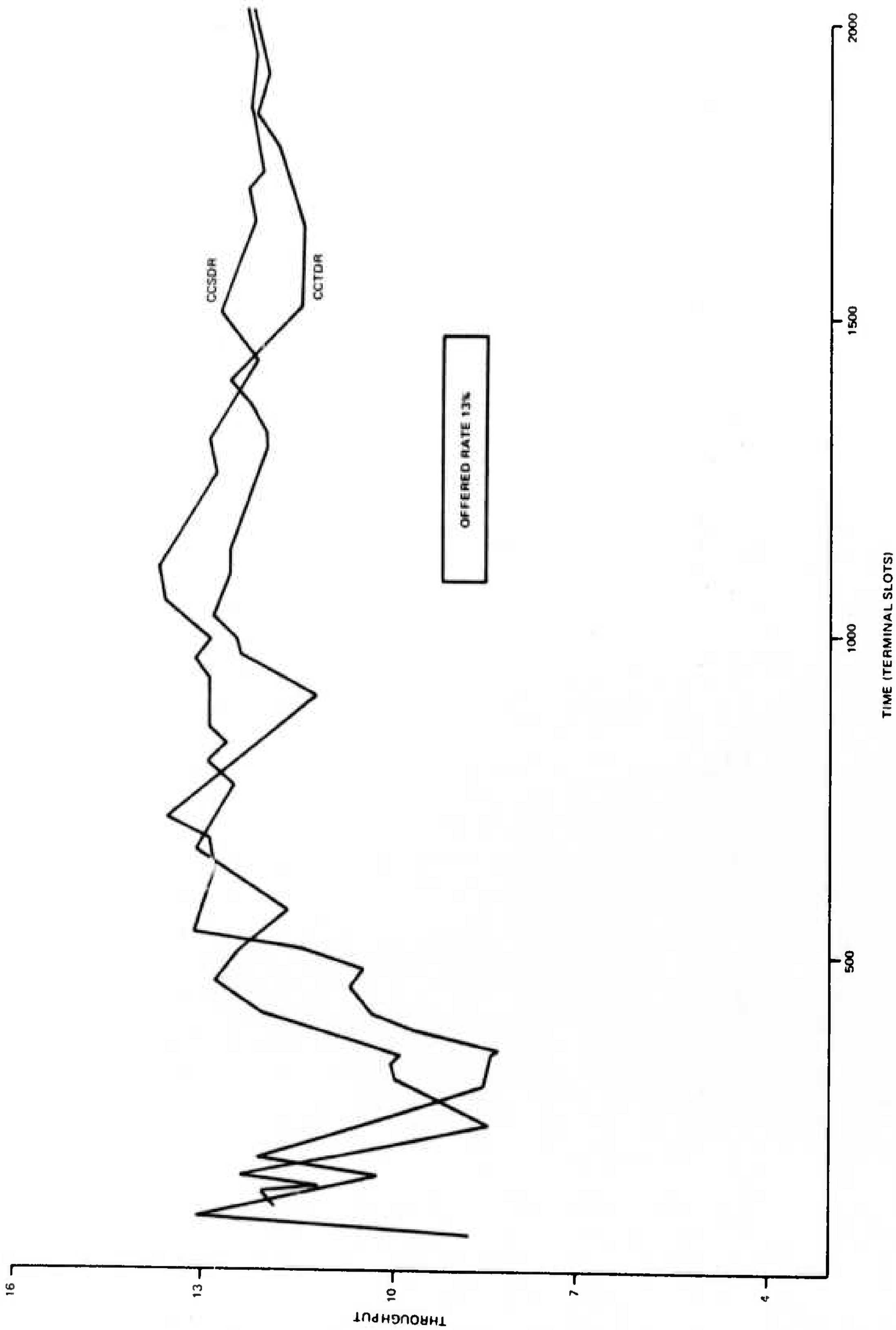

CCSDR
CCTDR
OFFERED RATE 13%
THROUGHPUT
16
13
10
7
4
TIME (TERMINAL SLOTS)
500
1000
1500
2000

Figure 6.

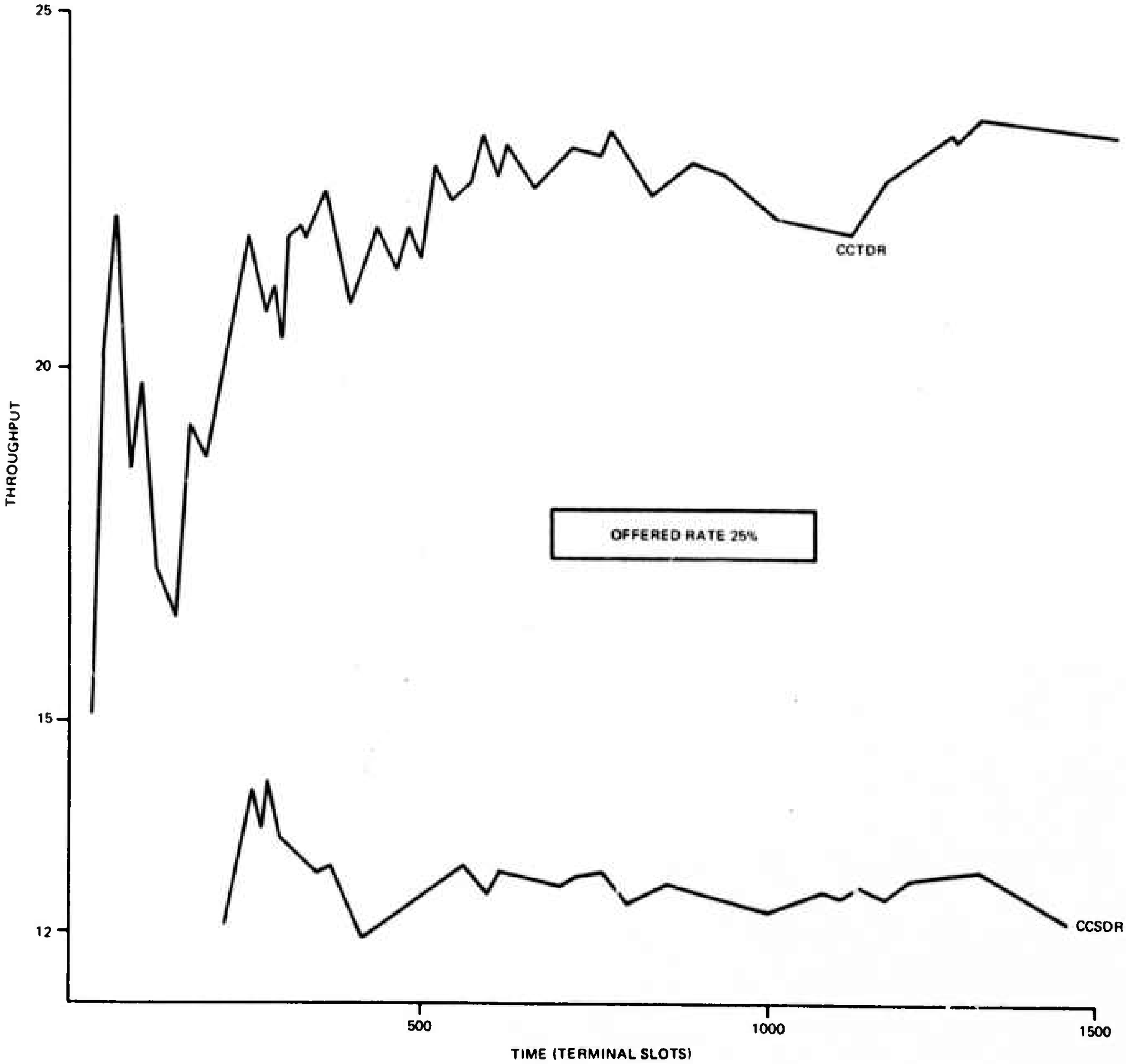

25
20
15
12
THROUGHPUT
CCTDR
OFFERED RATE 25%
CCSDR
500
1000
1500
TIME (TERMINAL SLOTS)

Figure 7.

Table 2

	OFFERED RATE (%)	THROUGHPUT (%)	DELAY OF IP TO STATION (Terminal Slots)	RATE OF STATION RESPONSE	PROB STATION BUSY	% OF IP BLOCKED	TOTAL % OF IP LOSS	TERMINALS REMAINING
CCSDR	13	12.14	16.61	2.06	0.50	10.63	11.41	10
	25	12.20	34.97	1.61	0.48	29.50	32.95	23
CCTDR	13	12.39	4.91	1.99	0.26	1.59	1.59	9
	25	23.33	11.51	1.97	0.31	3.31	3.31	34

to the CCSDR system, in terms of throughput, delay, and other measures. One can see that the CCSDR system is saturated at an offered rate of about 13%.

Effect on Blocking Level

In Table 2 one can see that one reason for the relatively low throughput of the CCSDR system at an offered rate of 25% is due to blocking. Furthermore, the station busy period has decreased. This may suggest that the station may be able to handle more terminals providing they are able to enter the system. To examine this point we ran the CCSDR system with offered rate of 25%, and relaxed the constraint for entering the system. Rather than resulting in better performance, this step resulted in reduction in blocking and increase in delay. The throughput increased to 12.63%, the blocking decreased to 18.35%, and the total loss decreased to 30.73%. On the other hand, the delay increased 57.82%, station busy period increased to .57, and the rate of station response decreased to 1.32.

5.0 MAXIMUM THROUGHPUT, LOSS, AND DELAY OF CCSDR AND CCTDR SYSTEMS

In the packet radio system there is an absolute maximum throughput (independent of loss and delay) because of the interference characteristics. Similar to curves of throughput versus channel traffic, when the relation is known analytically [4], we draw the curves of system throughput versus offered rate to estimate maximum throughput. Figure 8 shows the throughput versus offered rate for CCSDR and CCTDR systems. The curves are linear for low offered rates and saturate when the offered rate increases.

For the CCSDR system one can see that the throughput is practically the same when the offered rate is increased from 13% to 25%. This and the other measurements (see Table 2) (for example the rate of station response) shows that the system is overloaded at a 25% offered rate. On the other hand, the system seems to operate at steady state at an offered rate of 13% (rate of station response 2.06). A rough estimate of maximum throughput for this system would be between 12% and 15%. Similar observations of the performance measures lead to an estimate of between 27% and 30% for the maximum throughput of the CCTDR system. The average delay of the first information packet from terminal to station, and the total loss, as a function of offered rate are shown in Figure 9.

There are many parameters in the simulation program which we have not experimented with (or tried to optimize) and which affect the quantities discussed above. One parameter which is significant in determining the maximum throughput is the average number of response packets from station to terminal. The affect of this parameter has been analyzed in [6] for a slotted ALOHA random access mode. It has been shown that the maximum throughput is increased in the common channel system when the rate of response increases, and the maximum throughput tends

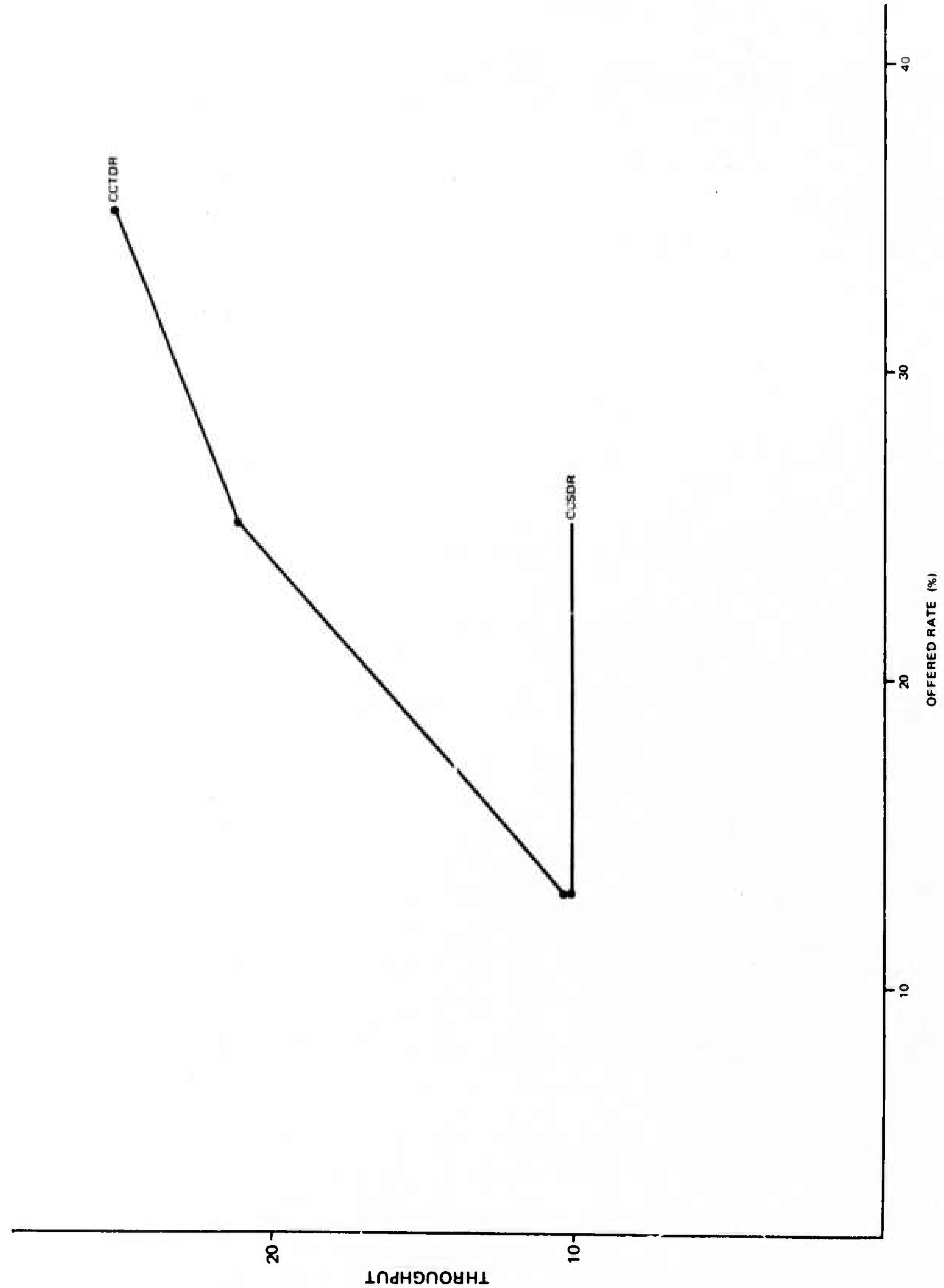


Figure 8.

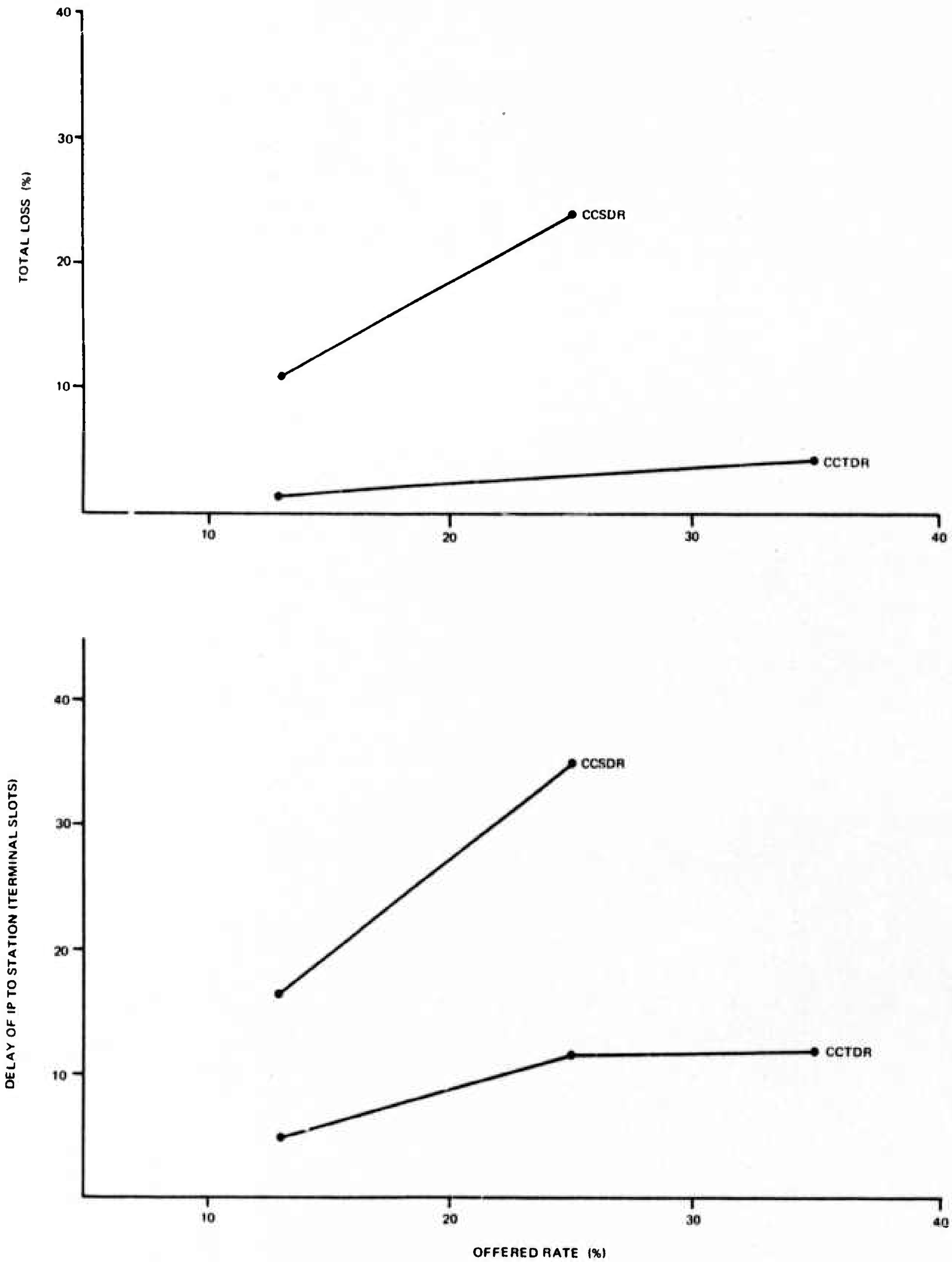

TOTAL LOSS (%)
40
30
20
10
CCSDR
CCTDR
10
20
30
40
DELAY OF IP TO STATION (TERMINAL SLOTS)
40
30
20
10
CCSDR
CCTDR
10
20
30
40
OFFERED RATE (%)

Figure 9.

to infinity. We expect that this parameter has a similar effect for the mode of access simulated. In the results reported here the rate of response is 2.0 which is small compared to usual estimates for terminals interacting with computers. Furthermore, the relatively short terminal interaction increases the traffic overhead of the search procedure, per information packet.

A1. GENERAL SYSTEM DESCRIPTION

Devices: The system simulated contains 48 repeaters and one station. These devices are assigned to corner points in a grid and their exact location is randomly distributed in the neighborhood of the corner points as shown in Figure A1. Terminals originate in time according to a Poisson process. Their geographical locations on the plane are random variables with a two dimensional uniform distribution. A terminal departs from the system when unable to establish a communication path due to congestion, after completing its communications, or after partial communication if delays are high or packets are lost.

Channel and Access Mode: The communication channel is shared both for transmitting in the direction to the station and from the station to terminals. The channel access mode is the non-persistent carrier sense multiple access (CSMA) [4]. That is, when a packet is ready for transmission, the device senses the channel and transmits the entier packet if the channel is idle. If the channel is busy, the device sets a random time after which it senses the channel again.

Capture: A zero capture system is simulated. That is, whenever the reception of more than one packet overlaps in time, none of the packets is correctly received.

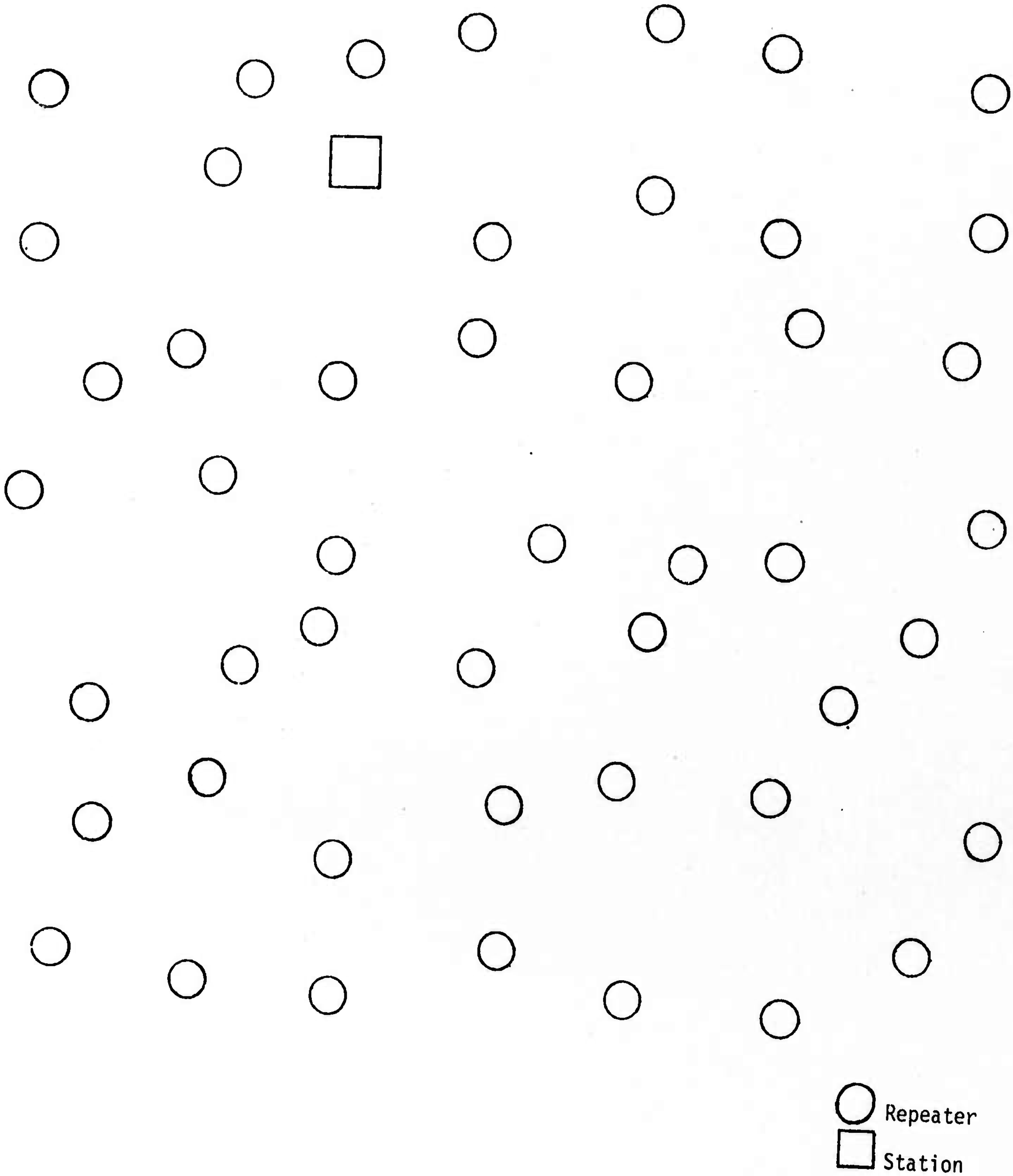


Figure A1. Device Locations.

Packet Types - Simulated

IP	Information Packet
ETE	ETE ACK Short packet (assumed to be 10% of the length of an information packet.)
SP	Search Packet, transmitted by terminal or repeater to all devices and aimed to identify a specific receiver (short packet).
RSP	Response to a Search Packet, transmitted by repeater or station which received a SP and is available for handling packets. This packet contains the label of the transmitting device and is addressed to all devices.

A2. ROUTING AND ACKNOWLEDGMENTS

A formal presentation of the routing and acknowledgment schemes is given in [3]. Devices are assigned hierarchical labels which result in directed (point-to-point) transmission in the packet radio system. Information packets are addressed to particular devices whenever possible, devices use omnidirectional antennas; thus, an information packet is received and interferes with all devices within the effective range of the transmitting device.

Two acknowledgments are used. An End-to-End Acknowledgment (ETE ACK) between terminals and station to ensure message integrity and a Hop-by-Hop Acknowledgment (HBH ACK) along the communication path. The HBS ACK is a passive "Echo" acknowledgment. That is, when device i transmits a packet to device j, it waits sufficient time to allow device j to repeat the packet to the next device; when this transmission by device j is received by i, it considers it as an acknowledgment.

If device j is an end device (station or terminal) which need not repeat the packet, it specifically repeats the header for this purpose.

As discussed in Reference [3], a repeater can attempt "forward" or "forward and backward" search when encountering blocking on the shortest path. In the simulation we use one SP which is equivalent to "forward and backward" search. This simplifies the repeater since it need not construct the various search packets or identify these when receiving one. Also, the search packet has the same format whether transmitted by a repeater or by a terminal.

The Response to Search Packet (RSP) is transmitted by repeaters after a random waiting time. This time randomization is essential in a no-capture system. Otherwise, if more than one receiver wishes to respond to the SP, the transmissions will overlap at the searching device. The station, on the other hand, transmits the RSP immediately after receiving the SP (note that the channel is idle at this time since otherwise the SP would not have been correctly received). The above enables searching devices within range of an idle station to communicate directly with the station.

A repeater makes one attempt to transmit a RSP and if the channel is busy it discards it, rather than store the RSP for further transmissions. This allows control of the level of terminal blocking by specifying the number of transmissions of the SP by a terminal. Thus, when the system is congested in the geographical neighborhood of the terminal, it will not be able to "enter" the system. This feature also makes repeaters more available for handling information packets.

The Maximum Handover Number (MHN) was initially introduced into a packet header as a control to prevent packets from cycling in a loop of repeaters or being propagated to a large distance [3]. The MHN is decremented at every hop and the packet is discarded when the MHN reaches zero. The MHN also has an important role in the HBH Echo ACK scheme and its implications.

In a point-to-point network such as the ARPANET, the channels are fixed so that when node j receives a packet on channel k, it knows the device, say i, which has transmitted the packet to it, and thus can transmit a specific HBH ACK to device i. In the packet radio system, however, this information is not available. Therefore, the HBH ACK must be <u>independent of the path that the packet travels on.</u> The HBH Echo ACK test used includes the following:

(i) identification of the packet.

(ii) tests that the MHN of the Echo received is <u>smaller</u> than the MHN of the packet stored. This ensures that the packet has advanced along the path, rather than being a retransmission from devices that had the packet previously.

The HBH Echo ACK has several advantages over a specific HBH ACK:

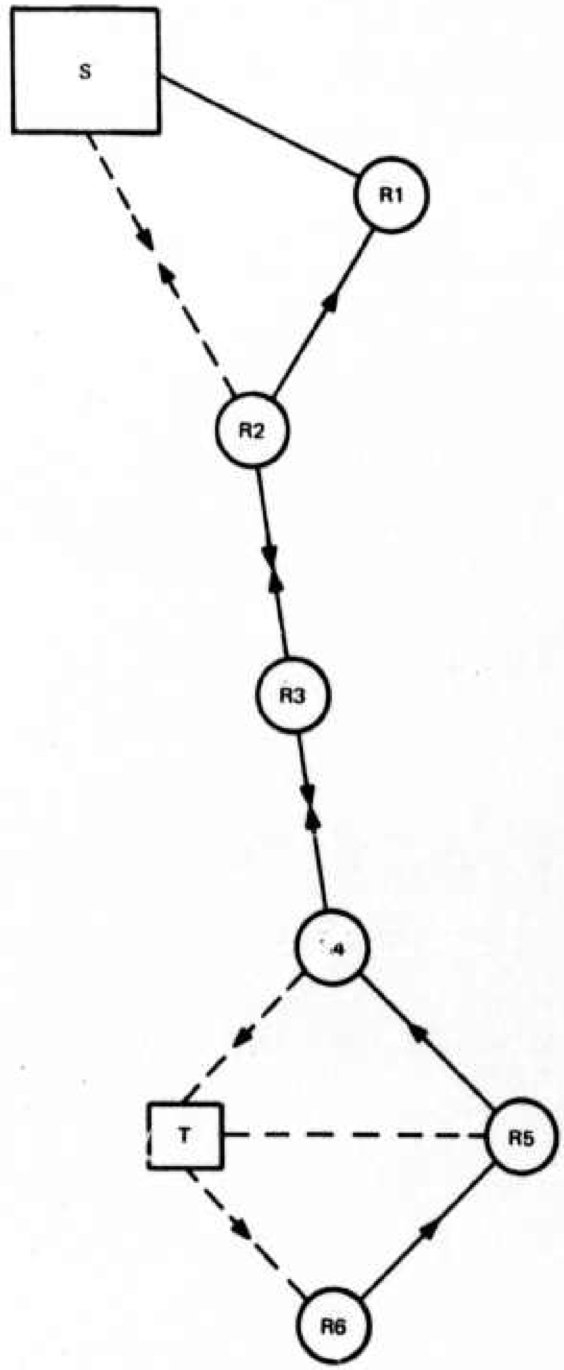
(i) it simplifies the repeater (hardware and software) so that it need not construct and manage acknowledgment packets.

(ii) it reduces the traffic overhead of transmitting specific acknowledgments. This is most significant in a broadcast network.

(iii) it enables acknowledgment of several devices at a time; in particular, all devices which store the packet with a MHN larger than that received are acknowledged.

(iv) it enables shortening of the transmission path, as described below.

Since the RSP's by repeaters are randomized in time, a terminal frequently does not identify the repeater nearest the station within range of the terminal. In fact, if two repeaters are labelled on the same path to the station and both are within range of the terminal, there is a higher probability that the terminal will identify the repeater farther away from the station since, on the average, the latter handles less traffic. Suppose a single data rate channel is used throughout the transmission path between terminal and station, the terminal identifies a packet transmitted to it by its specific terminal ID, and the station can recognize any packet destined for it. Then a communication path as shown in Figure A2 may be established. In Figure A2 the terminal is within an effective range to R4, R5, and R6; and the station within an effective range to R1 and R2. The terminal shown identifies R6 as the repeater to which it transmits. The path from terminal to station will usually be: T-R6-R5-R4-R3-R2-S; and from the station to the terminal S-R2-R3-R4-T. The end devices, terminal and station transmit the Echo ACK immediately after receiving the packet, and <u>transmit it with MHN = 0</u>; thus, they acknowledge all devices which still store the packet. In particular, when R4 transmits a packet toward the terminal, it is addressed to R5; however, the terminal may receive this packet and acknowledge both R4 and R5 simultaneously. Similarly, when R2 transmits toward the station it addresses the packets to R1; when the packet is received by the station it acknowledges both R1 and R2 simultaneously.



NOTE:
THE SOLID LINES INDICATE THE LABELED PATH BETWEEN THE REPEATERS AND STATION. THE DASHED LINES INDICATE THE EFFECTIVE CONNECTIVITY OF TERMINAL AND STATION TO REPEATERS.

Figure A2.

A3. LOGICAL OPERATION OF DEVICES

Terminal: When a terminal originates a message, it begins to transmit and retransmit a SP to identify a specific receiver. If it does not identify a specific receiver after a specified number of transmissions, it departs from the system. We say that such a terminal is blocked. When a terminal identifies a specific receiver, it substitutes the label and MHN sent by the receiver into its IP and begins to transmit its IP. The IP is retransmitted after short waiting periods of time until a HBH Echo ACK is received. At that time the terminal stores the IP for a longer period of waiting after which the IP is reactivated if an ETE ACK is not received.* The terminal is expecting several IP's from the station, each of which is ETE acknowledged by the terminal. When all IP's from the station are received and ETE acknowledged, the terminal departs from the system.

Repeater: A repeater does not distinguish between IP's or ETE ACKS, except for their transmission time. When an IP or ETE ACK is received by a repeater (addressed to it) and the repeater has available storage, it stores the packet, decrements the MHN, modifies the packet label according to the routing and begins to transmit and retransmit the packet, awaiting an Echo ACK. When an Echo ACK is received the repeater discards the packet. When a repeater is not successful in transmitting along the "shortest" path it begins to search for an alternate receiver by transmitting SP's. When one is found it transmits the entire packet to it; otherwise, it discards the packet. When a repeater receives an SP it checks whether it has available storage; if it does, it makes one attempt to transmit a RSP and then discards it. When a repeater receives a RSP it tests whether it needs one; if it does, it uses the label, otherwise it discards it. The repeater currently simulated has buffer storage for two packets; one exclusively

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* We use the term retransmission when a device waits a relatively short period of time (less than 2 IP slots) and is awaiting a HBH acknowledgment. We say that a packet is reactivated when an end device stores the packet, awaiting an ETE ACK. When a packet is reactivated it repeats the complete retransmission process.

used for packets directed toward the station and the second for packets toward the terminal. In addition the repeater can inspect all packets that it receives, which are stored in common arrays by the simulation program. Thus, from a practical viewpoint, buffer storage for three packets can easily be allowed by the simulation program.

Station: The storage organization in the simulated station is shown in Figure A3. There are two queues for active packets. Packets in these queues are active in the sense that they are retransmitted after short random periods of waiting until an Echo ACK is received. The active queue for long packets contains IP's from the station to terminals. Once an IP is Echo acknowledged it is stored in the passive queue for a longer period after which it is reactivated if an ETE ACK from the terminal is not received. The active queue for short packets contains ETE acknowledgment packets to terminals, and these have priority over the long active packets. The ETE ACK packets are, obviously, discarded once an Echo ACK is received. The point-to-point (PTP) network queue simulates the interaction of the packet radio network with a PTP network. When a new IP is received from a terminal, it is stored in the PTP network queue for a random time, after which a response message containing several response IP's to that terminal is generated and placed into the active queue for long packets. The station responds immediately to SP's, and ignores RSP's.

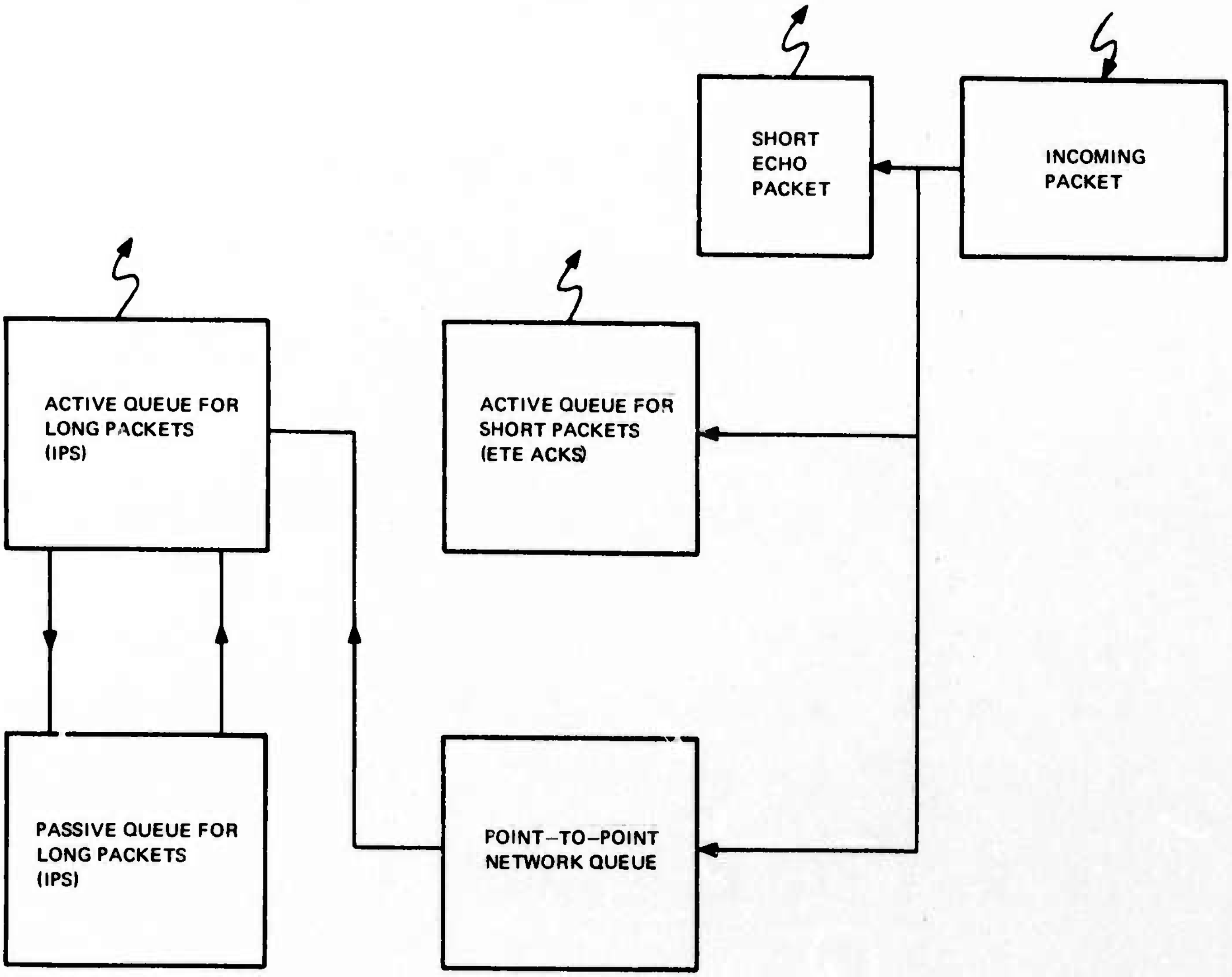

Figure A3. Storage in Station.

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4. L. Kleinrock and F. Tobagi: "Carrier Sense Multiple Access (CSMA)," ARPANET Network Information Center, Stanford Research Institute.

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Appendix B.1

RF Channel Capacity Considerations

I INTRODUCTION

The packet radio network concept was born to meet the requirements for more efficient and rapid communication from mobile digital terminals to central computing resources. Although there may be other possible means to interconnect mobile users, the radio-channel seems to be the only practical one. It is hoped that many of the packet-switching techniques devised for the ARPANET and the ARPA Satellite System will be applicable to Packet Radio Networks; however, it is important to understand that the use of mobile radio terminals places severe constraints on packet radio nets, not experienced by the other ARPA packet nets.

Perhaps the most severe constraint is placed by the limited peak and average r.f. power which can be supplied in a mobile package. Since we desire the package to be hand held someday, the constraint is severe indeed. A second constraint which is much different is the fact that the r.f. spectrum is a limited resource, and must be used efficiently. In digital nets using common carrier facilities the system designer is given his choice of bit-rates, not bandwidth. The common-carrier must worry about using his resources efficiently to provide a digital channel. In the packet radio net the system concept must use bandwidth efficiently. The third constraint considered in this note involves the propagating medium which places constraints on the packet radio channel in the form of noise, multipath, and signal attenuation.

The purpose of this note is to consider the limitations of the r.f. channel as they affect packet radio networks, and to highlight the more important design possibilities.

II GAUSSIAN CHANNEL RESULTS

Shannon's capacity theorem for the Gaussian channel is not directly applicable to typical packet radio channels; however, it can provide some useful insight into some of the myriad tradeoffs which must be considered in design of a packet radio net. Furthermore, the results derived using Shannon's theorem can be adjusted to apply to practical channels.

A. The Capacity Theorem

Shannon's theorem states that the capacity C in bits/sec which can be achieved by a transmitter/receiver pair operating at a received signal-to-noise ratio P_R/N_oW over a bandwidth W Hz is given by

$$C = W \log_2 \left(1 + \frac{P_R}{N_o W}\right)$$

In this case P_R can be interpreted as the received signal power (average) if N_o is the (white) noise spectral power density of Gaussian noise.

Shannon has shown that this capacity cannot be exceeded for signaling at arbitrarily small message error rate, and that a signaling scheme exists which will achieve this rate at arbitrarily small error rate.

B. Bandwidth Limited Channels

Figure 1 is a plot of C vs. $\left(P_R/N_o\right)$ for fixed values of W. For fixed W, large P_R/N_oW C varies as the $\log_2\left(P_R/N_o\right)$. In this region increasing the received power has very little effect on capacity, and the channel

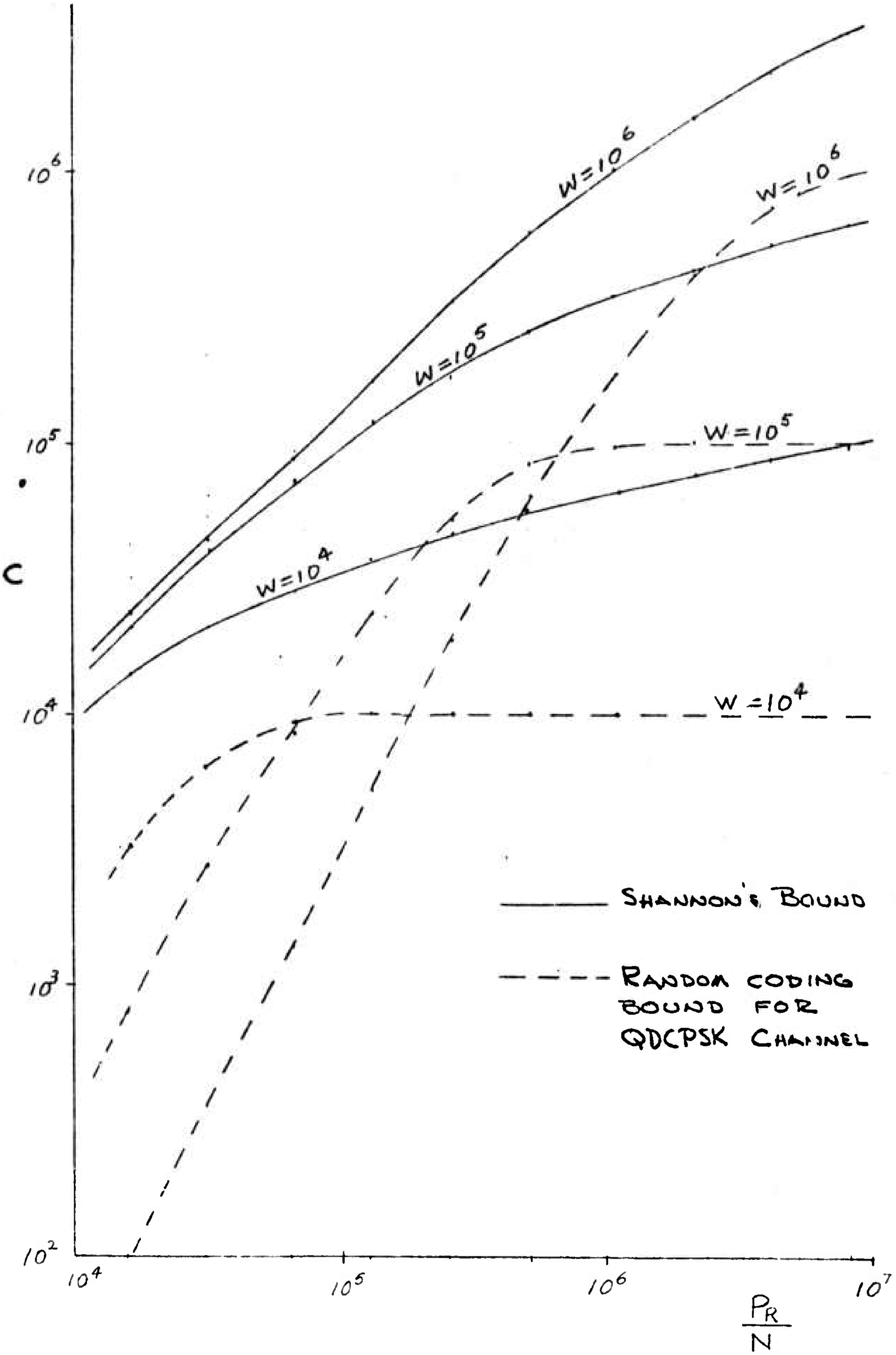

FIGURE 1 BANDWIDTH LIMITED CHANNEL CAPACITY

is effectively bandwidth-limited. Bandwidth-limiting is much more pronounced on practical channels as illustrated by some of the other curves of Figure 1, which are derived in a later section of this note.

C. Power-Limited Channels

Figure 2 shows the capacity variation with bandwidth for fixed P_R/N_o. As bandwidth increases the capacity is limited by $P_R/N_o \left(\log_2 e\right)$ bits per sec. As shown in the figure, more realistic channel models result in a capacity which peaks at some bandwidth, and decreases with increasing bandwidth past this peak. These channels and their implications will be discussed later.

D. Time Sharing Versus Frequency Sharing

The fact that a power-limited capacity region exists suggests that in some cases frequency-sharing a channel may provide greater capacity than time-sharing the channel.

In the packet radio case, the terminals will undoubtedly be both peak- and average-power limited. Usually the terminals will be required to share a channel to some central point such as a repeater. In this case we can derive some interesting results using very simple arguments

Assume a population of n power-limited users of a Gaussian noise channel which is bandwidth limited to W, where the users have sufficient power to cause a received power-to-noise-density ratio of P_R/N_o at a central receiver. Define the channel capacity as the sum of the capacities from each user to the central receiver, $C_S = \sum_{i=1}^{n} C_i$.

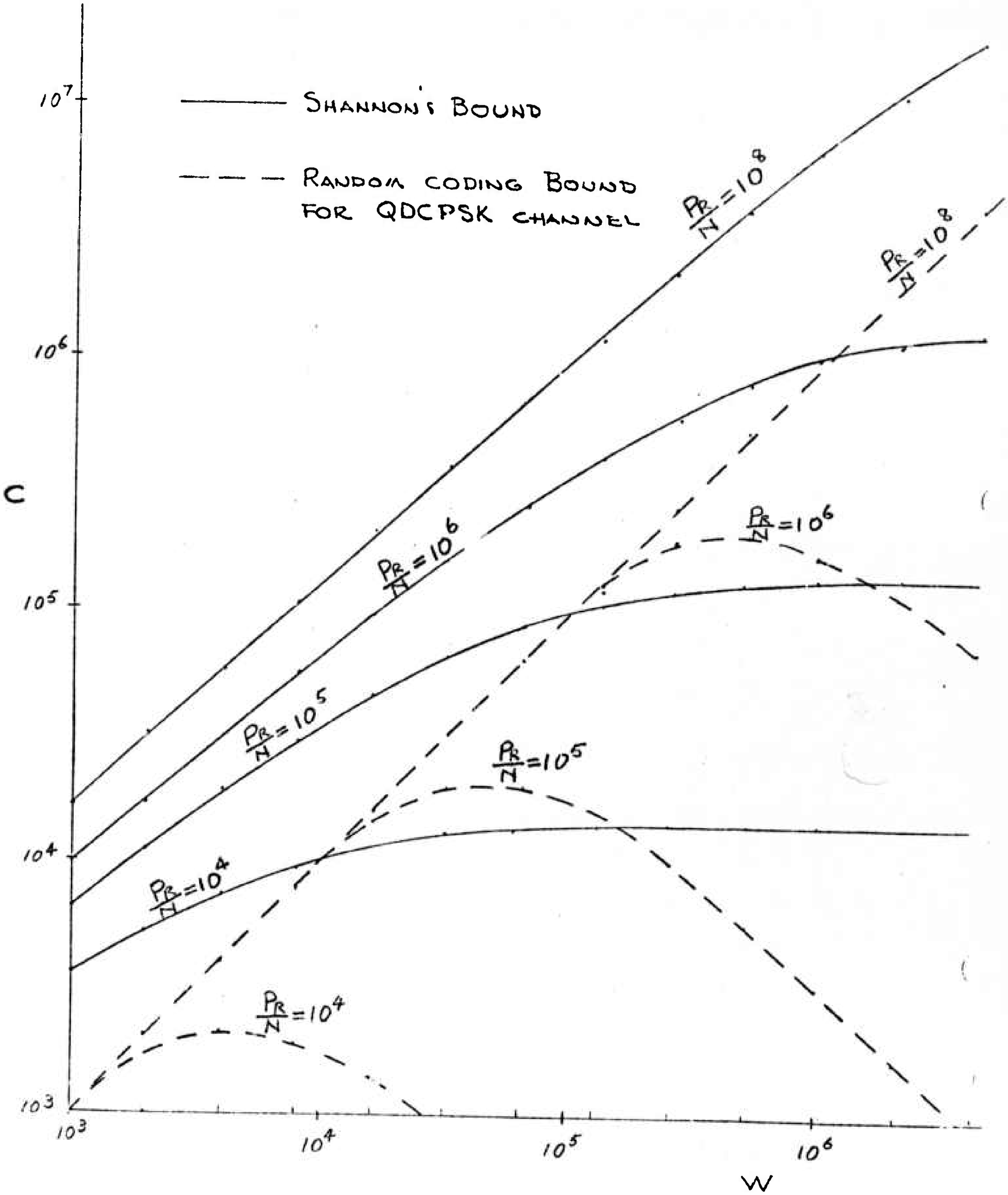

FIGURE 2 POWER LIMITED CHANNEL CAPACITY

Then the capacity of a single user is $C = W \log_2 \left(1 + \frac{P_R}{N_o W}\right)$ if he uses the whole channel, all the time. If he uses the channel a fraction α_i of the time he will have capacity

$$C_i(\text{t-share}) = \alpha_i C$$

the total capacity of the channel, used in a time-sharing mode is

$$C_S(\text{time-share}) = C \sum_{i=i}^{n} \alpha_i \leq C$$

On the other hand, if each user uses α_i of the frequency band, he will have capacity

$$C_i(\text{f-share}) = \alpha_i W \log_2 \left(1 + \frac{P_R}{N_o W \alpha_i}\right)$$

Note that
$$\frac{C_i(\text{f-share})}{C_i(\text{t-share})} = \frac{\alpha_i W \log_2 \left(1 + \frac{P_R}{N_o W \alpha_i}\right)}{\alpha_i W \log_2 \left(1 + \frac{P_R}{N_o W}\right)}$$

$$= \frac{\log_2 \left(1 + \frac{k}{\alpha_i}\right)}{\log_2 (1 + k)}$$

Since

$$0 < \alpha_i \leq 1$$

q
$$C_i(\text{f-share}) \geq C_i(\text{t-share})$$

Hence $C_S(\text{f-share}) \geq C_S(\text{time-share})$, so that frequency-sharing theoretically dominates time-sharing for the inverse broadcast channel. This comes about because more power is available, on the average, in the frequency-sharing mode since we have limited the peak-power of the individual transmitters.

Practically, frequency-sharing requires guard-bands so that some efficiency is lost; and time-sharing requires time-guard bands so that some efficiency is lost.

It is possible to combine frequency and time-sharing in order to ease the implementation problem. If we divide the channel into m sub-channels and time-share the subchannels, then a capacity of

$$\hat{C}_S = \sum_{i=1}^{n/m} \alpha_i \frac{W}{m} \log_2\left(1 + \frac{P_R m}{N_o W}\right) \leq W \log_2\left(1 + \frac{P_R m}{N_o W}\right)$$

if the channel is power-limited, so that

$$\frac{P_R m}{N_o W} < 1$$

then

$$\hat{C}_S \cong m \frac{P_R}{N_o} \log_2(e)$$

This is an increase by a factor of m over simple time-sharing.

If the channel is bandwidth-limited, so that $P_R/N_oW > 1$, then frequency-sharing will not be significantly better than time-sharing.

This suggests that if the terminal power, range, noise, etc. are fixed, and we have the opportunity to use more bandwidth, we should time-share the band until we are power-limited; i.e., reach a signal-to-noise ratio of about 1 $\left(P_R/N_oW = 1\right)$. At that point we should start splitting the band into sub-channels; i.e., we always maintain the signal-to-noise ratio greater than 1.

Practical channels reach the power-limited region at much higher signal-to-noise ratios (smaller bandwidths) than the Gaussian channel, so that band-splitting will be efficient at smaller bandwidths.

E. Line of Sight (L.O.S.) Range

Shannon's bound can also provide some insight into the question of L.O.S. range. Suppose that we assume that a single repeater must serve an area in which there are many terminals. Let

δ = density of terminals

B = source rate per terminal (bits/sec)

S = channel utilization

Then the capacity needed in a single repeater

$$C_N = \delta \pi R^2 \frac{B}{S}$$

to handle all terminals in an area of radius R.

The typical signal strength available from a terminal at range R varies as $\frac{1}{R^4}$. Thus the capacity available within a circle of radius R is

$$C_A = W \log_2 \left(1 + \frac{P}{N_o W R^4}\right)$$

where P is the power received at 1 mile.

It is logical to choose an L.O.S. range so that $C_A = C_N$ (ignoring interference). This can be found by plotting C_A and C_N vs. R, as in Figure 3.

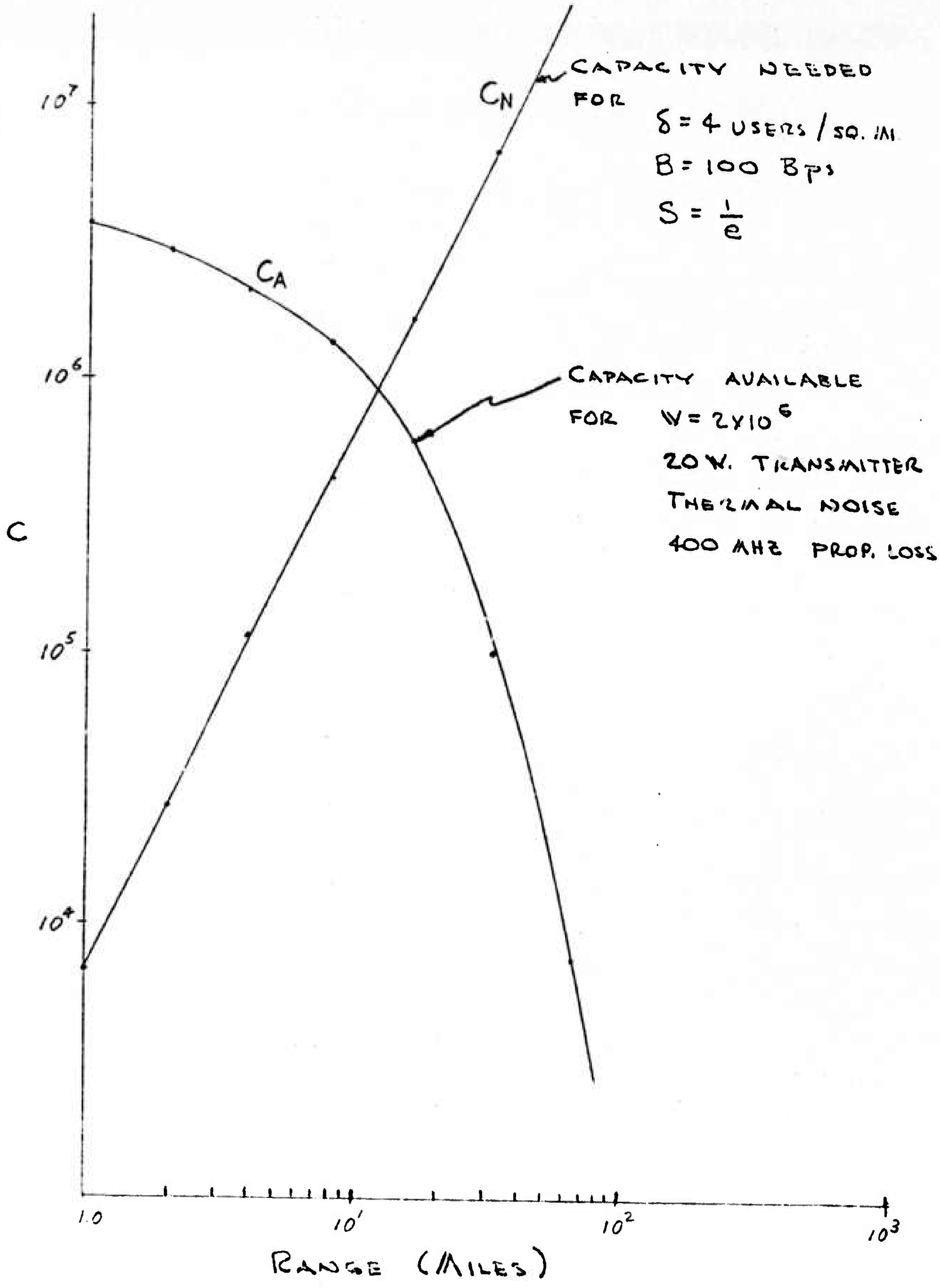

FIGURE 3 CAPACITY VS. MAX. RELAY-TERMINAL RANGE

In this figure we have assumed

$$\delta = 4 \text{ terminals/mile}$$

$$B = 100 \text{ bps}$$

$$S = 1/2e$$

$$W = 2 \times 10^5 \text{ Hz}$$

$$P/N_o = 9 \times 10^{10}/R_4$$

This corresponds to a terminal transmitter power of about 20 watts at 500 MHz depending on antenna design, etc. (PRN #4 example uses 4 W at 500 MHz to achieve this P/N_o).

The results from this figure would indicate that a range of about 10 miles is suitable for this example. We should not take this number too seriously until a more realistic channel model is introduced; however the computation illustrates clearly that terminal peak power, terminal density, source bit rate, bandwidth and utilization are all interrelated in a rather simple and direct way in determining L.O.S. range.

The capacity of less ideal channels is not so easily calculated. Several realities must be considered:

1. The real channel will include additive non-Gaussian as well as Gaussian noise.
2. The real channel will have random amplitude (space-fading).
3. The real channel will have multipath effects.
4. Modulation and coding schemes to achieve capacity are too complex and expensive for implementation.

Although the first effect will be very important, we must defer consideration of non-Gaussian noise to another note, and restrict this note to Items 2, 3, and 4 above.

An excellent discussion of these considerations is given in Wozencraft and Jacobs,[1] and need not be repeated here. More to the point is a specialization to the packet radio problem. For this discussion we introduce the functional block diagram of the channel shown in Figure 4.

The transmitter must accept a message from the source and select a waveform to represent the message. For our purposes a message is a packet. The waveform-selection process can be broken into two steps: encoding and modulation.

The waveform is frequency-translated, amplified and radiated. The propagation channel distorts the waveform and adds noise. The receiver uses the distorted, noisy replica to try to decide which message was

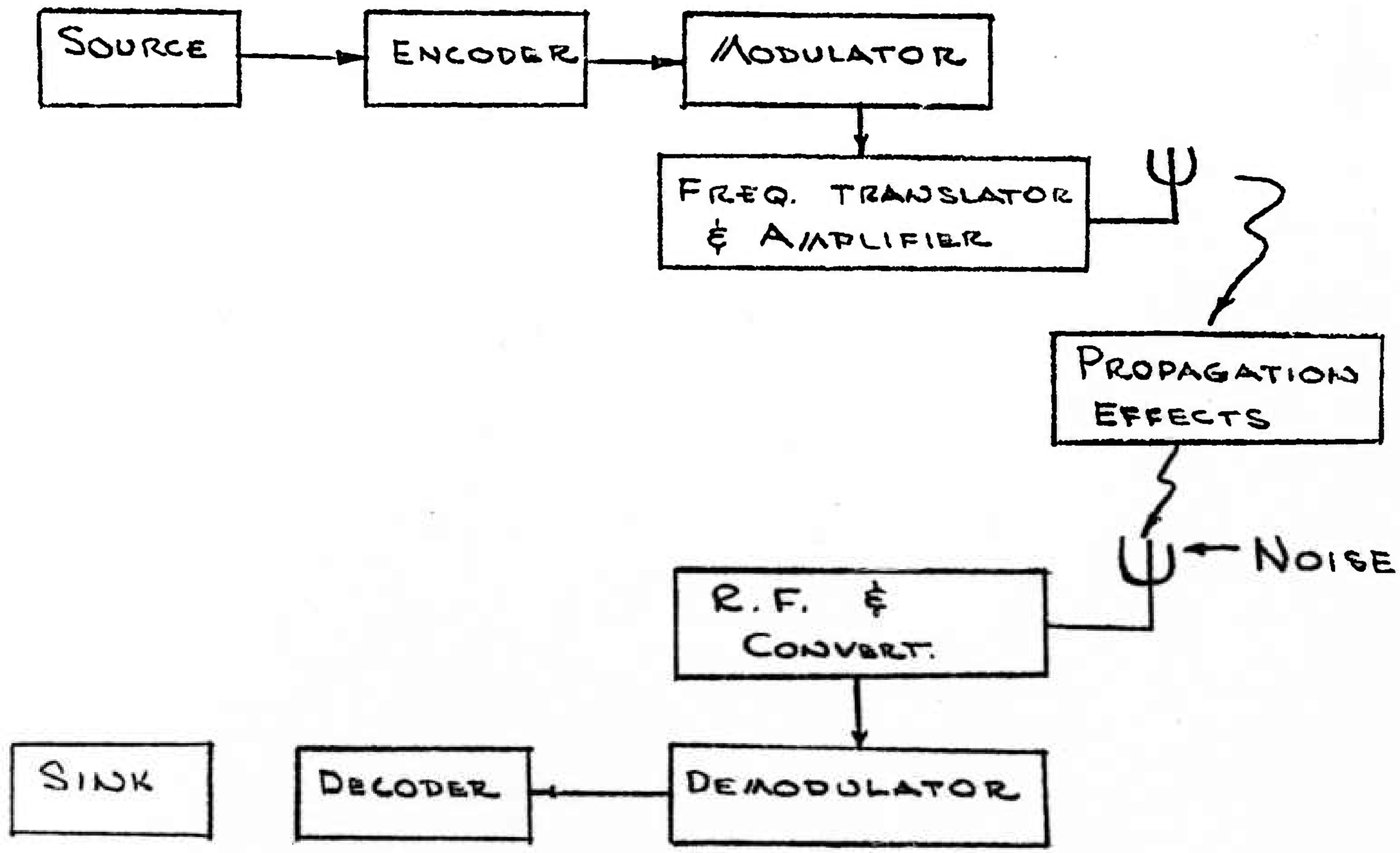

Figure 4 FUNCTIONAL BLOCK DIAGRAM OF CHANNEL

sent. In the most general case it is not possible to break the decision process into demodulation and decoding, but for our purposes this breakdown will be justified.

Design of the rf part of the system will require selection of the coder/decoder and modulator/demodulator as well as selection of frequency, peak power, antennas, etc. Selection of coder and modem should be made to best use the propagation channel.

Our purpose here is to discuss some of the important considerations for rf design and estimate their relation to the maximum rate at which messages can be communicated with some specified error rate.

A. A Low-Risk Channel

Before discussing more realistic achievable rates, it is worthwhile to calculate the rate which can be achieved with a simple low-technical risk terminal-to-relay link. For this purpose we define such a link below:

Transmitter Power	13 dBW (20 W)
Transmitter Antenna Gain	2 dB
Receiver Antenna Gain	4 dB
Receiver Noise Figure	7.8 dB
Operating Frequency	400 MHz
Modulation Type:	Noncoherent F.S.K.

B. Propagation Considerations

In order to obtain a conservative bound on link performance, we assume that the link operates in an urban area where the propagation losses vary with range as $1/R^4$ and the received field strength at range R is a random variable with a Rayleigh distribution. The mean propagation loss is taken from packet radio Note 4 as 145.6 dB at 10 miles.

Measurements made by Turin[2] and Schmid[3] indicate that the multipath delay-spread in an urban/suburban area is 4 to 10 μs (let us assume 5 μs for this note). So long as the bit durations exceed about twice the delay-spread, performance will not be significantly degraded by multipath; hence FSK bits rates up to 100 kbps can be supported without resorting to any form of multipath-equalization. (Higher bit rates are possible with other techniques - see below.)

C. Communication Rate

Suppose that we define acceptable performance in terms of the probability that a 1,000 bit packet will be incorrectly received. If we require a packet error rate less than 1 in 10^{-2}, and if we assume independent bit errors, then we will require a bit error rate of less than 1 in 10^{-5}. A non-coherent FSK modem will provide this error rate for all SN Ratios exceeding about 13 dB if the noise is Gaussian and there is no fading or multipath distortion. Although the received signal strength in an urban area varies randomly as the receiving antenna is moved over a small area, it does not fluctuate in time, except very slowly, so long as the receiving antenna is stationary. We assume that the slowly-moving user moves his antenna to avoid deep nulls, or that he is provided with some form of diversity to overcome the deep nulls. That is, we ignore the fading effects. In this case, a 10^{-2} message error rate will be achieved so long as

$$\frac{E_S}{N_O} \geq 23$$

For the 20 W transmitter assumed,

$$\frac{P_R}{N_O} = \frac{9 \times 10^{10}}{R^4}$$

For rectangular pulses of duration T, and amplitude A

$$P_R = \frac{A^2}{2} \quad \text{and} \quad E_S = \frac{A^2 T}{2} = P_R T \ .$$

Hence $$\frac{P_R T}{N_O} \geq 23 \quad \text{or} \quad T \geq \frac{23}{9 \times 10^{10}}$$

so that the maximum bit rate will be

$$B = \frac{1}{T} \leq \frac{4.10^9}{R^4} \ .$$

Considering multipath to provide an upper limit of 100 kbps, the achievable bit rate for a low-risk link is:

$$B \leq \min \left(10^5, \frac{4 \times 10^9}{R^4} \right)$$

Figure 5 shows B versus R for the non-coherent FSK link described above. The bit rate is multipath limited out to a range of about 14 miles.

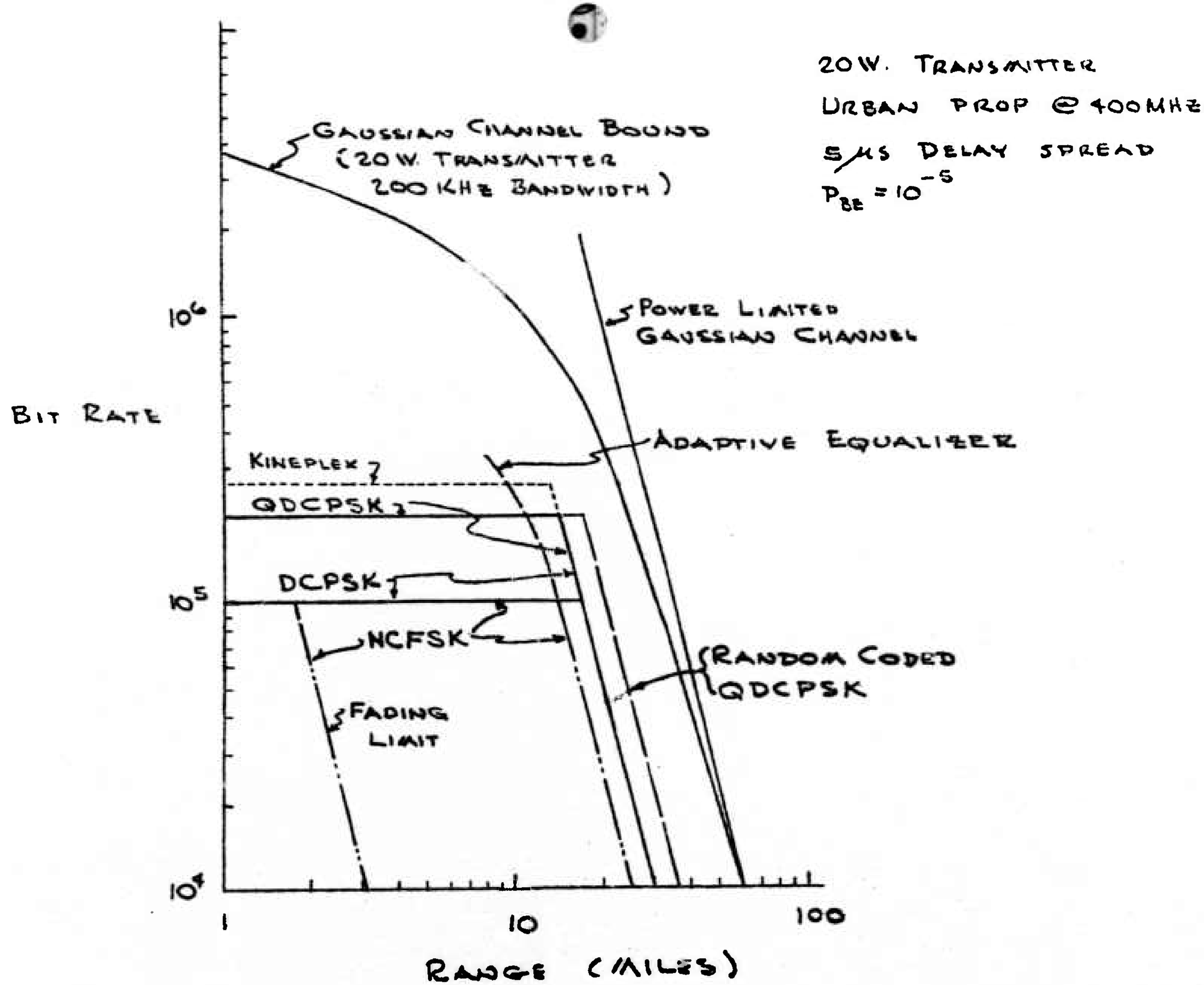

Figure 5 COMPARISON OF SOME RF CHANNELS

Beyond this range it becomes power-limited, and bit rate decreases rapidly with range.

An FSK signal occupies about 3B Hz where B is the bit-rate. Thus in the multipath-limited region the width of the frequency band used by this low-risk link is about 300 kHz. For comparison we have shown Shannon's bound for a 200-kHz bandwidth Gaussian channel, where the same transmitter power, propagation losses, etc., have been assumed. It is clear that the non-coherent FSK system can be improved upon.

To show that things could be worse, the curve for FSK performance over a fading channel is also shown. This would apply if, for example, the users moved rapidly, or were not allowed to avoid nulls.

D. Improved Channels

Also shown on Figure 5 are curves of achievable bit rate versus range for a variety of improvements. Techniques to use power more efficiently include better modulation, such as differentially coherent phase shift keying (DCPSK) with two or four phases (QDCPSK) and the application of error correction coding. Both DCPSK and QDCPSK have error probabilities given by

$$P_e = 1/2 \exp \left(\frac{- P_R T}{N_O} \right)$$

where T = duration of one bit

hence they require 3 dB less power than NCFSK. Coherent PSK is not shown because it is more complicated to receive than DCPSK, and it does not provide significant improvement at the required message error rate.

The random coding curve for DCPSK with a quantized receiver can be found by considering the capacity of a binary-symmetric channel with error probability p:

$$C = B\left[1 - \log_2\left(1 + 2\sqrt{p(1-p)}\right)\right]$$

If the expression for P_e is substituted in this equation we can obtain an expression for C in terms of B and P_R/N_0R^4. For a fixed P_R/N_0R^4 there is an optimum value of B which we found by plotting a series of curves of C versus B. The curve shown on Figure 5 is the maximum value of C for each R, with P_R/N_0 fixed.

Practically, it is very difficult to achieve this curve, since it requires complex decoding, and assumes infinite block lengths. On the other hand the performance of an uncoded QDCPSK system is almost as good and relatively easily achieved.

We should note that in the power-limited region there are many types of modulation which reduce to DCPSK or QDCPSK and will provide equivalent performance. Among these are Kineplex, MSK, and several varieties of spread-spectrum modulation.

At shorter ranges, each modulation scheme runs into a multipath limit. We have shown this on the curves at a delay-apread of 5 μs assuming that intersymbol interference becomes too great whenever a symbol is shorter than twice the delay-spread. This is probably a conservative bound, and will be investigated in more detail, but it is probably not off by more than a factor of 2.

The multipath limit is raised by modulation schemes which transmit multiple bits per symbol. The limit for QDCPSK is just twice that for DCPSK because the 4-phase signal carries 2 bits per symbol. Also shown is the limit for a 16-tone QDCPSK Kineplex modem. In this case one pulse on each tone carries 2 bits, and since there are 16 tones, there are 32 bits per symbol; however, for acceptable operation the delay spread cannot exceed about 2 percent of the symbol duration. Thus the multipath limit is about 260 kBps. Lest someone think that this is too favorable a comment on Kineplex (invented by Collins) I should note that it is a complex modem to construct, and I am not sure that anyone has built one at bit rates exceeding about 2×10^4 B ps. Kineplex is like MSK in that it achieves efficient use of the bandwidth, since it achieves roughly one bit per Hz. QDPSK can be filtered to approach this, but usually takes slightly more than one Hz per bit. DCPSK uses roughly 2 Hz per bit, orthogonal FSK uses about 3 Hz per bit.

Modulation schemes which perform at rates exceeding 1 bit per Hz are possible. Such schemes require multi-level symbols. These are undesirable for the urban channel where amplitude will be random; however, at very high received power levels (short range) it might be possible to use multilevel signals. It would probably be necessary to equalize the channel to use such schemes. If the user moved slowly so that the channel did not vary from bit to bit, an adaptive equalizer might be possible. The use of such an equalizer should be studied if operation at short ranges (or with higher power) is considered.

A curve showing the performance achievable with an adaptive equalizer is also shown in Figure 5. This was found from Monson's[4] results simulating a decision-feedback equalizer in multipath. The modulation scheme assumed by Monson is binary phase-shift keying with coherent demodulation. This modulation technique cannot be used on the packet radio channel where the received phase is unknown, but for low error probabilities DCPSK performance should be approximately the same.

IV CONCLUSIONS

Examination of supercurve leads to some interesting conclusions about the rf design.

a. The terminal-to-relay link will be either power-limited or multipath limited depending upon the terminal power, and the ratio of the terminal density-source bit rate product to channel utilization. For practical terminals this break-point will occur when the $\delta B/S$ ratio is about 300.

b. For smaller ratios the link will be power limited and only an increase in power or an improvement in modulation scheme will improve performance. In this region the modulation technique should probably be some version of QDCPSK (including Kineplex, MSK, and Spread Spectrum formats).

c. For larger $\delta B/S$ ratios the link will be either bandwidth limited or multipath limited, depending upon the available bandwidth. In this region only the use of anti-multipath techniques will improve performance. Use of modulations transmitting multiple bits per symbol such as QDPSK or Kineplex is important in this region, since such signals can increase the bit rate by a factor of 2 or more. Adaptive equalizers might increase bit rate by an additional factor of 1.5 to 2. This amounts to an increase in the relay range by a factor of 1.2 to 1.4. Anti-multipath techniques may not be worth the complexity, and a probable upper bound on the practical bit rate is 200 kBps.

d. If we are power-limited to 20 W and have 200 kHz available, then so long as our user distribution is such that we wish to operate each relay at ranges less than about 15 miles we should use a single frequency per relay.

e. If our user density is such that longer ranges are desired, then we should use two frequencies per relay at 15 miles, three at 19 miles, etc.

f. In general, for a bandwidth W, a range R, and a power P_R we can find curves of maximum bit rate versus range such as Figure 5. We will best use the bandwidth and power if we use n channels when the (QDCPSK) maximum bit rate is between W/n and W/n+1. Of course complexity considerations may override considerations of efficient channel bandwidth usage.

g. It is of critical importance to the RF design to determine the peak power available in a hand-held terminal, the bandwidth which might be allocated to such a system, and the user geographic distribution and average bit rate.

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Appendix B. 2

Measurement Program for Packet Radio
Channel Characterization

Measurement Program for Packet Radio Channel Characterization

I BACKGROUND

Many new or previously untried concepts enter into the development of a full-scale packet radio network. Among the first questions that must be resolved are those basic to the ability of a wideband system--as packet radio is presently conceived to be--to operate in an urban/suburban situation. This environment may be inimical to the packet radio network from two standpoints:

- The impulsive man-made noise radiated from power lines from automobile ignition systems and from other sources may cause high packet error rates.
- The existence of many radio-reflective surfaces, such as buildings, in the vicinity of a receiver may result in unacceptable multipath.

The published literature does not provide adequate information to determine the effects of impulsive noise and multipath upon the packet radio network; thus, answers to these questions must be sought in a measurement program.

We intend to perform the necessary noise and propagation measurements for the packet radio network in the actual urban and suburban locations that will serve as points in the packet channel test system, thereby ensuring the applicability of the measurements to the later experimental situation. Much of the equipment used in the measurement program will be smoothly integrated into an eventual test facility. In particular, a mobile lab vehicle, needed to house the measurement equipment, will, along with its power generator, antennas, and test equipment, be phased into the test facility as a mobile terminal or perhaps as an interim repeater. A judicious selection of the minicomputer to be used as a part of the data-acquisition system will allow this same machine to be used, after the completion of the first year's measurements, as the controlling computer at a packet repeater site or as a data-collecting and processing unit for further link tests.

The measurements to be made in the first year are those pertaining to the noise environment and those concerning the propagation path. They will be made at typical urban, suburban, and rural locations, and near the two candidate frequencies (approximately 430 and 1325 MHz) at which the packet radio network may operate. The receivers and transmitters for propagation measurements are to be provided by ARPA and integrated into a data-acquisition system. As mentioned, this system will be housed in a van containing a power generator; the van will also contain a minicomputer for control and data processing, and other test and maintenance equipment.

Both the noise and propagation measurement programs will be designed to provide results useful not only to packet radio but to other urban radio systems as well. The results will be published for wide distribution, as they will extend the knowledge of the urban/suburban radio channel.

II QUANTITIES TO BE MEASURED

A. Noise

At the locations and at the frequencies that a packet terminal or relay is expected to operate, the major sources of radio noise are the highly impulsive incidental radiations from man's activites and equipments. Terminal locations can be expected to be fully immersed in this noise environment. Preliminary measurements by SRI at 400, 1200, and 2900 MHz indicated that automobile and other man-made noise was almost always present at the lower frequency and less likely as the frequency was increased. The 2900 MHz frequency was relatively quiet but the impulsive noise greatly increased when automobiles passed by.

Focusing on this impulsive type of noise then, we postulate a working model to guide our selection of quantities to be measured.

To determine precisely what noise measurements are needed we must first create a model of how noise affects the detection process. We do this in the context of estimating the probability that at least one binary error occurs in a packet. Let us divide the problem into two parts. First the probability of a bit error is estimated given the receiver/detector impulse characteristics, certain descripters of the impulse such as its amplitude (A), its width (W) at some important amplitude threshold, and its time delay relative to the bit interval. From these descriptors, and some assumptions about relative phase of noise vs. signal, $P_B(A, W)$ can be calculated where

$$P_B(A, W) = P_r \left[\text{bit error} \mid \text{impulse characteristics } A, W\right] .$$

It is also assumed that one noise spike affects but one bit. In that case the probability of an error for any bit is given by

$$P_E = \int\int_0^{\infty} P_B (A, W) p(A, W) \, dA dW$$

where $p(A, W)$ is a joint probability density on A and W. To the extent that the impulses are bandlimited there will be a deterministic relation between A and W. In this case $p(A, W)$ becomes a single density on pulse amplitude $p(A_p)$. Our measurement should examine this condition.

The second part is to find the probability of an error in a packet, given P_E. To do this we can choose either to model the process as a more or less continuous one looking at each bit interval, or we can model the process as one in which large but infrequent noise spikes cause essentially all errors. If we choose to examine each bit, and assume that each examination is independent (regards noise) from all others, the probability of at least one error in m bits is

$$P_{PE}(m) = 1 - (1 - P_E)^m .$$

A more realistic model on the other hand would consider only discrete large-noise-spike events. If the number k of such spikes within time T is given by $P_k(T)$, then

$$P_{PE}(T) = \sum_{k=1}^{\infty} P_k(T) \left[1 - (1 - P_E)^k\right]$$

We shall proceed with the second, more realistic model and therefore must measure the following statistics:

(1) $p(A, W)$--the joint probability density on amplitude and width. This measurement may point up a relation between A and W in which case a joint distribution would no longer be necessary. The joint density can be defined by taking the distribution of width at several amplitude thresholds. The complex impulse response of the receiver must also be obtained.

(2) $P_k(T)$--the probability of exactly k pulses in time T. This will be compiled by determining the distribution of intervals between noise spikes, p(T), which is expected to be exponential.

But if the noise impulses are always band limited, or more precisely their spectrum is flat over the receiver bandpass, then it is not necessary to specify them as a joint amplitude/width distribution. The impulse noise process can be viewed as a collection of randomly weighted impulse functions with random arrival times. Or

$$i(t) = \sum_{-\infty}^{\infty} a_j \delta(t-t_j)$$

where a_j and t_j are the random weight and delay of the jth impulse. Putting this process into a receiver filter results in

$$f(t) = \sum_{-\infty}^{\infty} a_j h(t-t_j) e^{i\phi_j}$$

where h(t) is the impulse response of the filter. This is the same as the filter output assumed by Bello and Esposito.[1] The phase term is included to account for the phase of the impulse resulting from the band-limiting action of the filter.

Because of the random phase (time) relationships between separate noise impulses, envelope detection will prevent our associating output with input values whenever pulses are closer in time than the inverse of the filter bandwidth. If on the other hand $(t_j - t_{j-1})$ is greater than the reciprocal of the bandwidth, then the phase term is of no consequence and, knowing the filter impulse response, the jth term of f(t) can lead to the impulse weighting factor a_j.

Thus another variable whose measurement would help define a minimum-parameter characterization of impulse noise is the distribution of amplitude weights a_j. Specifically, we seek $p(A_p)$ the probability density on peak amplitude. With $p(A_p)$, $p(T)$, the impulse response of the receiver, and assumptions about uniform phase distribution among noise pulses, a wideband noise process can be synthesized.

The joint density $p(A, W)$ will be used to test the role of the receiver impulse response in determining the width W at various amplitude levels. It will also be used to test the 1 bit/noise-spike assumption. It should not be overlooked that to the extent that the output of the envelope detector used in the measurement system simulates the input to the eventual decision device, no extrapolation to input noise parameters is necessary. In this instance even overlapping noise pulses can be included in the noise characterization.

In addition to these impulse measurements, we will also measure two broadly useful basic noise parameters. These are the mean noise power and the ratio of rms to average noise voltage. The latter is a measure of the "impulsiveness" of the noise. Both have been widely measured for atmospheric noise and are coming to be recognized as useful in the area of man-made noise also.

B. Propagation

To adequately characterize the packet radio channel we must add a set of propagation descriptors or models to the noise parameters already mentioned. Most obviously a knowledge of attenuation rate must be obtained to determine areas of coverage for certain radiated power levels and antenna heights. Furthermore, any useful description of the channel must eventually relate to the possible distortion a packet may suffer in proceeding from antenna to antenna.

A review of previous material concerning propagation in the .5 and 1 GHz bands reveals that some of this information exists--particularly regarding the use of these bands in the land mobile service. Of the needed information, that concerning field strength as a function of distance for various terrain conditions is the best defined. The extensive model of Okumura et. al.[2] appears adequate for our purposes and is in substantial agreement with other investigations[3,4] both of which are quite recent. The Okumura model was derived from measurements and spans the frequencies of interest to packet radio. Because of the wide variety of possible terrain (e.g., the degree of urbanization) in which packet radio terminals may operate, we intend to compare some amplitude data against the Okumura model--if for nothing else than to associate different environments with the model.

As mentioned earlier we anticipate the need for characterizing the channel in terms relevant to vary wideband signals. To this extent available information is somewhat limited[5,6,7,8,9] and in some instances does not exist at all. All of the parameters we state below as needed for adequate channel characterization relate to the RF signal design.

We intend to treat the channel as a linear filter which, in the case of fixed terminals, is probably not sifnificantly time varying. Time invarance is presumed principally because the expected Doppler shifts are small relative to the allowable phase-variation over a typical detection period (approximately 10 μsec.). Our first attempts, however, will be directed toward determining whether or not motion of multipath reflectors is significant. This concern stems from the possibility that rapidly moving reflectors such as aircraft may be important. Even if they do not become important to packet radio, the effects of lower velocity reflectors such as automobiles will nevertheless be documented.

Another important effect imposed by the propagation medium is the limit on coherent bandwidth due to multipath. We shall attempt in this measurement program to determine not only the amount of excess delay due to multipath but also some of its shape characteristics that will be important to possible time-delay capture properties of wideband signals.

In considering the type of modulation to be used in the propagation measurement two choices become apparent: 1) a modulation with low ambiguity in both time delay <u>and</u> Doppler shift or 2) two separate modulations each having good ambiguity properties in one of the respective dimensions. We have chosen the latter for three reasons. First, the real-time on-site digitization of a wideband signal with low ambiguity in both dimensions requies rapid A/D conversion together with direct memory access into the computer--a somewhat elaborate and expensive hardware and software problem. Second, the information available from treating the two dimensions separately provides all the information necessary for communication system modeling--namely a measure of <u>spread</u> in time delay and in frequency. Third, we fully expect the channel Doppler characteristics to be of little consequence to the high data rates presently contemplated for packet radio.

We therefore intend to measure each of the following quantities for both frequencies and for various categories of terrain:

1. The distribution of maximum excess delays (i.e., beyond free space) for various terrain characteristics. This is needed to guide the selection of the maximum bit rate that avoids intersymbol interference. As indicated in Figure 1, this quantity will be given in multiples of the standard deviation above the mean amplitude. A second important excess delay measure is time-delay spread. This is defined as the second central moment of the impolse response and is perhaps the most useful one-number characterization of time delay for error probability estimation.

2. The shape of the multipath delay. This is a sample of the channel impulse response and in particular an attempt to quantify the resolvability of individual multipath components. Results will be given in terms of plots such as those in Figure 2. Resolvability as indicated in this manner will be explored at RF bandwidths of 20 and 40 MHz.

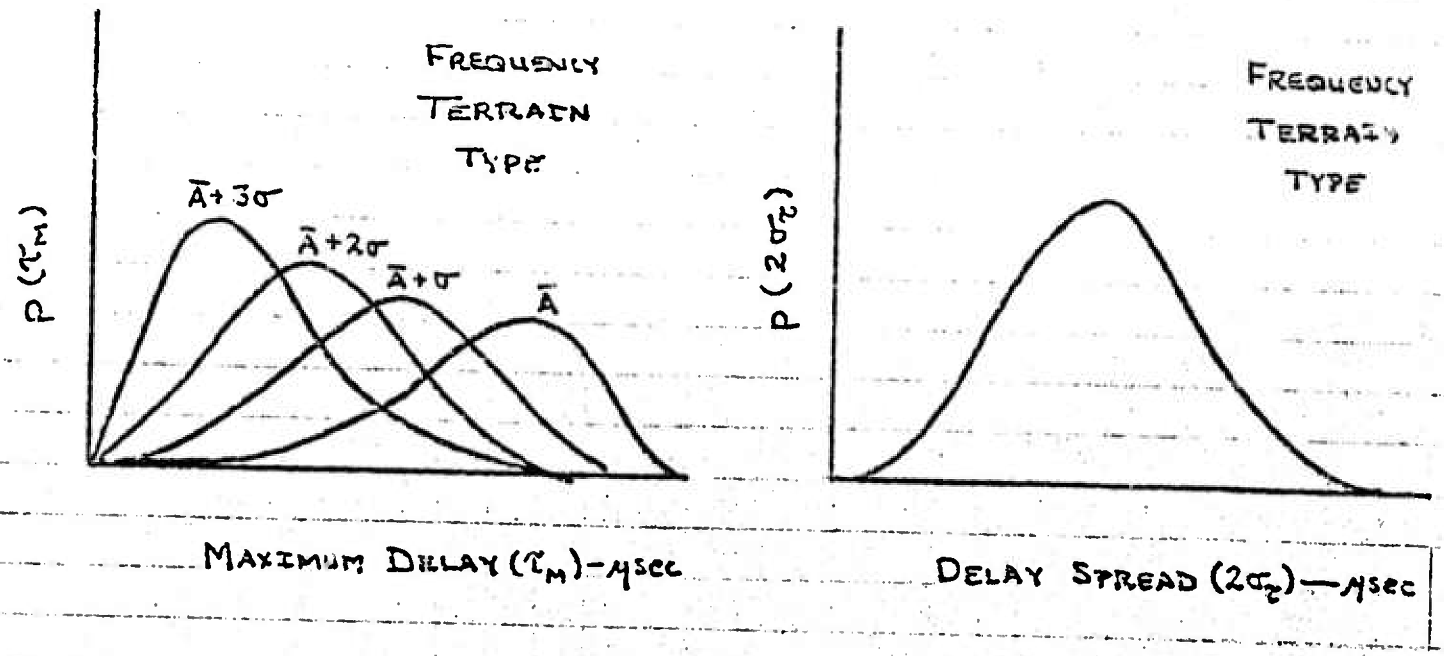

Fig. 1. Maximum Delay and Time-Delay Spread

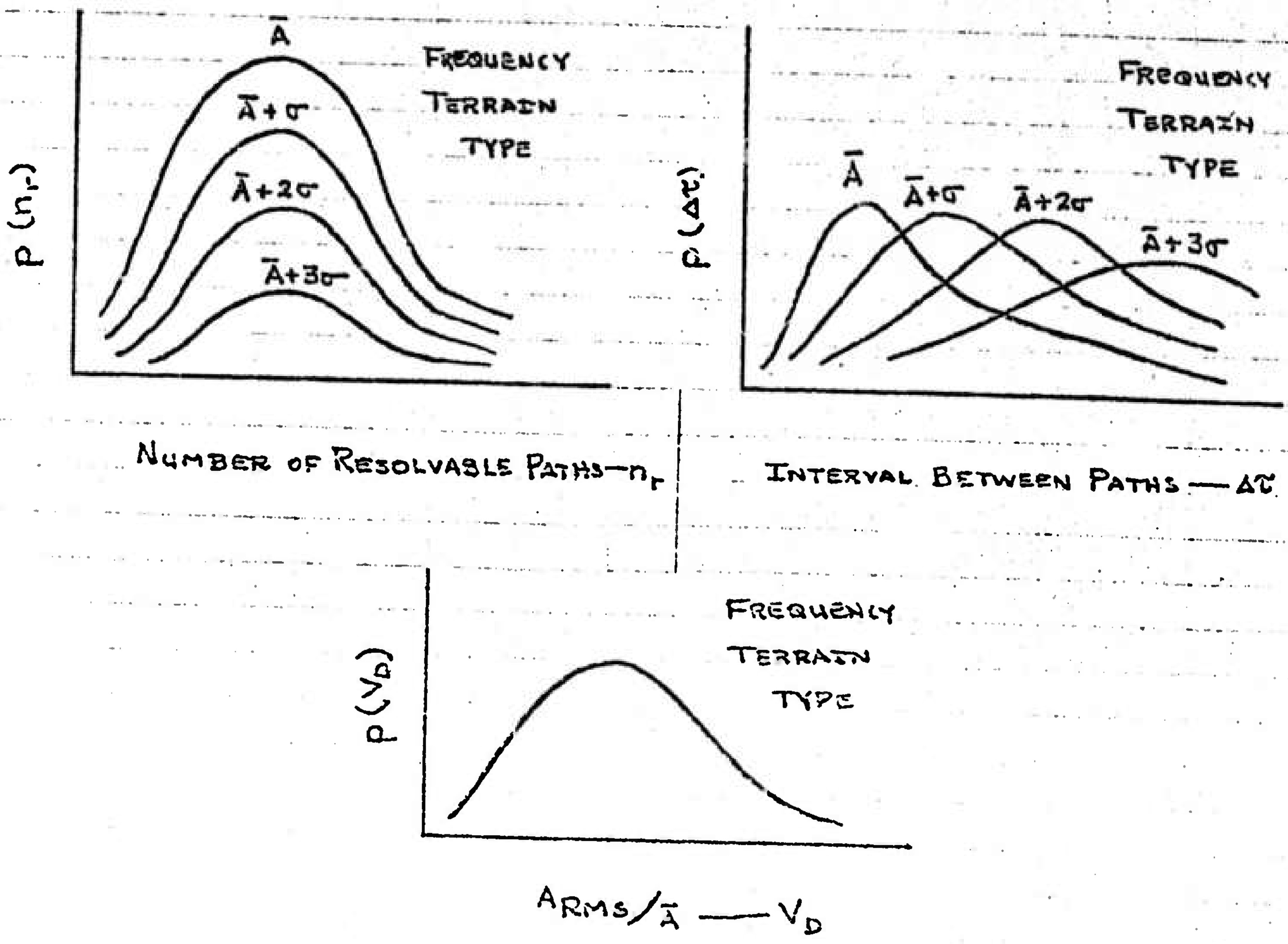

Fig. 2. Descriptors of the Shape of the Impulse Response

3. The time variability of the channel. This measurement will be approached from two standpoints: (a) the Doppler spectrum at frequency shifts less than ±100 Hz, the bandwidth within which most multipath targets are expected to lie, and (b) the Doppler spectrum at frequency shifts approaching those that would affect a 100 kbps DPSK system. Since no measurable effects are expected from (b) most emphasis will be placed on the narrowband case. The variability will be given in terms of power spectra similar to that shown in Figure 3 and by the spread of the spectra about its mean Doppler shift.

4. The spatial variability of the channel. This measurement, important to mobile terminals, is perhaps best defined in terms of systems that can or cannot resolve multipath. For those that cannot, a narrow band spatial frequency spectrum (see Figure 3) will be able to describe the Doppler spread associated with a given vehicle velocity. The measurement will be made by driving the mobile terminal at a constant velocity (governor controlled) while recording the output of the same narrowband quadrature detector used in the time variability measurement.

   For a system that does resolve multipath, the wideband signal will be used to track individual multipath components as the vehicle moves. This may be done by digitally recording the received pulse at specific increments in space.

5. Effect of Directive Antennas. Directive antennas are normally thought of as being able to reduce either required power levels or interference. In this instance we would propose to explore their use in reducing multipath spread. A receiver located in a highly relfective urban environment will probably receive multipath components from a wide range

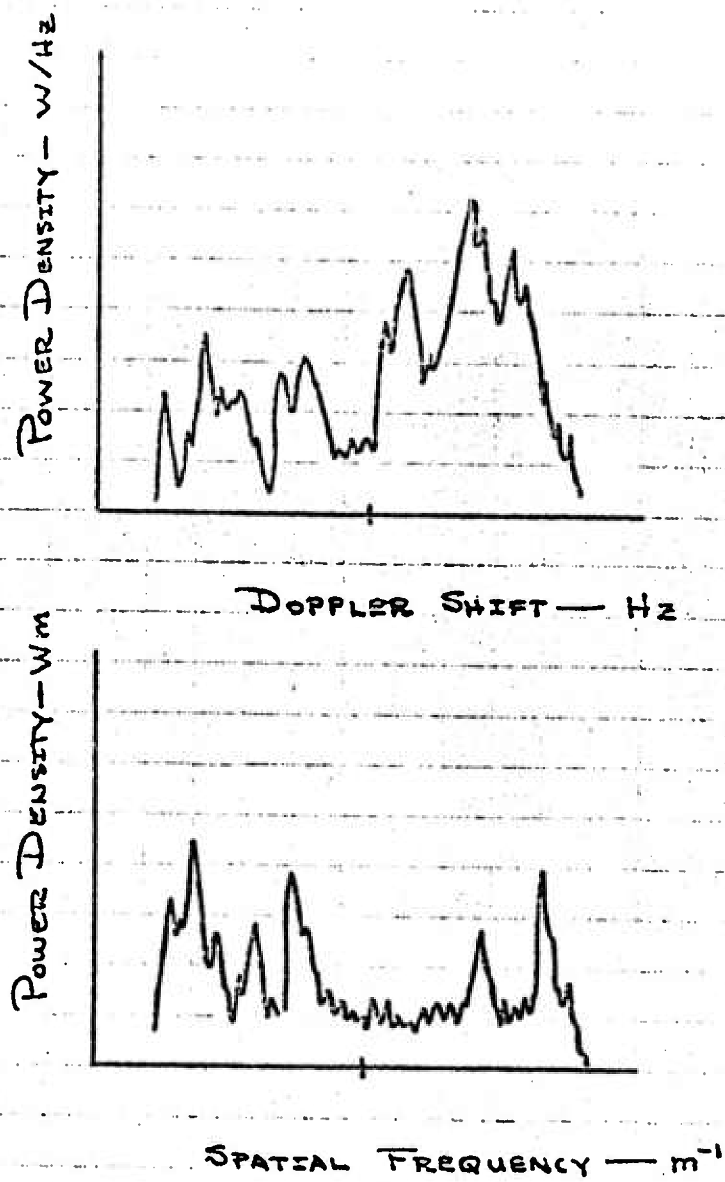

Fig. 3. Doppler and Spatial Frequency Spectra

of angles. To reduce the number of those components in azimuth, it will perhaps be adequate to judiciously place a null as opposed to trying to obtain directive gain. This approach would be exploratory since it is not clear that for terminal-to-relay communications (or the reverse) directive antennas are useful. Elevation patterns at the receiver can be altered somewhat by varying the antenna height above the top of the van (ground plane). By placing a null at low angles the noise received from cars can perhaps be lessened. This possibility will also be explored.

III MEASUREMENT LOCATIONS AND RANGES

A. Noise

Noise measurements in support of packet radio should be made not only at all the urban and suburban locations that are logical choices for the measurement of the propagation path characteristics but under some additional circumstances also. While gross multipath characteristics of a particular path would not be expected to change appreciably during the period of a day or a few hours, man-made noise is highly variable over these times. Within a city, for example, we have observed a considerable increase in the noise, identifiable with the passage of greater numbers of automobiles during the commuting periods. We expect the urban noise environment to differ greatly from day to night, requiring measurement at intervals around the clock to characterize the urban radio noise. Further since the noise at the frequencies considered for use by packet radio is generally man-made, we may expect to find a work-week dependence within urban areas that would not exist for naturally-occurring noise processes. Weather conditions are also an important variable for radio noise originating from high-voltage power transmission and distribution lines. Foggy or damp weather has been observed to raise the average noise power from power lines by more than 20 dB over that measured under dry conditions. To the extent that the aggregate urban noise is composed of that from power lines there will be a weather dependency in the urban noise environment.

So far we have suggested the measurement of noise only at those locations that are being considered for propagation measurements, but for additional times. Now we suggest some further noise measurement locations, selected to isolate particular noise sources rather than to accept the aggregate noise of the urban/suburban environment. To help relate these noise measurements to others, we plan to seek out and to measure the noise from a power line in an isolated area where there are no other noise sources and to similarly measure the noise near a freeway where the automobile ignition systems will be the only noise sources.

Table I will summarize the noise measurement situations that are being planned.

Table I

NOISE MEASUREMENT SITUATIONS

Location	Times	Other Conditions
Urban/Suburban	Intermittently for 24 hours	
Urban/Suburban	Intermittently for 4 weeks (including week-ends)	
Urban/Suburban		Wet and Dry Weather
Isolated Freeway		Heavy and Light Traffic

B. Propagation

The locations at which propagation measurements are necessary are those which are categorical. That is, they constitute a differentiable set that would apply to a wide range of packet radio operating environments. At this time we anticipate that packet radio terminals and repeaters will operate mostly in urban/suburban environments, which in many respects may constitute the most challenging operational environment. We will therefore concentrate on that type of "terrain" using various points in the San Francisco Bay Area.

Since determining signal strength contours about a given transmitting antenna will not be our primary emphasis (observed strength will be checked against existing models, however), locations will be chosen first on the basis of expected multipath severity and the likelihood of direct line-of-sight propagation. An example of one categorization is given in Table II where for completeness a rural class is also included. Possible locations in the Bay Area satisfying these categories are indicated in Table II and on Figure 4.

Propagation ranges or areas of coverage used in this measurement program are intended to be of the order of those contemplated for packet radio. The coverage area, as far as equipment is concerned, is defined by the transmitter energy per bit, the relative antenna height and the local noise level. We will be using power levels, modulation types, and antennas similar to those proposed for the early packet radio systems. Depending upon the type of terrain and relative antenna heights, coverage radii of between 3 and 50 km are anticipated.

Table II

CATEGORIES FOR PROPAGATION MEASUREMENTS

Category	Type of Propagation		Map Location
	LOS	In Defilade	
Dense Urban			
More than 20-story buildings		X	A
Moderate Urban			
3 to 20-story buildings			
Hilly	X	X	B
Flat	X	X	C
Moderate Urban			
Geographically limited	X		D,E
Suburban			
Hilly		X	F
Flat	X		G
Rural			
Hilly		X	H
Flat	X		I

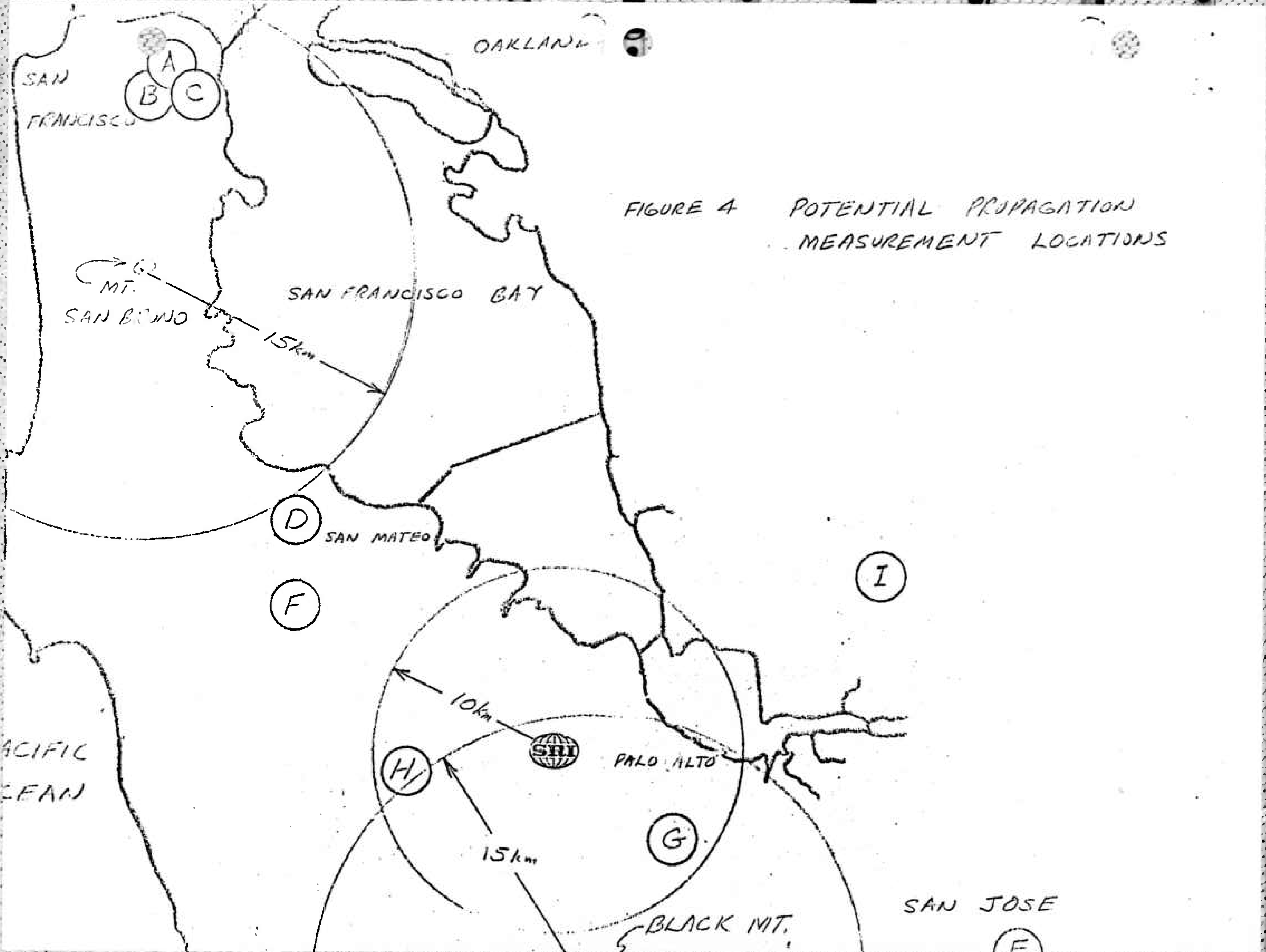

FIGURE 4 POTENTIAL PROPAGATION MEASUREMENT LOCATIONS

IV INSTRUMENTATION AND PROCESSING

A. Radio Frequency Instrumentation

1. Transmitter

The transmitter must be of sufficient power and the antenna of sufficient height to provide the coverage needed over the distances characteristic of the test area. These distances, a few of which are shown on the map of Figure 4, are typical of those expected in packet radio. Using the results of Okumura,[1] the effective radiated power (ERP) required for a given range may be determined. Assuming a 200 kHz RF bandwidth for noise, a 20 dB signal-to-noise ratio, and a 1.5 m receiver antenna height, Table III gives the expected range based upon median signal conditions. Various noise levels and transmitting antenna heights are considered.

Table III

TRANSMITTER POWER REQUIREMENTS

Frequency	Location	Trans. Ant. Hght.	Effective Input Noise	Radius	ERP
430 MHz	Urban	100 m	20 dB > kT_oB	2 km	16 dBw
	Urban	1000	20	.9	16
	Suburban	30	10	.4	16
	Suburban	1000	10	.34	16
1325 MHz	Urban	100	10	2	15
	Urban	1000	10	.9	15
	Suburban	30	5	.3	15
	Suburban	1000	5	.25	15

More complete data are shown in the Appendix along with the conversion of Okumura's curves to fit the particulars of this program.

The block diagram of the transmitter is shown in Figure 5. The transmitter will be switchable between CW and pseudo-random-phase switching operation. All oscillators will have a stability of at least a part in 10^{10} such that a resolution of .5 Hz may be obtained at 1.5 GHz.

Operation of the transmitter will be manual in the sense that switching between various frequencies and chip rates or between on and off will be by hand. The transmitter equipment will be easily portable since it will be sited temporarily at several different locations.

Transmitting antennas will be omnidirectional in azimuth. Vertical beamwidth will be narrowed below that of a vertical dipole for the 1325 MHz frequency because of lower available output power.

2. Receiver

As with the transmitter, the receiver will consist of separate RF sections for the two frequencies (see Figure 6). Manual switching will occur at the 140 MHz IF in order to provide for the various noise and propagation measurements. The dynamic range at this point should be greater than 50 dB. All oscillators must be stable to at least one part in 10^{10}.

a. Noise

Noise will be measured in bandwidths of 40 MHz and 500 kHz as well as at the output of the surface wave devices. Noise will be processed either through the Threshold Detector or digitally converted by switching to what is normally the pulse output.

b. Spread Spectrum Signal

Two 127-chip surface wave devices for 20 and 10 megachip rates will be used and are switch-selectable. The output of each is passed through an envelope detector and output to an A/D converter capable of digitizing at a rate of 10^8 samples per second.

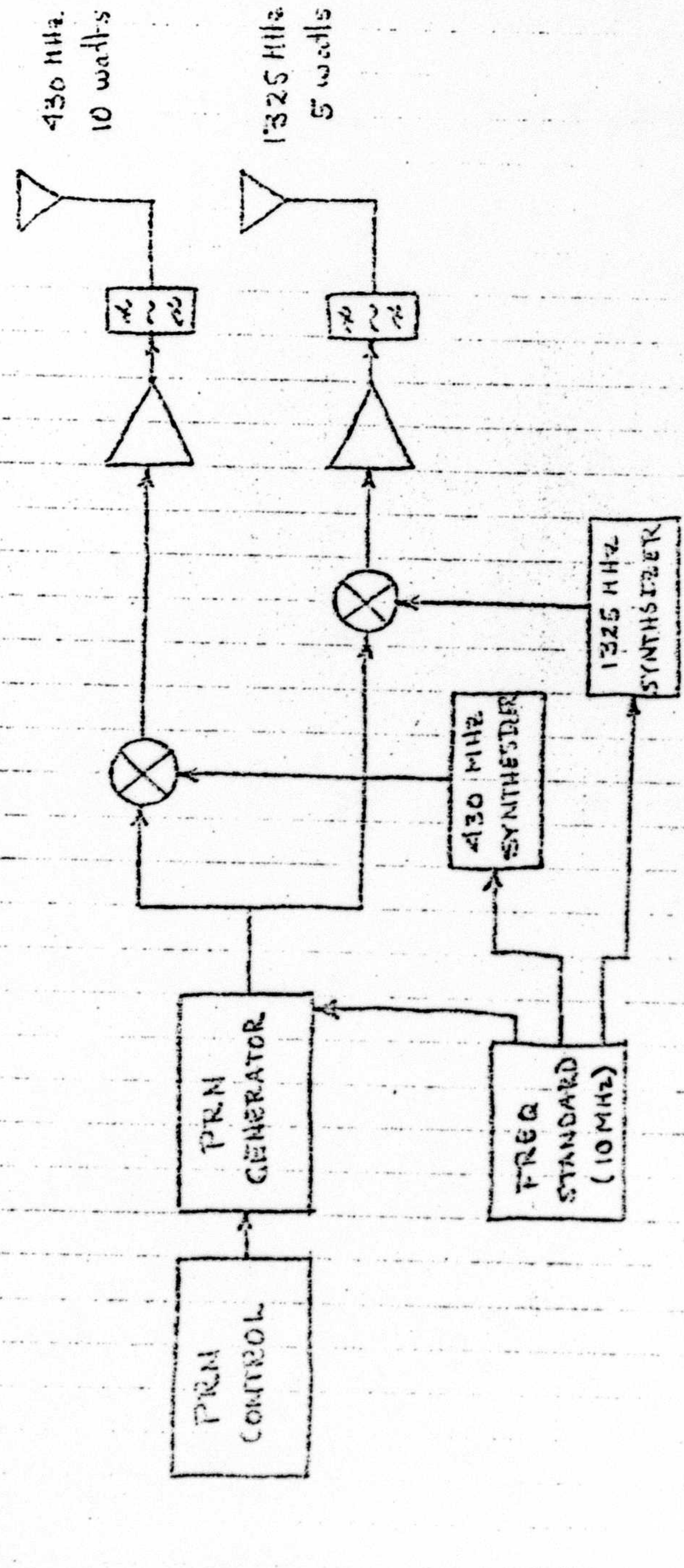

Fig. 5 Transmitter System

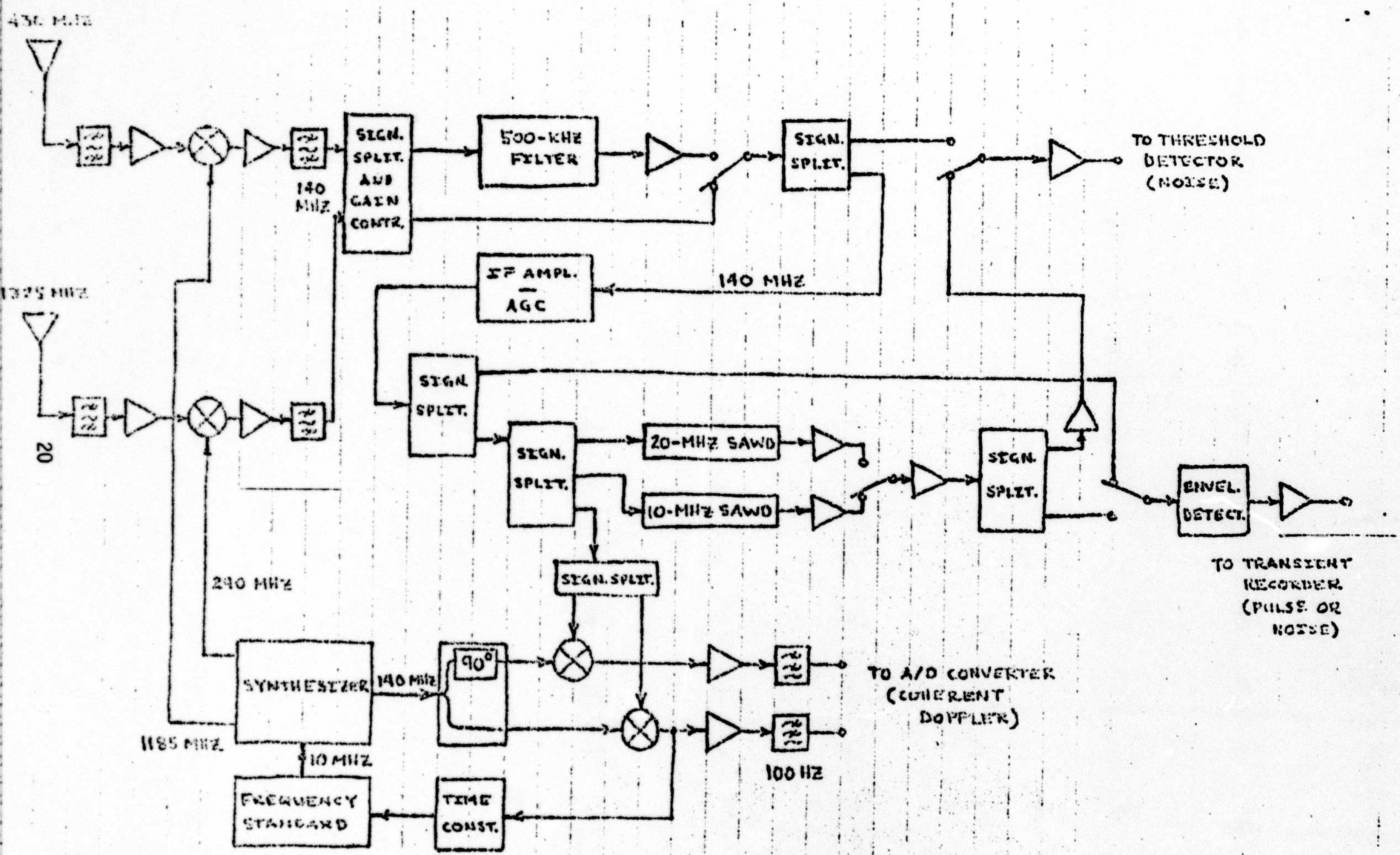

Fig. 6. Receiving System

c. Narrowband Signals

To determine the extent of channel variation a quadrature detector with two sets of post detection filters are provided. A nominal 100 Hz bandwidth will contain the spectral region where most variation is expected.

d. Receiving Antennas

Receiving antennas may be directive in two ways. For multipath reduction it may be adequate to manipulate a single null such as that formed by a cardiod. (Because of the small surface available on the mobile van, a large aperture antenna is not feasible. If signals propagate predominantly along an urban street, then a null may help reduce the signals from one direction.

A second use of directivity is to minimize the low-angle, nearby sources of noise or interference. This effect requires a vertical lobe that "cuts-in" at low-arrival angles.

Both of these antenna patterns will be used from time to time during the course of the measurements. Their use will probably be restricted to that sufficient for a comparative analysis with more conventional dipolar antennas.

3. Voice Channel

Since the experiment is not intended to function automatically, a duplex voice channel will coordinate the activities at transmitter and receiver.

B. Post Detection Processing and Recording

1. General

The processing and recording system proposed in this measurement program is centered around a minicomputer. A block diagram of the processing and recording system may be found on Figure 7. The minicomputer is to be shared among the particular noise or propagation

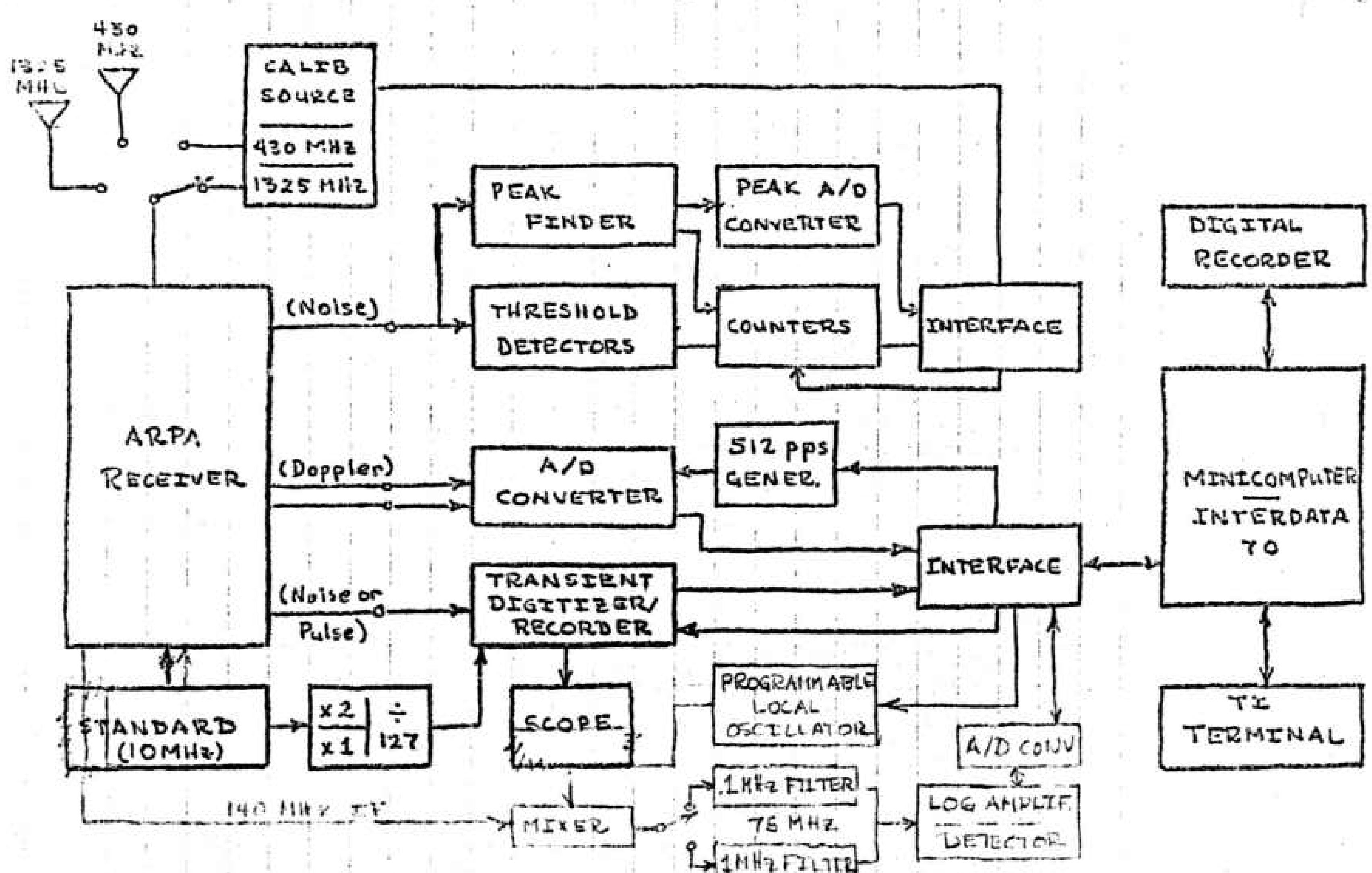

Fig. 7. Signal Processing and Recording System

measurements. Setting up for a given measurement will entail the manually switching in of the correct complement of hardware and, through the computer terminal, loading of the correct software. One of the several specific detector outputs shown in Figure 6, in combination with the appropriate peripheral devices, constitutes the processing and recording system for a given measurement. The various arrangements will be described in the following two sections.

2. Noise

a. General

Noise measurements will be made to identify the limited number of noise parameters mentioned in Section II. We are interested in three types of noise--each defined by the part of the receiver through which they pass:

(1) 500 kHz (rf) bandwidth noise

(2) 40 MHz (rf) bandwidth noise

(3) SAWD noise output.

Each measurement will either contribute to the characterization of the input noise or to the noise at the point of decision making circuitry. Each of the above types of noise will be analyzed using two techniques. The first is the use of a group of threshold detectors whose outputs are eventually digital counters. These outputs are processed in the minicomputer and output to the terminal. The second technique is to feed samples of the noise directly into a transient digitzer/recorder and thence to the mini and a digital recorder. We will use the digitally recorded samples to calculate amplitude probability densities and measures of the impulsiveness such as V_D, the ratio of rms to average noise amplitude.

b. The Threshold Detector

The outputs of six Threshold Detectors will be used to determine the following:

(1) Joint amplitude-width probability density, $p(A,W)$

(2) Probability density on pulse interval (irrespective of pulse amplitude), $p(T)$

(3) Probability density on peak amplitude, $p(A_p)$.

To obtain (1) we must determine the width of noise pulses at each amplitude threshold. To do this we will use the binary output of six comparators (each associated with a threshold) to start and stop the counting of a frequency of about 200 MHz. At a particular positive-going threshold crossing the counter associated with that threshold will be started. The counter will be stopped at the next negative-going crossing of the threshold. The counter will be dumped (the start circuit will be immobilized until dumping is completed) and the counter then enabled for the next positive-going crossing. All such pulses counted will not be contiguous. The widths (counts) of all pulses counted at each level will form the distribution for that level for that time/location. Sufficient counts will be taken to accurately define the distribution--approximately 1000 in number. Nonparametric estimates suggest that 1000 samples would define the cumulative distribution function within 4 percent with .95 confidence.

To obtain the pulse-to-pulse interval distribution one could simply measure the interval between pulses for each threshold level--the complement of the width measurement. A more general result, however, can come from first locating when each noise peak occurs. (Such a peak locator can be developed using all of the six Threshold Detector outputs.) Knowing this the interval between two consecutive peaks can be measured. A distribution of time interval between noise spikes can then easily be formed.

The peak locator can also be used to help obtain the third required distribution. A knowledge of when the peak occurs permits sampling of the peak value provided the signal to be sampled is appropriately delayed. A sample-and-hold circuit that can follow the wide 20-MHz post detection bandwidth plus a narrow aperture A/D sampler will be used to obtain sample peak-values.

c. A/D Conversion of Noise

Several noise measures are more simply obtained through digitization rather than the Threshold Detector. The Transient Recorder will be used to sample values of the noise output of either the 500 kHz and 40 MHz filters or the SAWDs. Sampling must occur at intervals substantially greater than the reciprocal of the bandwidth. The Transient Recorder will be pulsed by the mini to sample at the appropriate interval and the data will be calibrated and blocked before recording on tape.

Since the sampling interval is quite arbitrary we may consider performing most of the processing in real time and outputting only the results for later off-line plotting. Those parameters to be calculated are:

(1) Amplitude Probability Distribution

(2) RMS Amplitude

(3) Average Amplitude

(4) Ratio of RMS to Average Amplitude.

2. Propagation

a. General

As discussed earlier the main part of the propagation measurement consists of trying to measure and characterize (1) the impulse response of the packet radio channel (2) the extent of time variability, and (3) the spatial variability. The first of these measurements is wideband whereas the latter two are narrowband measurements. The two narrowband measurements are computationally nearly identical. One narrowband recording will be made while the receiver is stationary and one while the receiver is moving at a constant velocity. All propagation measurements are to be digitized and placed, after formatting in the minicomputer, onto magnetic tape. Signal processing therefore is done entirely off-line.

b. Digital Recording of Propatation Measurements

The Transient Recorder shown in Figure 7 will digitize and store 2800 8-bit samples at a maximum rate of one each 10 ns. Readout from the Recorder's memory is at a slower rate--one compatible with the computer/recorder electronics. At a given frequency, chip rate, and location approximately 200 pulses will be recorded on magnetic tape. The pulses will be processed for the impulse response characteristics outlined in Section II. In cases of low signal-to-noise ratio several pulse envelopes may be added together. Just how this is done depends upon the pulse-to-pulse variation of the received signal. Good time stability will be required to do this so triggers will be derived from a stable oscillator.

For fixed terminals the channel time variability in an urban/suburban environment is defined almost exclusively by automobiles. If so the maximum Doppler shift should not exceed say 250 Hz at 1525 MHz. The power involved on the extremities of the spectral spread will of course depend upon the target cross section afforded by a car as compared to the power reaching the receiver by other means. The use of an elevated transmitter would seem to minimize the effect of reflection from automobiles even when obstruction loss is considerable. An initial post-detection bandwidth will be 100 Hz although this may later have to be increased. A two second recording will be made at a 250 Hz rate on each of the quadrature channels.

The recording of the spatial variation proceeds in an almost identical manner to the time variability. The receiver in the van will proceed at a constant speed (±5%) for the duration of a recording period (2 sec). The record will be constituted exactly the same as the time variability record and the velocity of the van will be used to obtain a compromise between sufficient distance (wavelengths) and aliasing.

Velocities parallel to the wave normal of 30 and 15 mph will result in the following spatial record lengths:

Frequency	430 MHz	1325 MHz
Wavelength	.7 m	.23 m
Velocity	30 mph	15 mph
Duration	2 sec	2 sec
Distance	38.4 λ	59.2 λ
Maximum Doppler	±20 Hz	±30 Hz

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Appendix

USE OF OKUMURA MODEL[2] TO DETERMINE RANGE OF MEASUREMENT SYSTEM

A 127-chip sequence takes 6.35 μsec when switching at a 20 MHz rate. This results in a 157 kilobit transmission rate or about a 314 $\overset{K}{\text{M}}$Hz of bandwidth. Let P_M be the external noise power in a 314 kHz RF bandwidth in dB above kT_oB. The same power, then in dBm is

$$P = -174 + 55 + P_M \quad \text{dBm} \tag{A1}$$

For a 3 dB receiver noise figure,* a 20 dB SNR, and a $\overset{16}{3}$ dB failure to recover the time-bandwidth factor of 127, the signal power required is

$$P_s = P + \overset{36}{23} \quad \text{dBm} \tag{A2}$$

For a transmitter power and a gain of P_T and G_T and using Okumura's loss model, the expected range can be found from

$$P_R = P_s - P_T - G_T \quad \text{dBm} \tag{A3}$$

where

$$P_R = E_{OK} + G_R + 20 \log \lambda - 155.4 \text{ dBm per watt ERP} \tag{A4}$$

P_T and G_T are in dBW and dBi, respectively

G_R is the receiver antenna gain (dBi)

λ is the wavelength (m)

E_{OK} is the median field strength required $\left(\frac{\mu V}{m}\right)$

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* Since a 3 dB noise figure contributes less than 1 dB to the effective input noise for input noise power greater than 6 dB above kT_oB, we will disregard the noise contribution of the receiver.

Combining Equations (A1) through (A4),

$$E_{OK} = (P + 23) - P_T - G_T - 20 \log \lambda + 155.4$$

$$= P_M - P_T - G_T - G_R - 20 \log \lambda + 59.4 \text{ dB above } 1 \frac{\mu V}{m}$$

At 430 MHz, P_T = 10 dBW and $G_T = G_R = 0$, so

$$E_{OK} = P_M - 10 + 3.5 + 59.4 = P_M + 53 \text{ dB above } 1 \frac{\mu V}{m}$$

At 1325 MHz, P_T = 7 dBW, G_T = 8 dB, and $G_R = 0$, so

$$E_{OK} = P_M - 7 - 8 + 13 + 59.4 = P_M + 57.4 \text{ dB above } 1 \frac{\mu V}{m}$$

These two equations for E_{OK} were used to convert the ordinate scales of Okumura's Figures 41(b) and (d). The results appear in Figures A-1 and A-2.

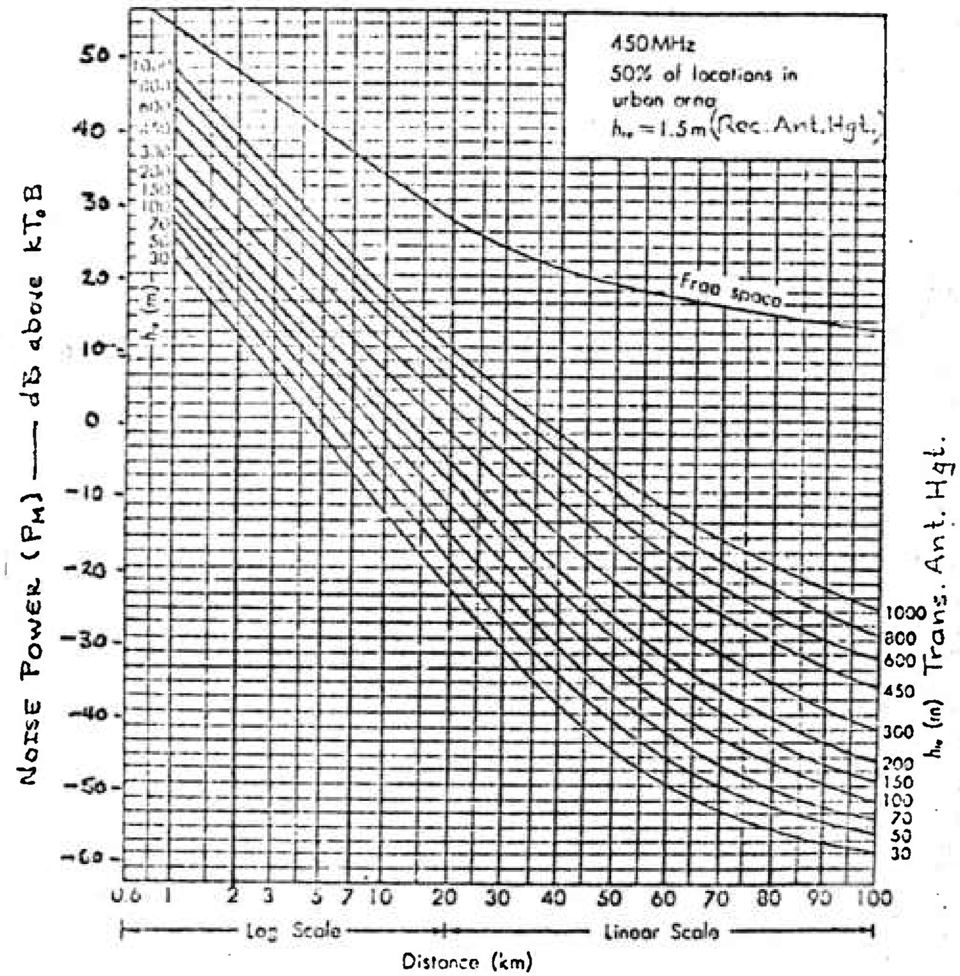

Fig. A1. Range Available for Various Input Noise Powers at 450 MHz

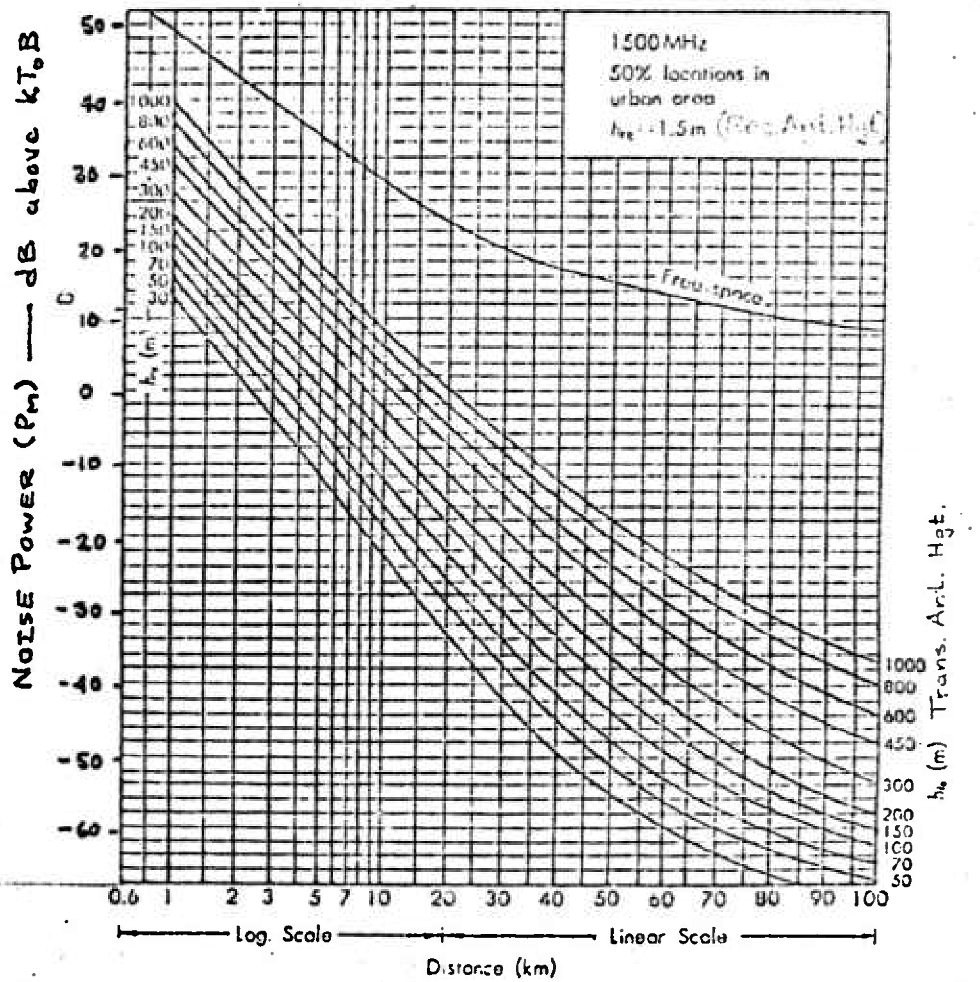

Fig. A2. Range Available for Various Input Noise Powers at 1500 MHz

For the type of modulation planned, satisfactory performance (probability of bit error $\leq 10^{-5}$) will be obtained so long as the received bit energy-to-noise density ratio exceeds 20 (13 dB). Received bit energy depends on propagation loss, and noise density depends on background noise level as outlined below.

The received power is determined from the equation:

$$P_r = P_t G_t G_r L_p$$

where

P_t = transmitted power

G_t = transmitter antenna gain

G_r = receiver antenna gain

L_p = propagation loss

The received energy per bit is

$$E_b = P_r T_b$$

where

T_b = duration of one bit.

The noise power density is

$$N = N_o(F_m + F_r)$$

where

F_r = receiving system noise figure

F_m = non-thermal noise factor

N_o = thermal noise density

Thus, the received energy per bit-to-noise density ratio is:

$$E_b/N = \frac{P_t G_t G_r L_p T_b}{N_o(F_m + F_r)}$$

For satisfactory performance $E_b/N > 20$. This requires that

$$\frac{L_p}{F_m + F_r} > \frac{20\ N_o}{P_t G_t G_r T_b}$$

The propagation loss, L_p, can be related to (Okumura's) curves, since the curves plot field strength received for ½ kW transmitted. The relation is derived in the previous section, so that

$$10 \log (L_p) = E_{ok} - 155.4 + 20 \log \lambda \quad \text{dB}$$

The four situations depicted in figures 6-2 and 6-3 were derived using the following link parameters

P_t = 10 W

$N_o = 10^{-20.3}$ W/Hz

$F_r = 2$

$\lambda = 0.2$ m

For the terminal-repeater link

$G_t = 9$ dBi

$G_r = 0$ dBi

$T_B = 10\ \mu s$

and receiver antenna height $h_r = 1.5$ m.

For the repeater-repeater link

$G_t = 9$ dBi

$G_r = 9$ dBi

$T_B = 2\ \mu s$

and $h_r = 200$ m.

Okumura's urban/suburban correction factor and antenna height correction factor have both been used to obtain suburban curves and the curves for $h_r = 200$ m.

Appendix B. 3

Dynamically Allocated Multiple-Channel Network Concept

I INTRODUCTION

The acronym DAMN (-FINE)* has been chosen to represent a Dynamically Allocated Multiple-channel Network (For Improved Network Efficiency). It might equally have been chosen as the DYNATAC-ALOHA Multiple-channel Network, since many of the concepts are drawn from these two multiuser networks. The DYNATAC network is a mobile-radio voice network proposed recently by Motorola on FCC Docket #18262. It consists of an array of fixed, independent transmitting and receiving sites interconnected by telephone cable in a networ with a centralized control, and a set of hand-held voice terminals which can be synthesizer-tuned under remote digital control to any of many narrow-band frequency channels. The DYNATAC centralized control automatically allocates channels to a terminal-transmitter site-receiver site triplet. These channels are fixed for the duration of a communications transaction although the transmitter/receiver may change. Thus the channel is not available to any other user in the same area, although it may be reused over and over on a noninterfering basis at other locations. The DAMN system concept draws upon the DYNATAC concept to provide many geographically reusable narrow-band channels; however, it discards the disadvantage of a hard-wired, centrally controlled network, using circuit switched channels. To avoid the inefficiency of circuit switching for digital data, the DAMN system draws upon the University of Hawaii ALOHA system as extended by the packet radio project. In the DAMN system each channel is time-shared among several mobile users and a repeater. The

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* Hereafter referred to as the DAMN system. The reader may introduce the parenthetical (FINE) if he is so inclined.

hardwired network of transmitter-receivers is replaced by a network of portable (but fixed) repeaters which are interconnected by rf links, chosen from the same channel set. The repeater channels are also time-shared a-la ALOHA (with a carrier-sense modification). The centralized control is replaced by an adaptive channel search and selection algorithm in each repeater and terminal which effectively distributes the control, but provides for very efficient dynamic allocation to match changing traffic and interference patterns. Because the DAMN system searches passively for channels, and because it is a frequency channelized system it can be readily overlaid on existing systems with minimal mutual mutual interference. It requires no active cooperation in allocation of channels. The DAMN system uses only the resources it needs, and carefully selects the resources from those which are not used by existing systems.

The purpose of this note is to suggest a design concept for the DAMN system which embodies much of the knowledge and research reported in the Packet Radio Notes, and which seems to meet all of the system design criteria so far voiced by the Packet Radio Working Group.

Briefly, the design criteria considered are:

1. The system should provide error-free communication of digital data between a mobile terminal and a station with minimum network delay.
2. The system should use the spectrum efficiently.
3. The system should be compatible with existing users.

4. The system should not stress technology unnecessarily.

5. The network should be readily expandable as users are added.

The degree to which the DAMN system meets each of these criteria will be discussed in a subsequent section after the system concept is described in some detail.

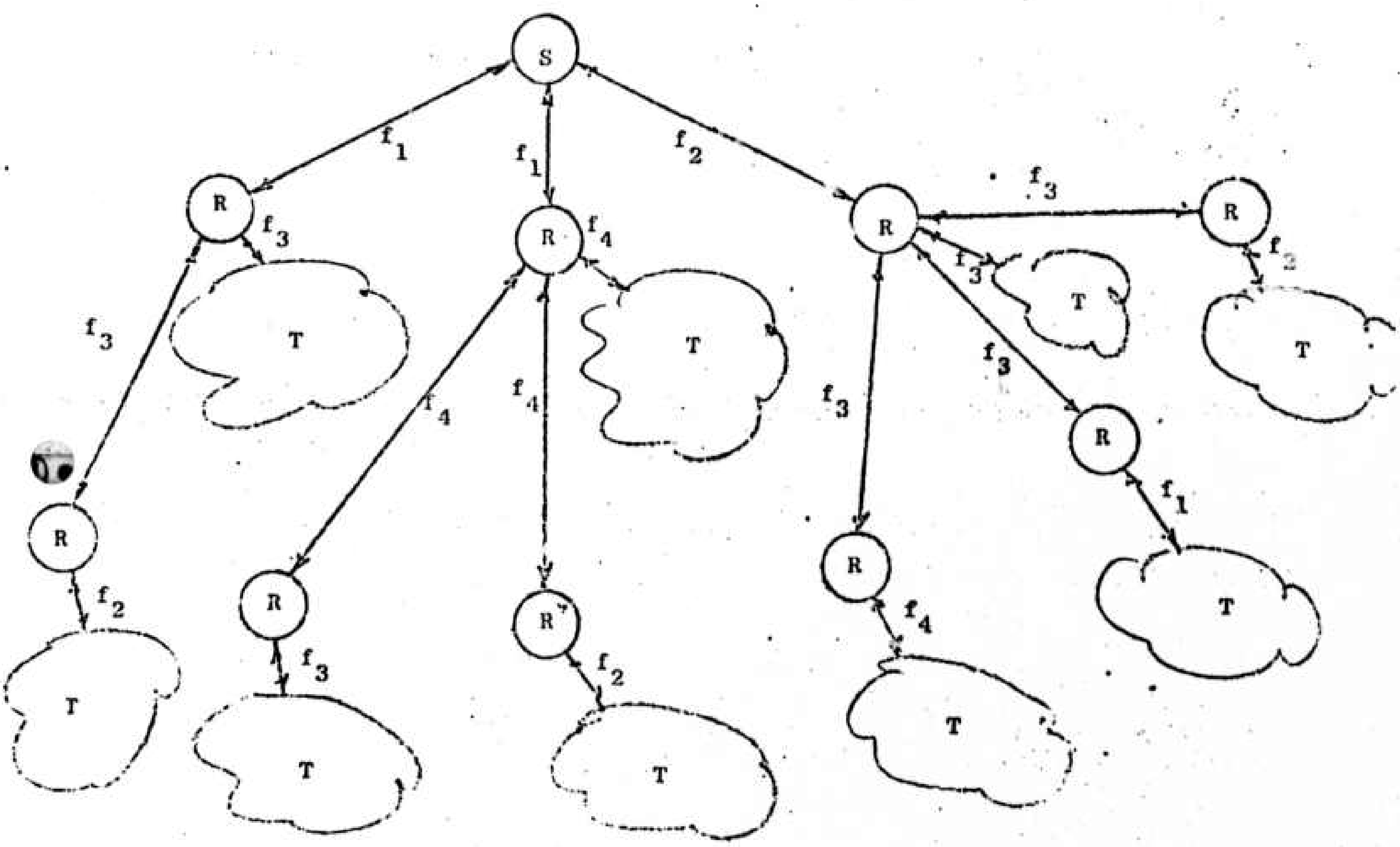
The suggested system includes many speculative, or undefined concepts in the area of network organization, routing, flow control, hardware design, etc. This has been necessary in order to quickly bring the concept before the Packet Radio Working Group. It is hoped that other members of the group will criticize, modify, add to and otherwise improve the idea if it has merit.

II DESCRIPTION OF THE DAMN SYSTEM

A. System Concept

Figure 1 represents the suggested DAMN system concept. It is a network with a station, a hierarchy of repeaters, and terminals which are repeater-associated. The station has a multichannel capability which is dynamically allocated to best fit the traffic patterns. The channels are narrow-band (100 kHz) Minimum-Shift-Keyed channels. The dynamic allocation is not centrally controlled, and requires no complicated control algorithms. Each terminal and repeater automatically select a channel which is least busy. Only when there are not enough repeaters in a given area to handle the area traffic will overload occur. Furthermore, if a high-duty cycle user enters the net he will cause other lower duty cycle users to avoid or leave the channel he selects, so that he will eventually obtain a channel well matched to his requirements.

As suggested in Figure 1, each channel is centered on a frequency which can be used in several noninterfering areas simultaneously. The DAMN system uses many relatively narrow-band frequency channels and the DAMN station has the capability of operating on several of these simultaneously. The repeaters can operate on any two frequencies and the terminals on any one. The frequencies may be selected at random anywhere within the allocated band. Figure 2 shows a hypothetical situation in which a heavily used mobile radio band has been selected in which to set up the DAMN system. Because of the nature of the distributed dynamic control, the DAMN system will automatically select



S = Station
R = Repeater
T = Terminal

$\xleftrightarrow{f_1}$ ≡ channel at frequency f_1

Figure 1 Typical DAMN System

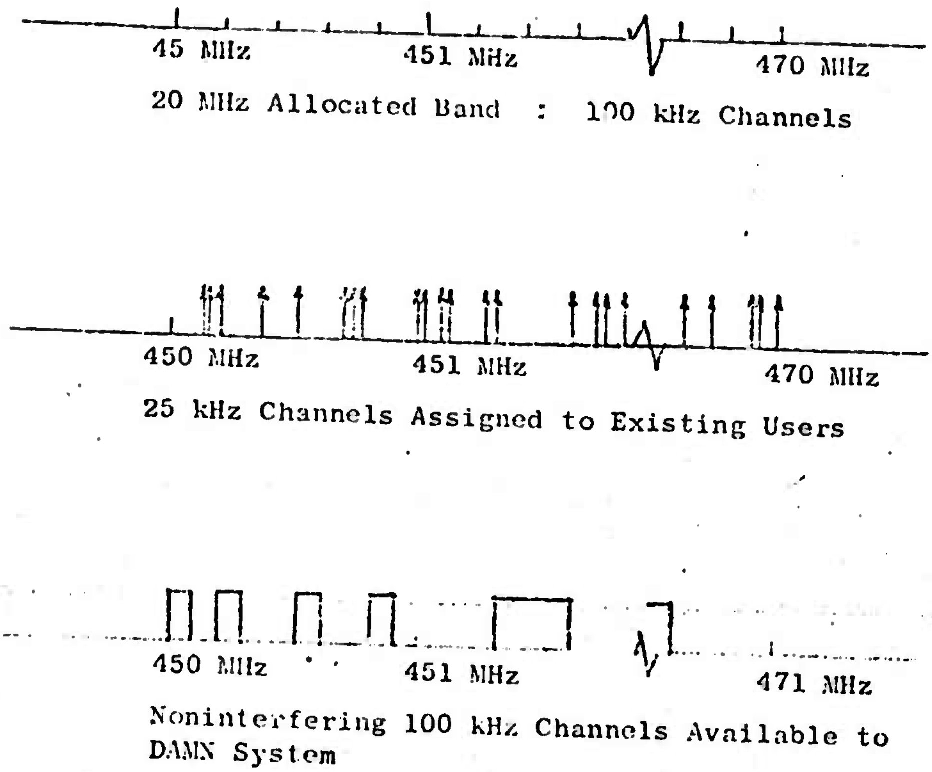


Figure 2 A Hypothetical Distribution of Channels

noninterfering frequencies. If an interfering user comes up on a DAMN frequency, the interferer will cause a reallocation so that the new frequencies will again be noninterfering.

Because the interference pattern will change from area to area and because the DAMN system can adapt to different interference patterns and reuse frequency channels in different geographical areas, it is possible that a DAMN system could operate effectively in an area where the frequency band would be considered completely allocated, or in a frequency band where the number of narrow-band signals would make operation of a spread-spectrum system impractical.

To understand the nature of the suggested distributed control suppose that a user at a mobile terminal wishes to initiate a communications-transaction with the station; that is, he wants to send a message and receive a response. Let us follow the signaling sequence through the network. The terminal is designed to periodically scan all possible frequency channels and to determine which channels are being used by nearby repeaters. The terminal also estimates activity in each nearby channel and selects that one which is least loaded. In this way the load is dynamically allocated to achieve an even distribution over the channels, but no centralized channel allocation control is needed. A possible channel-search algorithm is illustrated in Figure 3 to show the simplicity of this control method. Thus, when the user inputs his message, the terminal has already decided which channel to use. The message is transmitted and response received on this channel using a Carrier-sense access

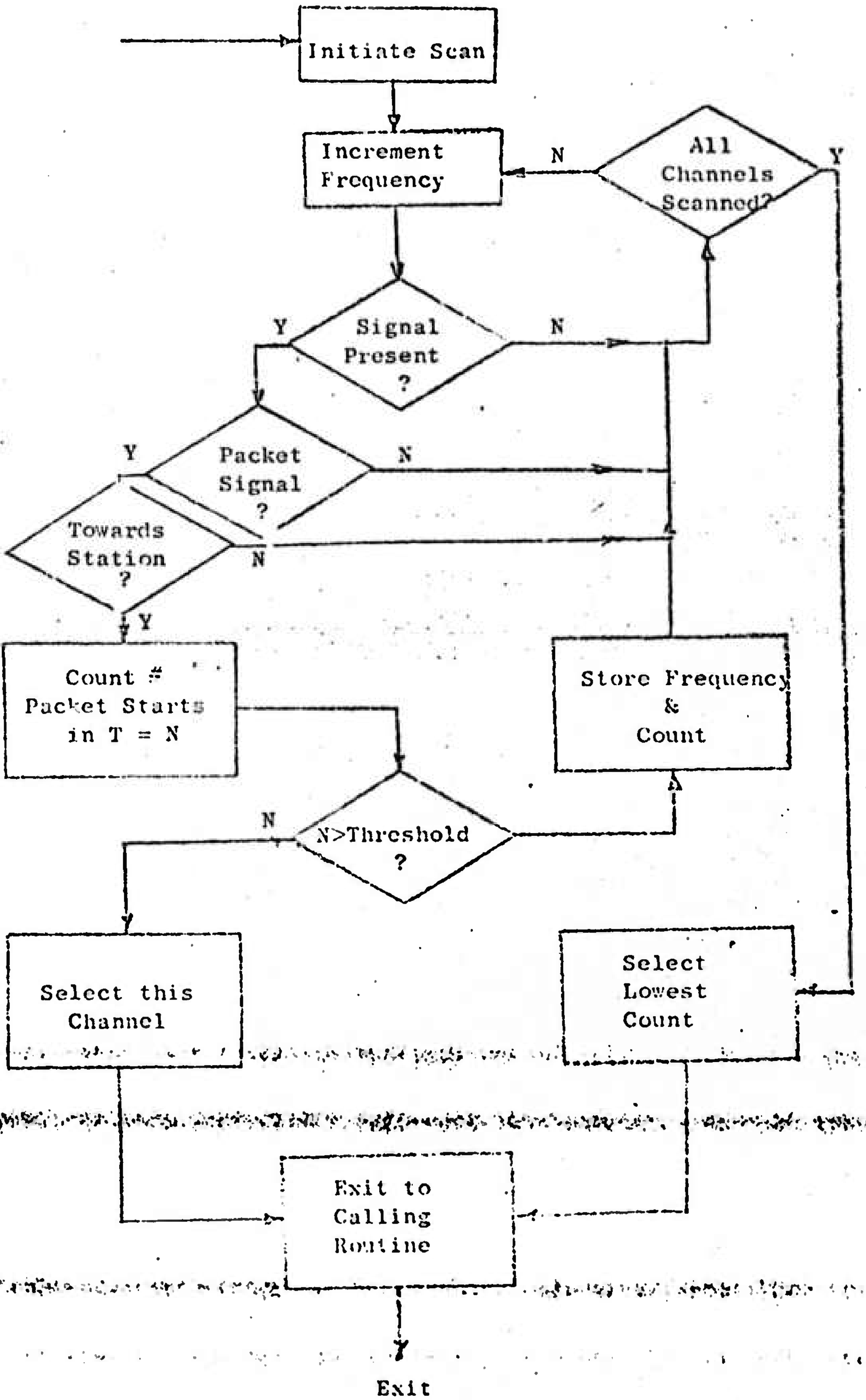


Figure 3 A Terminal Channel-Selection Algorithm Flow Chart

mode. The terminal will continue to use its first-chosen channel until the end-to-end delay becomes excessive. If this occurs it will initiate a scan to find another channel which may provide less delay either because the new repeater is less busy, or because the new route through the network is less busy.

Meanwhile the repeater which will receive the message has selected two channels for operation. One (the outbound channel) will carry traffic to and from repeaters farther from the station and to and from local-area terminals (this is the channel which our terminal will select); and the second (the inbound channel) will carry traffic to and from the repeaters next on the route to the station. The inbound channel is selected using the same criteria as used by the terminal. The repeater initially scans all channels to find the least-busy inbound repeater (or station) that can be reliably heard, and chooses this channel. The repeater will continue to use this channel until retransmissions become excessive at which time the repeater initiates a search for a new inbound channel. Retransmission may become excessive either because the inbound repeater channel is loaded, or because the inbound route is loaded. In either case, the repeater will automatically initiate a channel search, and the load will be reallocated, if a better channel is available, without the necessity of a centralized control.

The outbound channel is selected by the repeater initially by scanning the channels to find an empty one. This channel is monitored for a short period of time to assure that it is not being used by a nonnetwork member (such as a radar or air-to-ground voice communication system) with a low duty-cycle. Then the repeater begins to periodically broadcast "beacon" packets to tell terminals and other repeaters that this channel is available to be used for inbound traffic. As the inbound user load builds up a corresponding outbound load will also build up so that the dummy beacon packets need not be sent.

When the repeater receives the message sent by our user terminal, it will acknowledge receipt on the outbound channel and repeat the message on the inbound channel using a Carrier Sense mode.

The station must establish a link with each nearby repeater. A number of strategies are possible. For example the station might scan all channels, find an empty one (just as the repeater finds an outbound channel), and transmit periodic "beacon" packets until traffic builds up to the point where beacon packets are no longer needed, and finally, until the traffic approaches channel capacity. At this point a second channel is established by sending beacon packets on another empty channel. This requires that the station be able to operate on several channels simultaneously; however, any multichannel system must have the same capability. Initially a single repeater can be used as the rf part of a station, providing dual-channel capability. As the network grows so that more channels are needed, additional repeaters can be integrated into the station.

B. DAMN Hardware

1. The DAMN Universal Module (DUM)

Figure 4 is a functional block diagram of a DAMN universal module (DUM) for the system. This module is straightforward and can be constructed largely from off-the-shelf hardware. With the exception of the micro processor, the Motorola "Dynatac" hand-held terminal contains all of the DUM components plus a power supply and audio I/O components in a package with the following characteristics:

Size	1-7/8" x 3-1/2" x 9"
Weight	45 ounces.

This package was designed to operate in the 900 MHz mobile radio band, providing 200-400 frequency channels with 1 watt output power. The package contains audio components, and digital control logic which uses at least as much volume, weight, and power as the DUM microprocessor and control logic; thus it is apparent that an extremely small, lightweight DUM can be readily developed.

2. Application of the DUM to Network Components

Figures 5, 6, and 7 illustrate how the DUM is used to build up a station, a repeater, or a terminal. The major change in a DUM when operating as part of these three system components is in software in the microprocessor. This software, in fact, is relatively simple, since it need implement only the three search strategies required of station, repeater, and terminal; however, the software may be difficult to implement in minimum memory and minimum speed configuration probably required for off-the-shelf microprocessors. Nevertheless, the software development is a one-time effort and the cost is amortizable over many units.

Packaging of the DUM will undoubtedly be different for repeater, station, and terminal due to the differing mobility and environmental

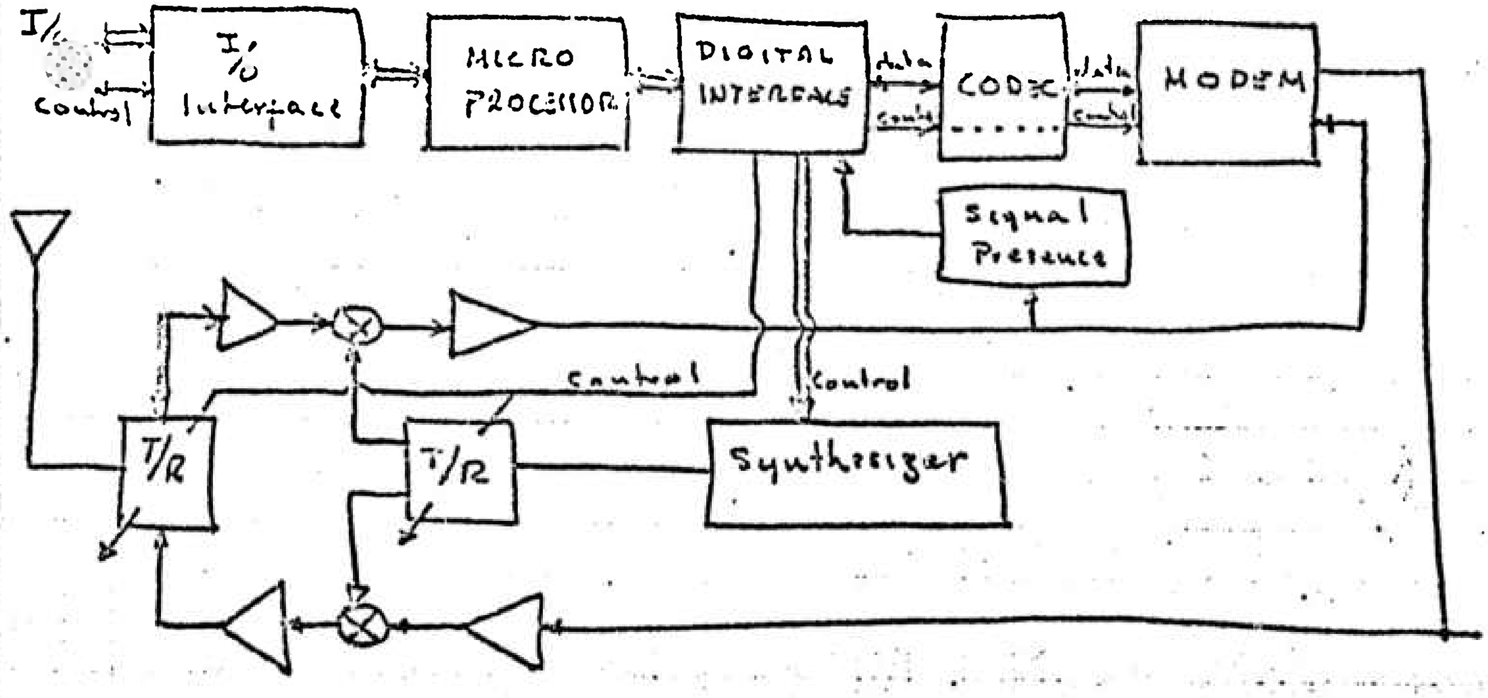


Fig 4 - DAMN UNIVERSAL MODULE

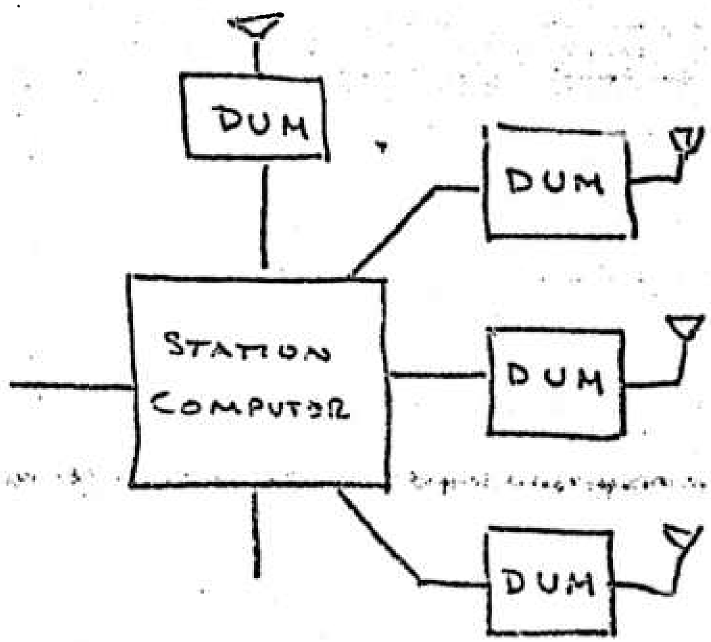


Fig 5 - STATION CONFIGURATION

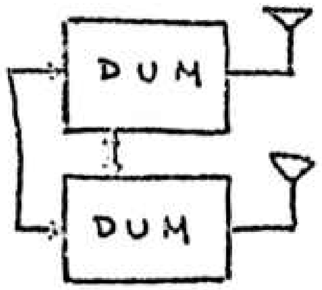


FIG- 6 REPEATER CONFIGURATION

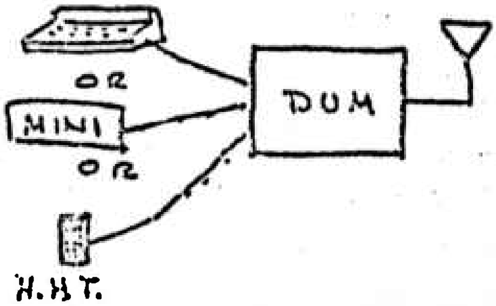


FIG 7 TERMINAL CONFIGURATION

factors. Also, it will probably be desirable to provide either an add-on module to increase the radiated rf energy at the station and repeaters, or to provide the DUM in two power-sizes; one, perhaps 10W transmitted power, for stations, repeaters and portable terminals where power-supplies are not critical; a second, perhaps 1W, for hand-held and ultrasmall portable terminals where size and weight consideration are critical.

3. The DUM Functions

a. I/O Interface

Returning to Figure 4--we see that the DUM I/O interface must be able to connect a wide variety of I/O devices to the microprocessor. The interface should be universal in the sense that it can operate with a large computer (station), minicomputer, keyboard-printer, or another DUM interface. It must be able to accept and offer data in parallel 8 or 16 bit bytes, and provide the necessary control signals to handle I/O functions.

b. Microprocessor

The microprocessor must perform many functions:

1. I/O
2. Formatting messages into packets
3. Addressing
4. Routing and flow control
5. Channel search and selection algorithms
6. Carrier-sense channel access control
7. Acknowledgment generation and checking.

Not every application of the DUM requires all functions, and the functions vary from one application to another, e.g. channel search algorithms are different for repeater and terminal. Thus the microprocessor must be programmable. Perhaps it is possible to develop a station ROM, repeater ROM, and terminal ROM which could be plug-interchangeable to provide the software flexibility.

c. Digital Interface

The digital interface must provide an interface between the parallel data operations of the microprocessor and the serial data operations of the codec and modem. In addition it must perform certain fixed-logic functions such as address checking which would require excess data handling in the microprocessor. It provides control signals to the synthesizer and the T/R switches, and interfaces the signal-present signal into the microprocessor. In certain application, such as at the station it may be desirable to be able to transfer data directly through the digital interface to the I/O port, so the digital interface may provide a DMA channel which bypasses the microprocessor. This last feature is not necessary and need not be implemented if it requires excessively-complex or power consuming circuitry.

d. The Codec

The codec encodes and decodes the digital data. If error detection proves adequate, the codec may be a simple convolutional coder and parity checker. If impulsive noise proves excessive, some more complicated error-correction codec may be required.

e. The Modem

The modem maps the encoded digital packet into a sequence of analogue waveforms suitable for rf transmission. The modem also maps received analogue waveforms into a sequence of binary digits for decoding. The modem must also develop the necessary synchronized clock signals to control all sequential operations of the codec and digital interface. The modem should probably use an MSK (minimum-shift-keyed) type of modulation in a phase-differential mode since this will provide for maximum reliability and use minimum bandwidth for a given radiated power, bit rate, and background noise level.

f. The rf Components

The rf portion of the DUM consists of the antenna, T/R switch, transmitter, receiver, synthesizer and signal-presence detector. The antenna should be plug-changeable so that omni and directional antennas can be interchanged if desired. This may be particularly useful at the station.

The T/R switches control the state of the DUM and must operate as rapidly as possible to provide for efficient channel utilization. It may be necessary to synchronize two T/R switches in the repeater DUM's if separate antennas, carefully designed IF filters, and channel selection algorithms cannot provide adequate channel isolation.

The receiver rf amplifier must cover the entire band of channels so that tuning is not required. Probably a double-conversion if will be necessary to provide the necessary selectivity and channel isolation. One of the conversions must be synthesizer-controlled so that all channels are digitally selectable. Channel bandwidth is nominally 100 kHz to pass the nominal 100 kB MSK signal. Since contiguous channels can be avoided, it is probable that some overlap of channels is acceptable.

The transmitter must up convert the modulated signal to the proper channel and amplify it to provide the required output power. The power amplifier should be power-controlled so that the microprocessor acting through the digital interface can adjust the ERP to the optimum level for network operation.

The synthesizer is the only difficult component in the rf section. It must be capable of generating a stable sine wave at any one of the possible frequencies of operation. Furthermore, during the search mode it must be rapidly switched from one frequency to another while the signal-detector is sampled to determine whether the channel is occupied. Such a synthesizor is clearly possible as demonstrated by the prototype DYNATAC terminal.

The signal-presence detector should be a simple energy-detector with a digitally-controlled threshold setting so that the microprocessor can define the energy level at which a signal is considered to be present. The analog level of the energy detector should be available to the microprocessor so that the relative signal quality on each occupied channel can be evaluated when the terminal is searching for a repeater channel.

III PRELIMINARY PARAMETERS AND ANALYSIS

A. Parameters

In order to provide a concrete basis for comparison with alternative system concepts a preliminary set of parameters has been chosen. No attempt has been made to optimize these except that those principles so far established in the Packet Radio Notes have been applied when possible.

1. Frequency of operation

   1300-1400 MHz

2. Bandwidth occupied

   20 MHz entire possible band

   100 kHz individual channel

3. Modulation

   MSK

4. Bit rate

   100 kbps

5. Terminal and Repeater power

   10 watts

6. Channel Access mode

   Carrier Sense.

B. Comparison with Design Criteria

It is too early for a detailed analysis, but we can make some preliminary estimates of performance for comparison with other systems.

1. Delay and Throughput Efficiency

The achievable delay and throughput will depend very much on the network details; however, we can determine the delay-throughput for a single repeater-terminal area. Since Carrier Sense will be used the results of Kleinrock/Tobagi PRN # 75 will apply. Because an MSK signaling scheme is chosen the throughput represents efficiency in bits/Hz.

Since the frequencies are geographically reusable it is likely that in a large network the efficiency for branches far from the station will be greater than one bit per Hz, even for small delays. Efficiency of the overall network will be determined by the efficiency of the bottleneck at the station. Even though the station has multiple channels, none can be used more efficiently than the carrier-sense access mode will allow. The overall network efficiency will probably approach carrier sense very closely; however, delay will depend on routing and flow control procedures and cannot be analyzed at this time.

It is clear that if multiple stations are used in a given area, the overall efficiency of bandwidth use can exceed one bit per Hz for very reasonable delays. This is not possible with many alternative systems which do not allow geographical reuse of channels.

2. Compatibility

The DAMN system is readily compatible with other types of users in the 1300-1400 MHz band. If sufficient time is given to each repeater to assure that the channel selected for operation is not in use, the system will not interfere with active users. The only type of interference experienced by other users will be on channels which have very low duty cycle. These channels will be abandoned by repeaters after a short period of contention. If the channel-selection mechanism is carefully designed the period of interference should not exceed the duration of a few packets, since the repeater can quickly tell when interference is not packet-network generated.

3. Practicality

The DAMN system concept requires much more study, research, and analysis, but the components can be constructed using off-the-shelf technology. The only development risk involved is one of software; however, careful packaging development effort is clearly needed.

4. Flexibility

The DAMN system seems to be extremely flexible. Furthermore, it makes efficient use of bandwidth in all size networks. It is readily adaptable to a single station with a few terminals, to a single station, multirepeater network, or to a multistation multirepeater network. A large network can readily be implemented one repeater at a time, and no interference need be experienced. As traffic concentrations occur due to changing user patterns the network is easily modified by adding or removing repeaters. Furthermore, so long as repeaters are not isolated by propagation, the network will not fail catastrophically due to the failure of a few repeaters.

5. Operation in Multipath

Since the system operates with 100 kHz, 100 kBps channels, and since excess multipath delay rarely exceeds 5 us (1/2 bit duration) the proposed system would be immune to the intersymbol-interference effects of multipath.

Because of the multiple frequency operation, the effects of fading on a terminal repeater propagation path can be largely avoided.

The DAMN system is almost multipath-proof.

IV SOME ALTERNATIVES

Although the suggested concept seems to be acceptable, there are several alternative designs that deserve examination.

A. Variable Bit Rate

It is possible that a more efficient network would result if each DUM had software-selectable bit rate, so that a single repeater could handle the traffic in a sparcely populated area. Decreasing the bit rate would extend the range of the repeater-terminal link. As pointed out in PRN #28, there exists an optimum bit rate for a given user density and fixed transmitter power. This relation could be used to select bit rate.

B. Spread Spectrum Modulation

The variable bit-rate might also be used to provide some security if it was implemented using a fixed-chip-rate of 100 k chips/sec spread-spectrum system. As the bit-rate is decreased, the spread-factor is increased and either energy per bit is increased, or the energy density in watts per Hz is decreased. The former provides greater range, or some anti-jam protection, the latter provides lower detectability. This mode of operation would use chip-synchronized code-generators to avoid the vulnerability of a fixed-code system.

C. Multiple Bit Rate

It would also be possible to provide the DUM with two or more bit-rates. One could be used for terminal-repeater traffic and the other for repeater-repeater traffic. This might reduce network delay significantly but would probably not affect efficiency.

D. Directional Antennas

The concept can be implemented using only omnidirectional antennas. It is possible that lower power could be used on the repeater-repeater channels if one DUM in each repeater (used to repeat inbound and outbound traffic) were provided with two directional antennas. This does not appear to be desirable since it would require some setup on initial installation, and would make the net more vulnerable to failure of a single repeater.

E. Other Access Modes

It might be possible to improve overall network efficiency by using some form of reservation channel in the link between station and first-echelon repeaters. Since these links will effectively limit the channel efficiency, and will carry the most traffic it is likely that a reservation scheme will improve the network significantly.

V COST

The DAMN system is based on the use of a DUM with each terminal, two DUM's with each repeater, and many DUM's with each station. Terminals will require some additional I/O components, and the station will require additional computing capability and interface to other communication networks. Since all packet-radio networks require the terminal and station elements, the cost of the DAMN concept can be compared to the cost of other concepts by examining the cost of the DUM, and comparing it to the cost of the RF part of other networks.

We cannot readily determine the cost for other networks, but we have a convenient cost measure for the DUM, since it is similar to the DYNATAC hand-held terminal. The exact cost of this terminal is not available in Motorola's literature; however, they are proposing a network in New York, for example, to include 189,000 users. They believe that such a market exists at the hand-held terminal price. This suggests that the hand-held terminal price must be no greater than current mobile-telephone instrument price which is on the order of $500 - $2,000.

VI IMPLEMENTATION

Initially an elementary network could be implemented as represented in Figure 8 . It should consist of one DUM programmed to act as the outbound half of a repeater, and several DUM's programmed to act as terminals. This network can serve to work out hardware bugs in the DUM, evaluate and develop inbound and outbound channel search algorithms, and examine the impact of a real RF environment on the concept. Many of the basic concepts of dynamic allocation cannot be evaluated with such a small network, but RF compatibility and individual channel performance can be tested.

To evaluate some of the more complex concepts involving multirepeater operation, a multirepeater net would be required. The first step would be to implement two repeaters and a station as shown in Figure 9 . This would require at least six DUM's. In addition six or more DUM's programmed as terminals are needed.

The hardware for these elementary networks should be made as small and lightweight as off-the-shelf technology will allow. A DUM can certainly be built in a small attache case with little packaging effort. With some packaging effort a DUM should be smaller than the DYNATAC hand-held terminal. Perhaps this packaging effort should await the results of tests with a multiterminal network.

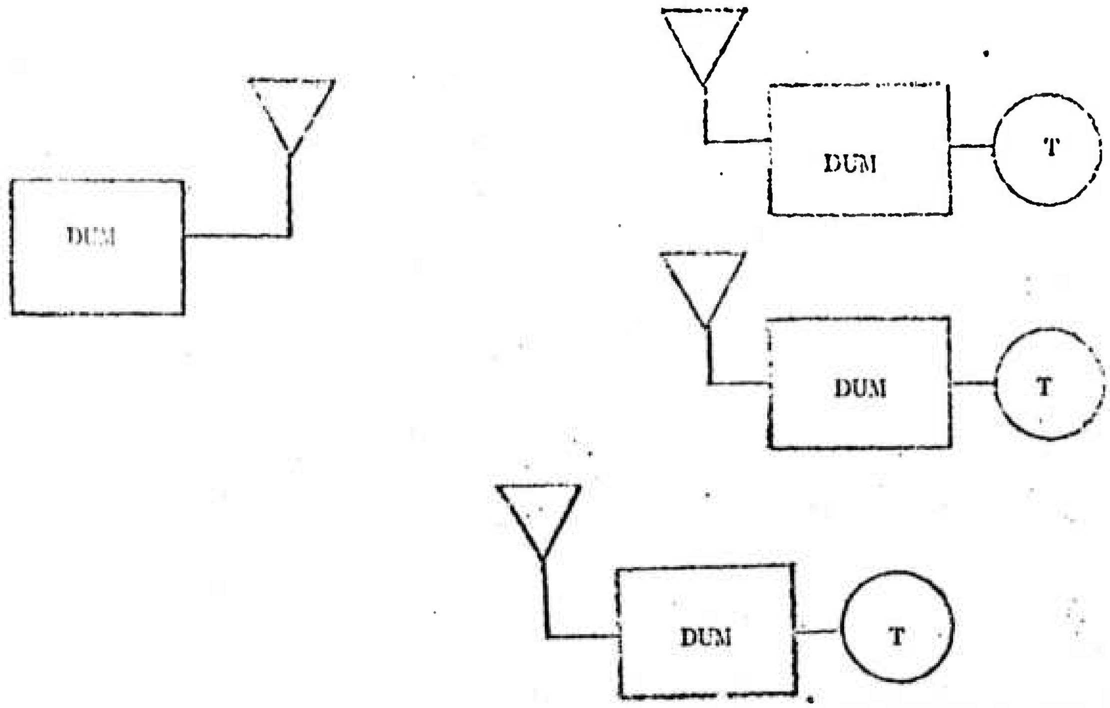


Figure 8 Elementary Net

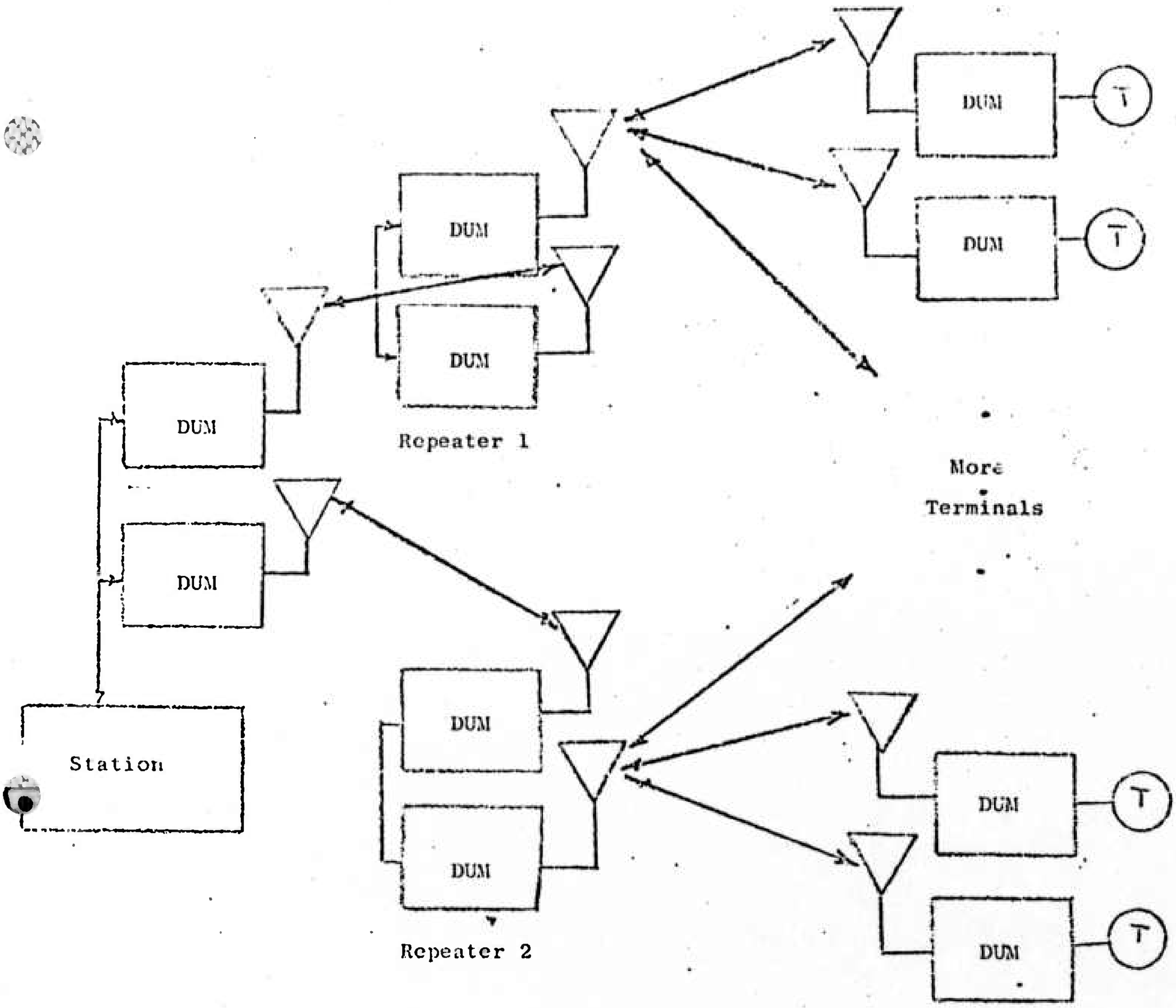


Figure 9 Two Repeater Experimental Net

Ultimately, to test all network routing and flow control techniques, to demonstrate performance in difficult RF-interference environments, and to shake out all hardware and software bugs a larger network must be implemented. This network should include the effects of a packaging effort by integrating smaller, repackaged DUM's, hand-held terminals, and multistations into the initial network.

Appendix B. 4

(Note on Radar Tests)

Appendix B.4

(Note on Radar Tests)

Introduction

As a part of our participation in the exploration of packet radio networks, SRI is performing a series of r.f. channel measurements. This series includes tests involving the transmission and reception of spread-spectrum signals in the 1- 2 GHz band to measure the channel impulse-response.

On January 7 and 8, 1974, a series of tests were made to determine the effect of these spread-spectrum signals on two nearby radar systems. The radars were an Air Force AC&W search radar at Mt. Tamalpais and an FAA in-flight surveillance radar at Scarper Peak. This report describes the tests.

Description of Tests

General

To facilitate channel measurements, SRI has recently modified a 14-foot GMC step-van to operate as a mobile laboratory. A packet-radio test transmitter, constructed for ARPA by Collins Radio Corporation and described in another temporary packet radio note was installed in the van for these tests. Dr. R. Kahn (ARPA) monitored the tests and observed the effects of interference on the Air Force radar at Mt. Tamalpais. Dr. D. Nielson (SRI) directed the tests from the van, and Dr. S. Fralick (SRI) observed the effects of interference on the FAA radar at Scarper Peak. We very much appreciated the help of Mr. R. Young (FAA site engineer) and M/Sgt. Jensen (Air Force) who acted as a test control group to assure that the tests would not cause operational problems.

The tests were performed by transmitting spread-spectrum signals from two sites (one site each day) and simultaneously observing the

PPI (Plan Position Indicator) and A-scope displays at the two radar sites. For various logistic and security reasons, it was possible to photograph the PPI display at the FAA site only.

Geography

The geography of the tests is shown in Figure 1. The FAA radar is located at Scarper Peak, the Air Force radar is located at Mt. Tamalpais, and the transmitter sites were at Vista Point on Highway 280 and Twin Peaks in San Francisco. Elevations and distances are shown in Figure 1.

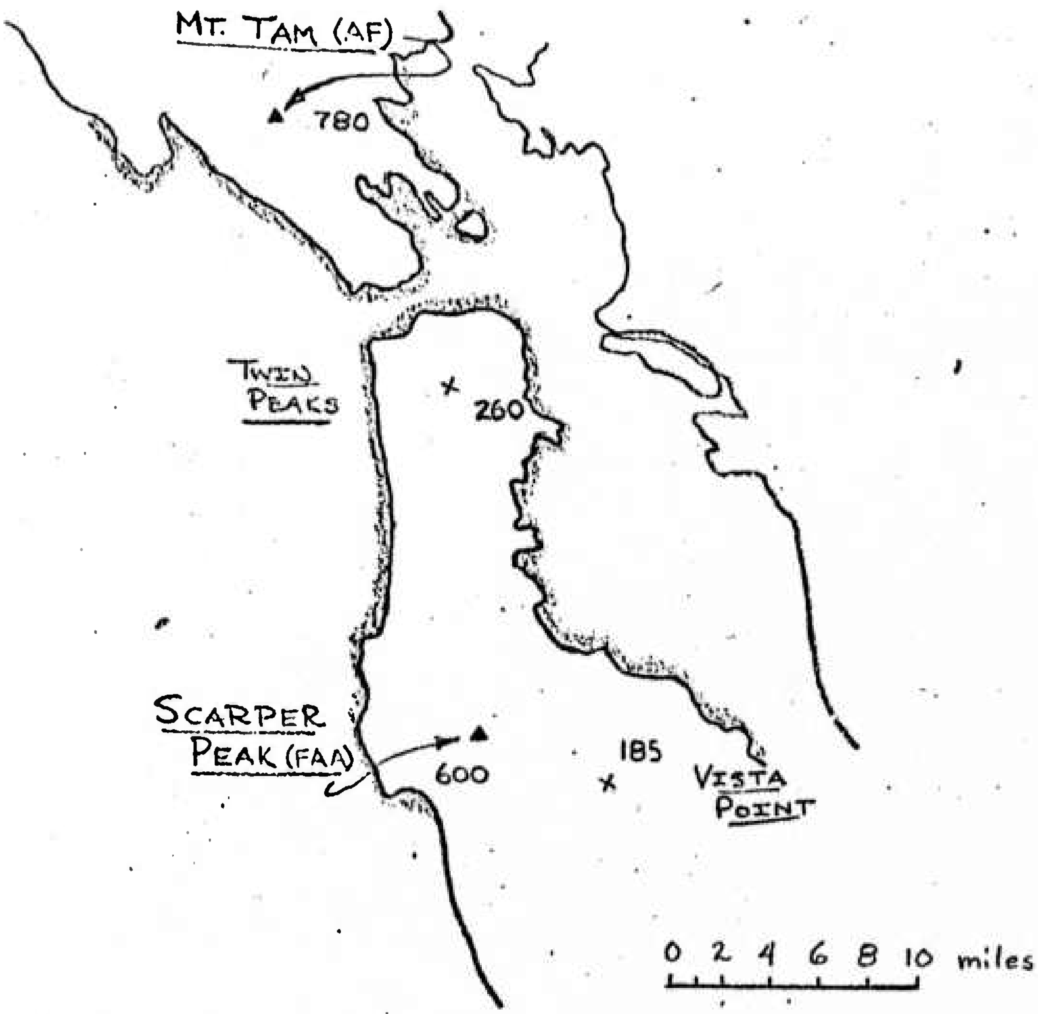
Transmitted Signal

The signal transmitted was a 6-watt spread-spectrum signal generated by the Collins packet radio test set. This test set generates a binary-valued non-return-to-zero pulse train from a 127 chip maximal-length sequence generator. The nominal chip rates are 10 and 20 megachips/sec. Each 127-chip sequence is called a bit, so the nominal bit-rate is about 80 and 160 kilobits/sec. Various bit-sequences are possible, but for most of the tests a continuous bit-sequence of 1's was used, so that the chip-sequence was repeated every 127 chips.

To transmit the signal a chip-sequence is used to biphase modulate a carrier at 1325 MHz. The biphase modulated carrier is filtered to reduce the transmitted energy outside a nominal 40 MHz band centered at 1325 MHz. The transmitted spectrum is shown in Figure 2.

The test-set transmitter was modified by Collins just prior to the tests so that the duty-cycle of the transmitted signal can be controlled either automatically or manually. This control is essentially an on-off key with independent control of the duration and repetition rate of the "on" periods. During an "on" period, the transmitter radiates a number of code sequences determined by the test-set controls.

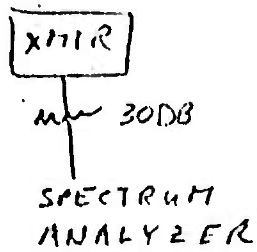
Two antennas were used in the test. One is an omni-directional antenna with nominal gain 2 dB. The other is a horn with nominal gain of 13 dB whose pattern is shown in Figure 3.



DISTANCES (km)

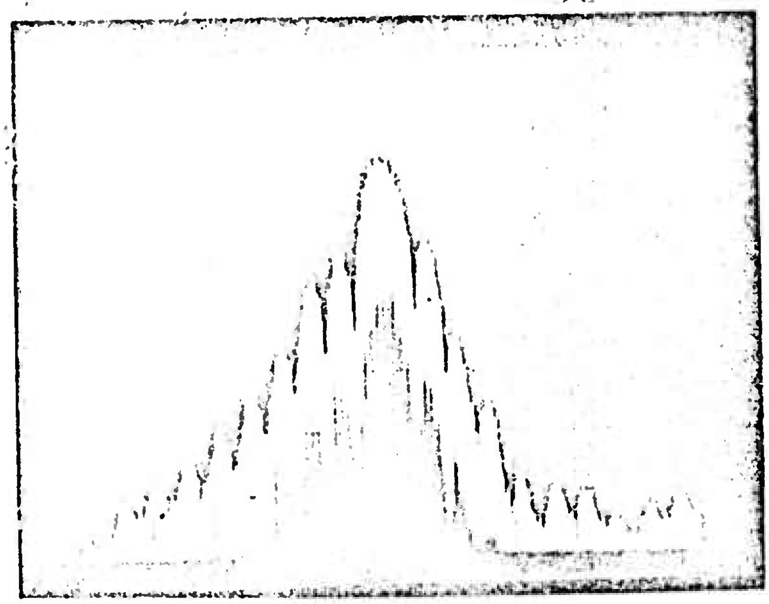
	SCARPER PK.	MT. TAM
TWIN PEAKS	22.5	21
VISTA POINT	8	48

TRANSMITTER SPECTRUMS CON'T 1325MHZ



10M CHIP CONTINUOUS CODE MODE

1402-809

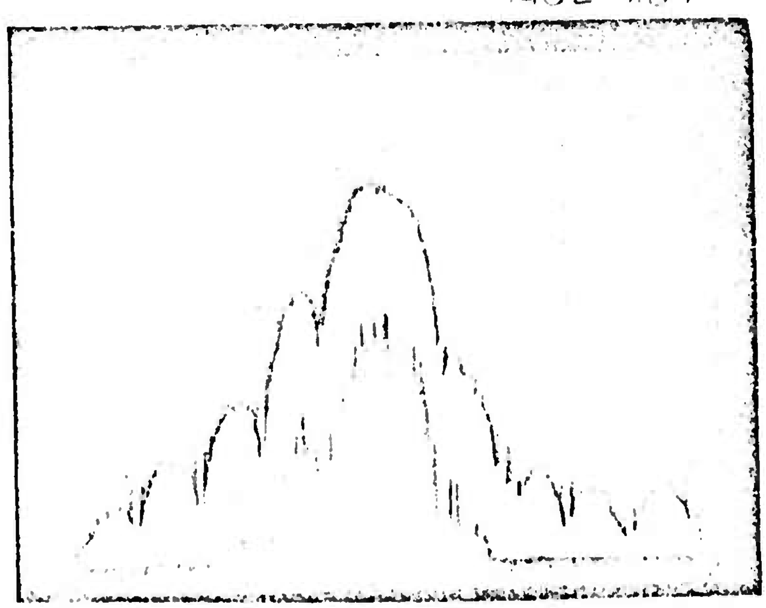


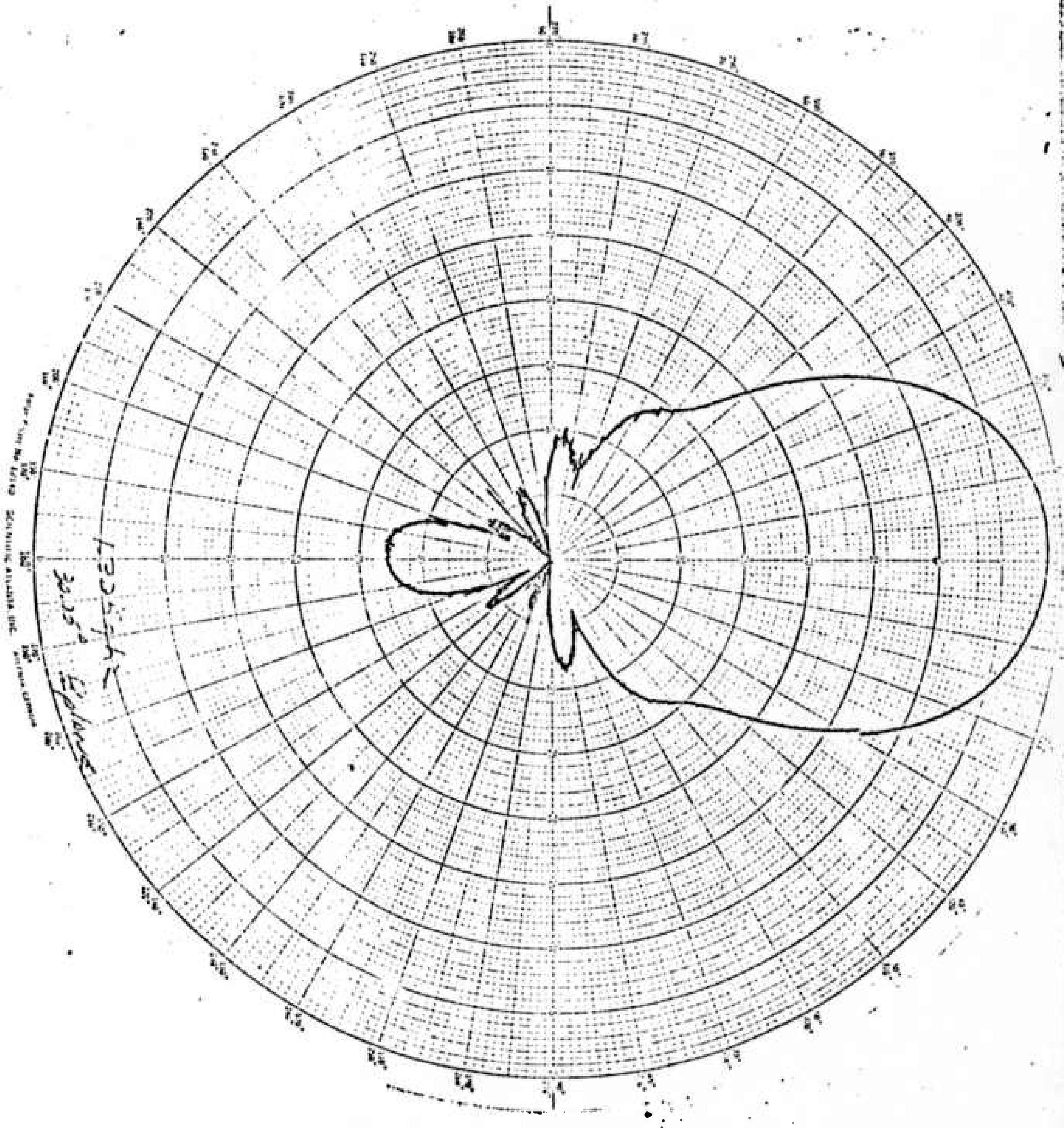
LOG REF +10DBM
BW 300kHz
f_c 1325MHZ (APPROX)
SCAN 20MHZ/DIV

20M CHIP CONTINUOUS CODE MODE

SAME TEST SET UP & SCALES

1402-809





Radar

The FAA radar is a general-search radar used for in-flight surveillance. During tests it was tuned to 1338 MHz. It transmits a 2 μs pulse with 4 MW peak-power on an antenna with 34 dB gain. The prf is nominally 333 pps. The receiver bandwidth is nominally 2 MHz.

If we assume a noise figure of 5 dB, then the receiver noise-level should be about -106 dBm referred to the antenna terminals or -140 dBm including antenna gain.

The Air Force radar is similar in all respects; however, precise details were not available.

Both radars are used to detect and track aircraft in flight. The antenna is rotated approximately five times per minute, and the raw-video return is processed digitally and displayed to flight controllers for the FAA.

The raw-video return at the FAA site was observed both on a plan-position-indicator (PPI) and on an A-scope. The A-scope is a standard oscilloscope with the horizontal sweep synchronized with the transmitted pulse train. The vertical displacement is proportional to the envelope of the received signal. Hence targets appear as pulses which occur at a time proportional to range. They will appear and disappear as the radar antenna sweeps past the target and illuminates it. Noise appears as random pulses which look somewhat like grass. An interfering signal, unsynchronized with the radar will raise the general noise level and may mask targets. The level of the interfering signal will vary with the direction the antenna points since it will be proportional to the gain of the antenna in the direction of the interfering transmitter. Although the A-scope is a good indicator of interference, and can provide some absolute measure of intensity, it does not provide the long-time integration effect seen by a radar observer, so it does not completely describe the effect of interference.

The plan-position-indicator (PPI) is a CRT display with the electron-beam position swept along a radial line, synchronously with the transmitted pulse so that the position of the beam radially from the center of the CRT is proportional to the range.

The PPI is a polar-coordinate display of the received video (r.f. envelope) signal. The electron beam is swept radially, once for every transmitted pulse. The azimuth of the radial sweep is continuously increased in synchronizm with the angular sweep of the antenna. The beam is intensity-modulated with the received video, so that a pulse will cause a bright spot on the phosphor. The phosphor persistance is usually set so that the bright spot fades almost completely in one complete antenna sweep. Targets appear on the PPI as a bright dot at an azimuth and range from the CRT center, proportional to true azimuth and range from the radar antenna. Noise appears as random bright dots on the screen.

The appearance of an interfering signal on a PPI depends on the strength and duration of the signal, and on the antenna direction when the signal is sent. For example, a 1 ms burst which was transmitted with sufficient amplitude to exceed the noise level would appear as a radial-line segment on the PPI. The length of the line would depend on the CRT sweep speed. If the sweep speed was adjusted so that the total radius (approximately four inches) = 300 miles, then the beam would be swept radially at a rate of 1.33 inches/ms. A 1 ms signal burst would appear as a line 1.33 inches long. A 100 μs burst would appear as a line .13 inches long.

Interfering signals which last longer than one inter-pulse period will be displayed as a sequence of bright radial lines which sweep out a pie-shaped segment of full radial-length and angular spread equal to 360 d/12 degrees where d is signal duration in seconds.

As the interference amplitude level decreases, the brightness of the display will decrease (within the phosphor dynamic range) until the line will no longer be continuous, but will be a sequence of dots, and finally will disappear altogether. The amplitude level will change as the antenna rotates, so that we would expect maximum brightness of the display when the antenna points in the direction of the interfering transmitters and darkness when antenna-pattern nulls point toward the transmitter. These effects were all demonstrated during the tests and are illustrated later.

Test Procedure and Results

On February 7 the packet radio transmitter was set up at Vista Point on Highway 280 (see Figure 1).

Our first objective was to determine if we could radiate low duty-cycle, full power signals on the omni-antenna without interfering with the radars. To accomplish this we transmitted full-power spread-spectrum pulses and varied the pulse duration. Pulse durations of 120 μs and 1 ms were radiated once per second. The results of these tests are shown in Figures 4 and 5, 6. We also manually generated one-second bursts, and once generated a 15 second burst. The one second burst is shown in Figure 6. The 15 second burst is not shown.

Figure 4 shows a triple exposure of the Scarper Peak PPI quadrant containing the azimuth of the transmitter. During this exposure we were transmitting 120 μs bursts. No interference was observed for this or any shorter duration test signal.

Figure 5 shows a triple exposure of the Scarper Peak PPI, same quadrant. During this exposure we were radiating 1 ms bursts of test signals. Although the interference is clearly visible, it does not appear to mask targets, or to be otherwise seriously objectionable.

Figure 6 shows a single exposure of the Scarper Peak PPI, Northeast quadrant. During this exposure we were radiating a one second burst. The variation in intensity is due to the fact that the antenna was pointed roughly at 90° to the test transmitter, and the interfering signal was being received on the antenna side lobes. Similar interference was noted on the back lobes. In fact, during the 15 second burst, the entire PPI was illuminated with bright radial lines. It is clear that such interference would cause serious problems.

Our second objective was to determine the maximum power which we could radiate without interference. This was accomplished by radiating manually controlled pulses of $\frac{1}{4}$-$\frac{1}{2}$ second duration, and attenuating the

TRANSMITTER AT VISTA PT., OMNI FULL POWER

1 SEC. BURST

FIG 4.

1 ms BURST

FIG 5.

120 μS BURST

FIG 6

power delivered to the test-transmitter antenna. The results of this series of tests are shown in Figures 7, 8, 9, and 10.

Figure 7 shows the effect of full-power bursts, Figure 8 shows the effect when 10 dB attenuation was inserted, Figure 9 shows 20 dB attenuation, and Figure 10 shows 22 dB attenuation. No interference was observed for attenuation in excess of 22 dB. Preliminary calculations of propagation loss indicate that the radar should receive a signal at the nominal r.f. noise level with 55 dB attenuation inserted. Since the background noise peaks are being displayed, one would expect to see interference down to 40-50 dB attenuation. This discrepency has not been satisfactorily explained.

During these tests, no interference was observed at Mt. Tamalpais. Since Mt. Tamalpais was six times as far away as Scarper Peak, we expected the test signal to appear at Mt. Tamalpais with an attenuation of about 14 dB relative to Scarper Peak. This did not occur. When we transmitted the test signal using a horn antenna with 13 dB gain, pointed at Mt. Tamalpais, we still did not cause noticeable interference. The reason for this is unknown; however, it may be due to special ECM circuitry at Mt. Tamalpais.

On February 8 we repeated the tests with the transmitter located at Twin Peaks. The pulse-duration test results were the same as those of February 7.

Figures 11, 12, and 13 show the result of the power-level tests. In this case, the pictures are of the quadrant containing true North, hence the azimuth to the transmitter. So long as at least 13 dB of attenuation was inserted at the transmitter, no interference was observed. This is to be expected since the range to Twin Peaks is three times the range to Vista Point, so the propagation loss should be about 9 dB additional.

The radar at Mt. Tamalpais did not observe interference during this sequence.

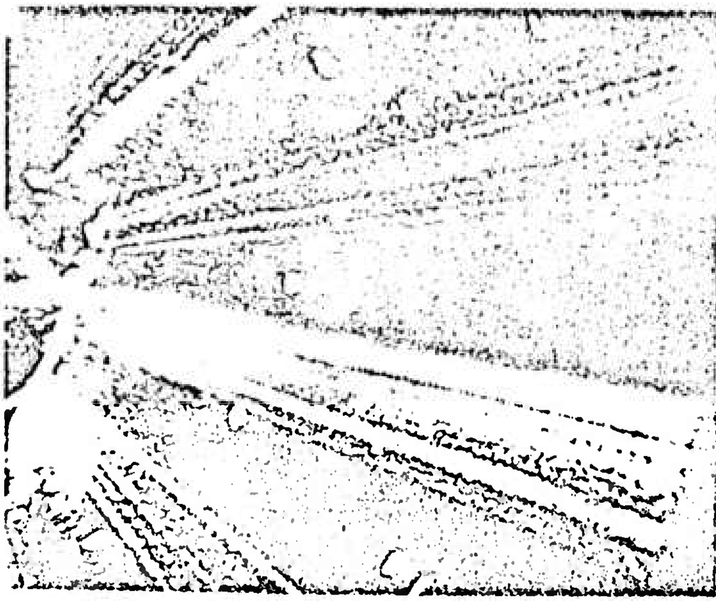


FIG 7. - FULL POWER

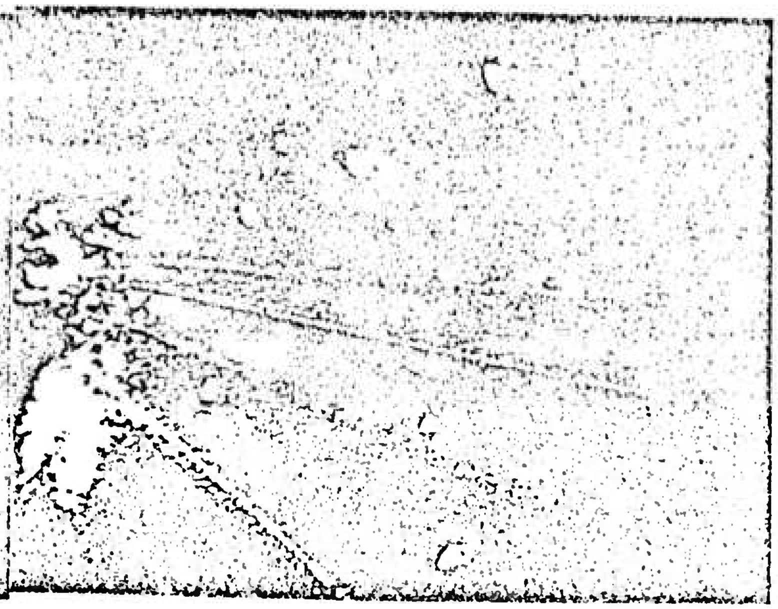


FIG 8. - -10 dB

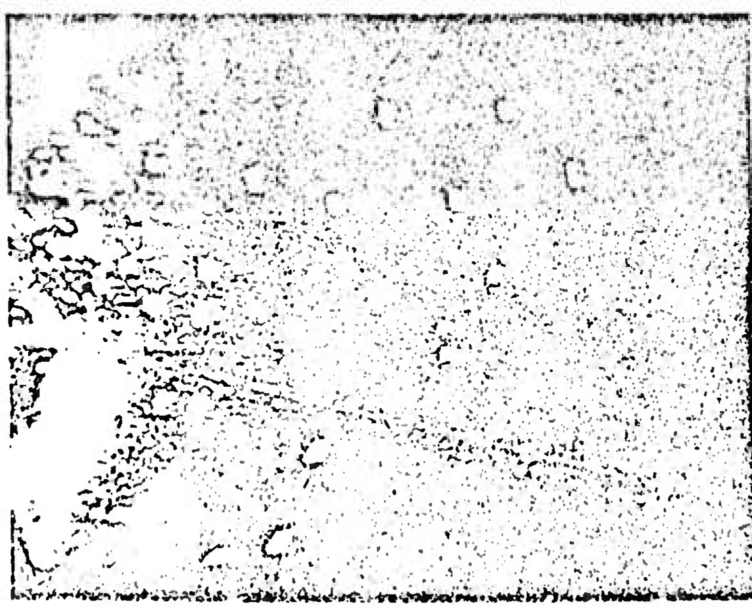


FIG 9. - -20 dB

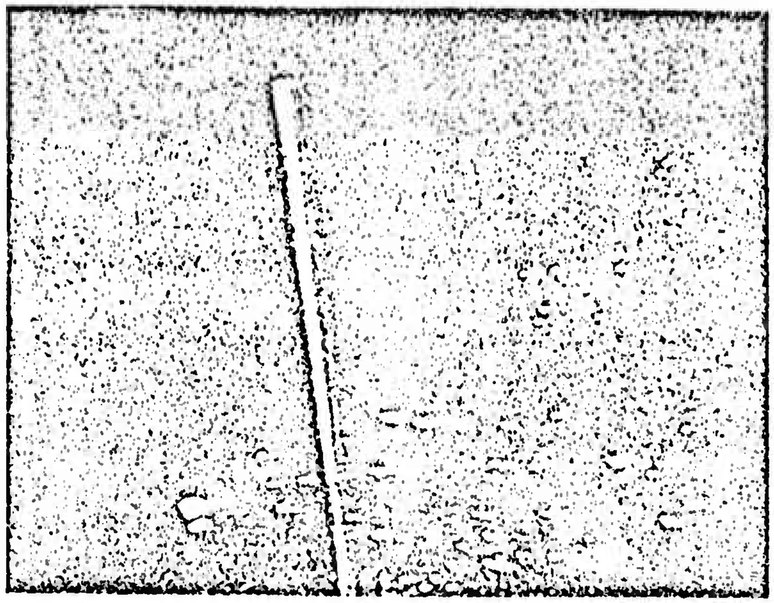
FIG 10. - -22 dB

TRANSMITTER AT VISTA POINT, OMNI ANTENNA, FULL POWER. MANUALLY SWITCHED ON-OFF FOR 1/4 - 1/2 SEC INTERVALS.

TRANS. AT
TWIN PEAKS, OMNI
CONT. CODE

FULL POWER

FIG 11.

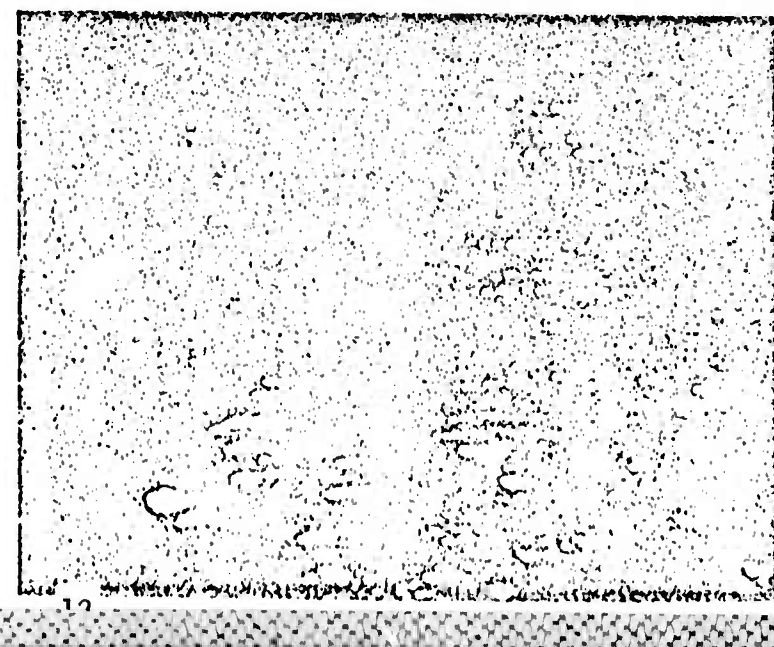


-10 dB

FIG 12.

-13 dB

FIG 13.



In order to generate interference at Mt. Tamalpais, it was decided to radiate using the horn antenna, and pointing it directly at Mt. Tamalpais. Using 1 second manually-generated bursts, the interference was at last observed. No pictures of this effect were made.

Finally, to determine whether we could radiate using the horn at Twin Peaks to illuminate San Francisco without interference to Scarper Peak, we radiated full power with the horn, and rotated it to determine the "no-interference" limits with Scarper Peak. These were roughly a 140° arc pointed away from Scarper Peak. A picture of the interference at the 180° point is shown in Figure 14. To check for anomolous horn gain we inserted 20 dB attenuation and rotated the horn 360° while radiating $\frac{1}{2}$ second bursts. The maximum interference occurred with the horn pointing at Scarper Peak. Results are shown in Figure 15.

Conclusions

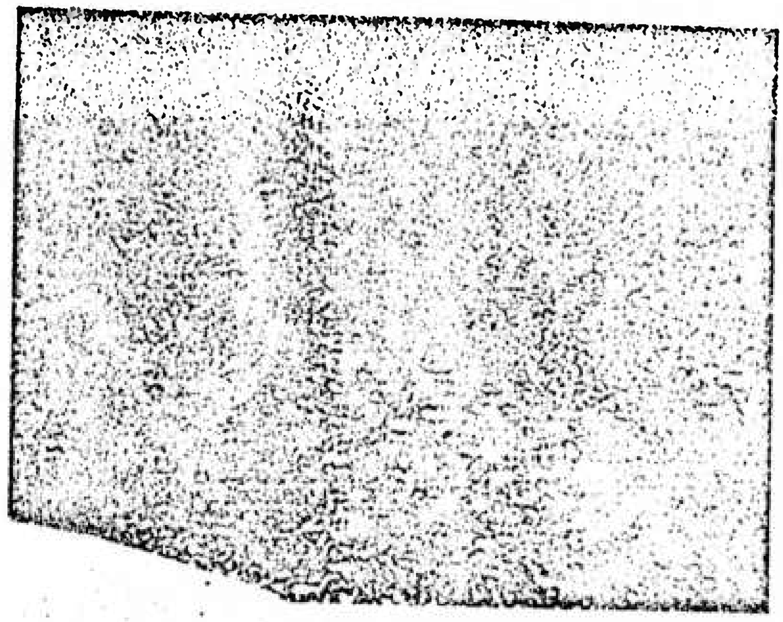
The results of these tests indicate that we can radiate enough energy for our propagation tests so long as we use bursts shorter than 120 μs once per second, or so long as we use a directional antenna which provides at least 13 dB (22 dB) attenuation in the direction of Scarper Peak from Twin Peaks (Vista Point).

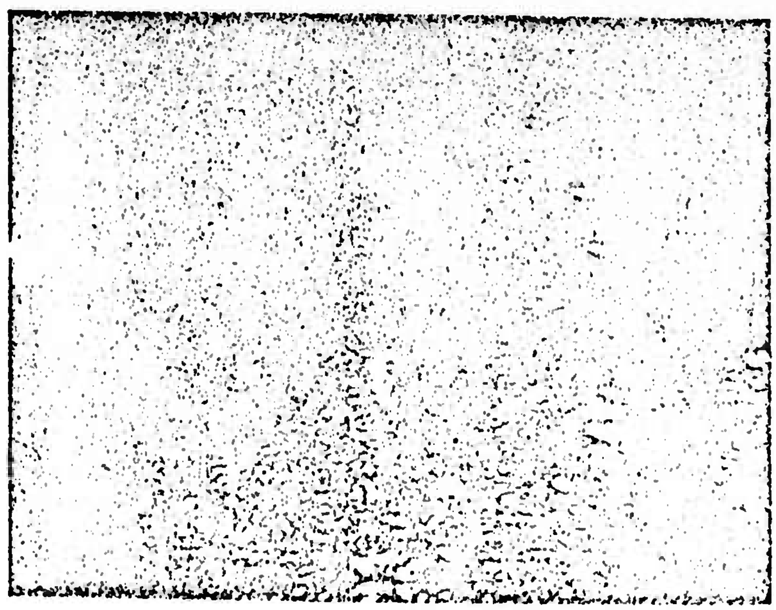
We assume that for future propagation tests these results may be scaled for range using a propagation loss proportional to the inverse squares of range, and for frequency offset by considering the spectral energy density of the transmitted signal relative to that transmitted at 1338 MHz when the carrier is 1325 MHz.

TRANSMITTER AT
TWIN PEAKS

HORN AIMED AT
SCARPER PK
WITH 20dB
ATTEN

FIG 14.





HORN, NO ATTEN.
AIMED AT SCARPER $\pm 90^\circ$

FIG 15.

NO NOTICEABLE EFFECT
FOR ALL AZ. BETWEEN
THESE LIMITS

Appendix C.1

Modulation Waveform Types

1. INTRODUCTION

Several basic pseudo-noise spectrum spectrum modulation approaches are considered as potential candidates for RF transmission over the ARPA Packet Radio Network. A comparison of these different modulation approaches is summarized in Figure 1. The modulation approaches considered are differentially coherent bi-phase PSK, differentially coherent quadri-phase PSK, coherent MSK, differentially coherent (DC) MSK, pseudo-orthogonal (PO) MSK and noncoherent (NC) 8'ARY MSK signalling. PSK and MSK refer to chip modulation, and coherency refers to bit modulation.

Each of these approaches will be examined with respect to their implementation and performance. Each approach considered is flexible in that the processing gain can be increased by using a larger number of chips per bit. Also, frequency hopping can be readily superimposed upon the pseudo-noise spread spectrum baseline concept.

Each approach is also shown implemented at both the modulator and demodulator with surface acoustic wave devices (SAWD). The advantages and potential of these devices appear very great, particularly with respect to considerations of fast acquisition, low power and size requirements. These factors are especially important for the hand-held packet radio transceiver.

2. PSK MODULATION/DEMODULATION

2.1 Differentially Coherent Bi-Phase Modulator/Demodulator

The differentially coherent bi-phase PSK modulation approach was discussed in the Packet Radio Temporary Note #33. The modulator differentially encodes the data. Each time a logic 1 bit is to be transmitted, the bit sequence is inverted. However, each time a logic 0 bit is to be transmitted, the code sequence remains the same. One implementation of the modulator is shown in Figure 2. For the baseline system, spread factors of 100 are assumed; therefore, a data bit "1" transmits a sequence of 100 phase modulated chips with each chip either 0° or 180°. The data sequence from the source may be represented by the following sequence

$$1\ 0\ 1\ 1\ 0\ 1\ 0\ 0\ .\ .\ .$$

After differentially encoding the data, the sequence becomes

$$1\ 1\ 0\ 1\ 1\ 0\ 0\ 0\ 1$$

The impulse generator drives the 100 tap SAWD with either a positive or negative impulse. The output of the SAWD at the point where the taps are summed is

$$s(t) = \sum_{i=1}^{100} d_{101-i} \cos (W_0 t+\phi) \ [u(t-(i-1)T_c) - u(t-iT_c)] \ldots$$

where d_i is either $\pm$ 1 depending upon the sign of the chip and T_c is the chip duration.

An alternate DPSK modulator is shown in Figure 3. The chip sequence for a data bit is exclusive-ored with the differentially encoded data bit. The impulses are generated on a chip basis yielding a phase modulated carrier at the SAWD output. A positive impulse may correspond to a 0° phase shift and a negative impulse corresponds to a 180° phase shift.

One of the shortcomings of bi-phase PSK modulation is its spectral density response which, as shown in Figure 22, rolls off as $1/f^2$ or 20 dB per decade.

The DPSK demodulator is shown in Figure 4. It is a matched filter at the IF. As shown, a single SAWD is matched to the chip sequence and the same chip sequence delayed one data bit time. The outputs of the taps are summed both for the present received bit and the previously received bit. This summing operation takes place right in the SAWD. The sum and difference is obtained from the two outputs. These are envelope detected, compared and the output is sampled. The sample time is established by the preamble detector and the sampling occurs at the peak of the baseband pulse. The sign of the sampled signal determines whether the data bit is a "1" or a "0".

Maximum Likelihood Detection

This type of detection assumes ideal bit timing and ideal sampling of the correlation waveform peak. Ideal timing is never achieved for bit timing. For non-ideal bit timing, Figure 5 shows the result of loss of signal envelope detection versus timing error. Figure 6 shows the loss of signal/noise (E_b/N_o) as a function of number of chips per bit with timing error Z as a parameter for T_B of 10μs (100 KBPS data rate).

Hence, for N = 100 and 10ηs bit timing error for 100 KBPS data, a 1 dB loss of E_b/N_o is incurred. Note that 10ηs out of 10μs bit period is .36° of timing accuracy and requires a 100 MHz clock if a counter is allowed to freerun from a master clock. A 1.0° error (28μs) results in 3 dB loss. Thus, bit timing acquirements are quite important for spread spectrum maximum likelihood detection.

Gated Peak Store Detection

An alternate method of data detection is achieved by sampling the SAWD matched filter output by a timing window that is $2T_c$ (T_c = chip interval) wide, storing the peak value during that timing window and making a bit decision at the end of the timing window. This method greatly alleviates the bit timing problems that the maximum likelihood detection requires since ideally, bit timing of only T_c accuracy is required (100ηs for N - 100).

It is estimated that this form of detection degrades performance by .5 dB (for a $2T_c$ window case) from the ideal timing. A $2T_c$ window is suggested for two reasons. One is that a $2T_c$ window is easy to generate. The second is that intuitively, a window equal in width to the correlation peak width would result in best performance.

Thus, for much easier timing requirements, a .5 dB performance penalty results. Hence, performance of DCPSK with $P_e = 10^{-5}$ requires E_b/N_o = 10.9 dB rather than 10.4 dB. A suggested method of implementing the detector is shown in Figure 7.

2.2 Differentially Coherent Quadri-Phase PSK Modulator/Demodulator

Quadri-phase modulation has certain advantages over bi-phase modulation. One is that the timing requirements are less severe since the symbols are twice as long as the bits. Another is that the bandwidth requirements are less as shown in Figure 22.

An implementation of the modulator is shown in Figure 8. The demodulator for quadri-phase PSK is similar to that shown in Figure 4 for bi-phase demodulation. An orthogonal channel is required which is phase shifted 90° from the other channel. This can be achieved by a 90° phase shift of the two SAWD outputs (A and B). These 90° phase shifted outputs are input to a hybrid, and the sum and difference of the outputs are envelope detected and compared in the same manner as the other channel.

Maximum likelihood detection and gated peak store detection criteria discussed for bi-phase apply for quadri-phase, except symbol intervals ($T_c/2$) must be used in the applicable equations and curves. Note that there are 2 chips per symbol for quadri-phase.

3. MSK MODULATION/DEMODULATION

3.1

An alternate approach to PSK signalling is sometimes referred to as Minimum Shift Keying (MSK). It is a two orthogonal channel amplitude weighting PSK modulation scheme. The in-phase and quadrature phase carrier channels are staggered by one chip and shaped with orthogonal cosine weightings of two chip length.

This is illustrated in Figure 9. To achieve a constant envelope signal, the quadrature channel is added to the in-phase channel as shown in the MSK waveform diagram. Also shown is the resultant phase and frequency for the example chip stream. Phase transitions occur at the null points of each sub-channel. This property results in phase continuity of the MSK waveform.

This property produces increased attenuation of the higher signalling frequencies. Its spectral density rolls off as $1/f^4$, or 40 dB per decade as shown in Figure 22.

3.1 Coherent MSK Modulator/Demodulator

Several approaches to MSK modulation/demodulation are being considered. The first of these approaches is termed coherent MSK. A data bit "1" consists of a sequence of 100 chips and a data bit "0" consists of the same sequence with inverted signs. Figures 10 and 11 illustrate two implementation techniques for the modulator. In Figure 10 the data source determines the sign of the impulse driving the SAWD. The input transducer has a cosine pulse weighting and has a duration of 2 chip periods. The SAWD has 100 taps, each tap corresponding to one of the chips. The even taps are summed for one MSK subchannel and its output is phase shifted 90°. The odd taps are summed for the other subchannel. The two outputs are then added together yielding the MSK signal. The signal can be represented mathematically as follows:

$$s(t) = 2P\,[\cos(W_0 t+0) \sum_{k=0}^{49} d_{2k+1} \operatorname{cosp} \frac{\pi}{2T} (t-2kT) +$$

$$\sin(W_0 t+\emptyset) \sum_{K=0}^{49} d_{2k+2} \operatorname{sinp} \frac{\pi}{2T} (t-2kT)] \;\ldots$$

where cosine pulse "cosp" is defined by:

$$\text{cosp}\ \frac{\pi t}{2T} = \cos \frac{\pi t}{2T}\ [\mu(t+T) - u(t-T)]$$

and sine pulse "sinp" is defined by:

$$\text{sinp}\ \frac{\pi t}{2T} = \sin \frac{\pi t}{2T}\ [u(t) - u(t-2T)]$$

and the chip interval is given by T.

The IF output is then up converted and amplified for transmission. The summing of the odd and even taps is accomplished in the SAWD. It may be feasible to also do the phase shifting in the SAWD. However, further investigation in this area is required.

An alternate approach to the modulator is shown in Figure 11. The SAWD is impulsed at the chip rate. The sign of the impulse determined by the exclusive output of the SAWD generates a cosine weighted pulse which exists for two chip periods. The impulse sign determines carrier phase in the cosine pulse.

The demodulator is shown in Figure 12. The demodulation process is coherent. The matched filter demodulator is identical to the modulator discussed above (Figure 10) except that the tap sequence is reversed. The summing of the two SAWD output results in an amplitude modulated carrier which is the correlation function. This IF signal must be sampled at exactly the right time because the phase of the IF carrier contains the information. Hence, near coherence is required which can be achieved by use of a phase lock loop which encloses the surface wave detector.

In a half-duplex application, it is possible to share the SAWD between the transmitter and receiver (c.f. Figures 10 and 12). The complexity of the switches must be compared to the complexity of a SAWD.

3.2 Differentially Coherent MSK (DC MSK) Modulator/Demodulator

To alleviate the severe timing requirements of the coherent MSK demodulator, an alternate approach is considered. With this approach the data is differentially coded prior to being modulated as was discussed in Paragraph 2. A data bit from the source is differentially encoded with the previously encoded data bit. The remaining part of the modulator (Figure 13) is identical to the modulator discussed above for the coherent MSK approach. The modulator can also be constructed using the 2 chip SAWD as shown in Figure 11. The only additional requirement is the differential encoding of the data source output.

A differentially coherent MSK demodulator is shown in Figure 14A. Two SAWD's are shown. One is a 10μsec (1 bit) MSK demodulator and the other is a 10μsec delay. The demodulator coherently detects the present data bit and delays the previously detected bit for one bit interval. These two outputs are then multiplied coherently and low pass filtered yielding a baseband output whose sign determines the data bit. The bit timing derived from the preamble is used to sample the mixer output and recover the data. A second DC MSK demodulator is shown in Figure 14B. The concept is the same as the DPSK demodulator shown in Figure 4. The advantage is the elimination of phase coherent mixer.

The two types of detection discussed for DPSK signalling are also applicable for MSK signalling. The effect of timing errors upon the loss of signal energy is shown in Figure 15. For MSK signals the correlation function is shown in Figure 15. For small timing offsets the loss factor is less than it is for 4-phase PSK signalling. For timing errors in excess of 1/4 of a chip the loss factor is less for 4-phase PSK signalling. With 2-phase PSK signalling, any and all timing offsets will be worse than 4-phase PSK or MSK.

3.3 Pseudo-Orthogonal MSK (PO MSK) Modulator/Demodulator

Another MSK modulation approach is illustrated in Figures 16 and 17. To modulate the pseudo-orthogonal signals requires the selection of one of two chip sequences. These chip sequences have approximately zero cross correlation. In Figure 16 a data bit "1" selects, for example, the upper SAWD for the generation of the MSK signal and the data bit "0" selects the lower SAWD. In Figure 17 the data bit selects the chip sequence which controls the impulse generator driving the MSK SAWD.

At the demodulator illustrated in Figure 18 the incoming signal is correlated by both SAWD's, each containing one of the chip sequences. Each SAWD is an MSK detector for a bit. One SAWD is matched to a data "1", the other SAWD is matched to a data "0". Each output is envelope detected and the output with the largest amplitude is selected. This approach requires two SAWD's each of 10 sec in length. The timing requirements are the same as those for the differentially coherent MSK demodulator. In this type of data detection, maximum likelihood detection and gated peak store detection may be used.

8'ARY MSK Modulator/Demodulator

Another signalling approach which yields a performance improvement over the approaches discussed above at low bit error probabilities is shown in Figures 19 and 20. The modulator utilizes 8 SAWD's. Each SAWD has 300 taps and the taps are arranged such that they correspond to the 300 chips for each code symbol. The cross-correlation between the eight 300 chip sequences is approximately zero. However, each code sequence has an autocorrelation function that has a large peak four chips wide and very low sidelobes. Three data bits from the source comprise a symbol and are used to select one of the eight chip sequences. For the chip rate in the channel to be 10 megachips per second requires 300 chips per symbol. Again, there is an input transducer for each of the SAWD's which has a cosine pulse weighting.

At the demodulator the IF signal is power split between eight SAWD's each of which is matched to one of the 300 chip sequences. Again, each is a matched filter for one of the eight possible 3 bit data sequences. The output of each filter is envelope detected and every symbol time the largest output is selected. The most likely bits are the three bits associated with the largest output. The timing requirements for this approach are more stringent than DC MSK since for 300 chips, the same correlation peak occurs. One of the drawbacks is the number of SAWD's required and the lengths of the SAWD's (30μsec). This can be alleviated in the modulator by using programmable SAWD's. The modulator could be implemented by using a programmable digital code generator driving a chip SAWD MSK generator.

4. PERFORMANCE COMPARISON

A probability of bit error performance comparison between the six modulation/demodulation approaches discussed above is shown in Figure 21. The chip rates in the channel are all the same. The comparison is made on the basis of probability of bit error (BER) versus the energy per bit to noise spectral density required to achieve a given error probability. For bit error probabilities in excess of 7×10^{-5}, coherent PSK or MSK is superior. For bit error probabilities less than 7×10^{-5}, 8'ARY MSK is superior. It requires .4 dB less energy per bit than coherent MSK at 10^{-5} BER.

Differentially coherent PSK and differentially coherent MSK have identical theoretical performance. At 10^{-5} BER their performance is 1 dB less than coherent MSK and 1.4 dB poorer than 8'ARY MSK. Pseudo-orthogonal (PO) MSK gives the poorest performance. Basically, this is because the correlation between the chip sequences is not positive N or negative N, but is either positive N or nearly zero. Hence, the signal detector space geometric separation between signals is not as great with the result of poorer performance. The performance of PO MSK is 3 dB worse than coherent DMSK as shown in the figure.

5. TRADEOFF PARAMETERS

Figure 1 contains some parameters which are important for deciding which modulation or signalling waveform approach is most desirable for the ARPA Packet Radio Network. The six modulation approaches are compared on the basis of performance. The best performance is denoted by 1 and the poorest performance is denoted by 5. These are qualitative indicators only.

The bandwidth utilization of the different modulation methods was also discussed above. The basic difference is between the PSK and MSK power spectrum for an equal chip rate. MSK has much lower out-of-band spectral energy and, hence, is superior where spectral containment is important. The PSK power spectrum rolls off at 20 dB per decade, whereas, the MSK spectrum rolls off at 40 dB decade. Quadri-phase PSK requires half the bandwidth of bi-phase PSK.

The timing requirements refer to the demands upon timing accuracy. Differentially coherent MSK and pseudo-orthogonal MSK have the least stringent timing requirements for small timing errors. 8'ARY MSK is more difficult because the slope of the correlation function per bit interval is greater. QPSK is superior for large timing errors. Bi-phase PSK is inferior for all timing errors. Coherent MSK requires the most accurate timing because one must sample the IF carrier in order to sense the sign of the data bit. To adequately recover the data from the coherent MSK modulated signal will require a phase lock loop.

The hardware limitations refer specifically to the device requirements relative to state-of-the-art technologies. For example, 10μsec SAWD's are relatively easy to make; whereas, 20μsec SAWD's are a little more difficult and 30μsec SAWD's require some extra development work. On this basis, coherent MSK is considered to have the fewest SAWD hardware limitations and 8'ARY MSK would have the most.

The judgments on the last three tradeoff parameters are more subjective and require more extensive investigation. The evaluation is based on first-pass engineering judgment.

The conclusions one might draw from such a tradeoff is that coherent MSK and differentially coherent MSK are the leading contenders. Coherent MSK has the problem of very accurate timing requirements since one must track the phase of the carrier for adequate detection. DC MSK circumvents this phase locking difficulty by coherently detecting two successive bit intervals. However, the price one pays is a slight performance degradation (≈1 dB). In actual practice, there will probably be .5 dB degradation due to phase tracking errors and jitter with coherent MSK.

Our recommendation at this time would be to assume differential coherent MSK as the baseline modulation approach.

TRADE OFF PARAMETERS / MODULATION TYPE	PERFORMANCE BER vs Eb/No @ 10^{-5} BER	BANDWIDTH UTILIZATION	TIMING REQUIREMENTS (BIT SYNCH.)	SAWD LIMITATIONS	EQUIPMENT COMPLEXITY	ESTIMATED COST	EQUIPMENT SIZE
DIFFERENTIALLY COHERENT BI-PHASE PSK	3	3	3	2	1	1	1
DIFFERENTIALLY COHERENT QUADRI-PHASE PSK	3	2	2	3	2	2	2
COHERENT MSK	2	1	5	1	4	3	4
DIFFERENTIALLY COHERENT MSK	3	1	1	2	1	1	1
PSEUDO-ORTHOGONAL MSK	4	1	1	2	3	3	3
8'ARY MSK	1	1	2	4	5	4	5

EVALUATION SCALE 1 - MOST DESIRABLE
TO
5 - LEAST DESIRABLE

FIGURE 1.

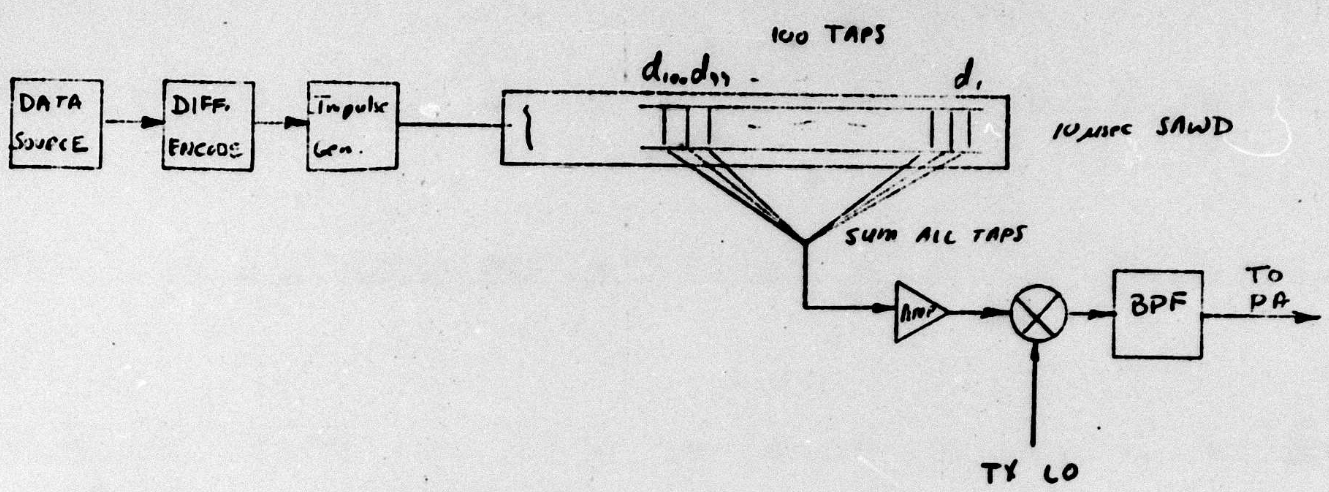


FIGURE 2 DPSK MODULATOR #1

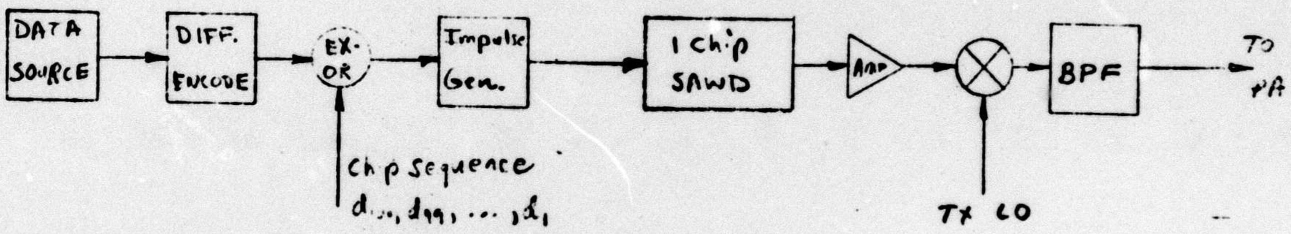


FIGURE 3 DPSK MODULATOR #2

W1

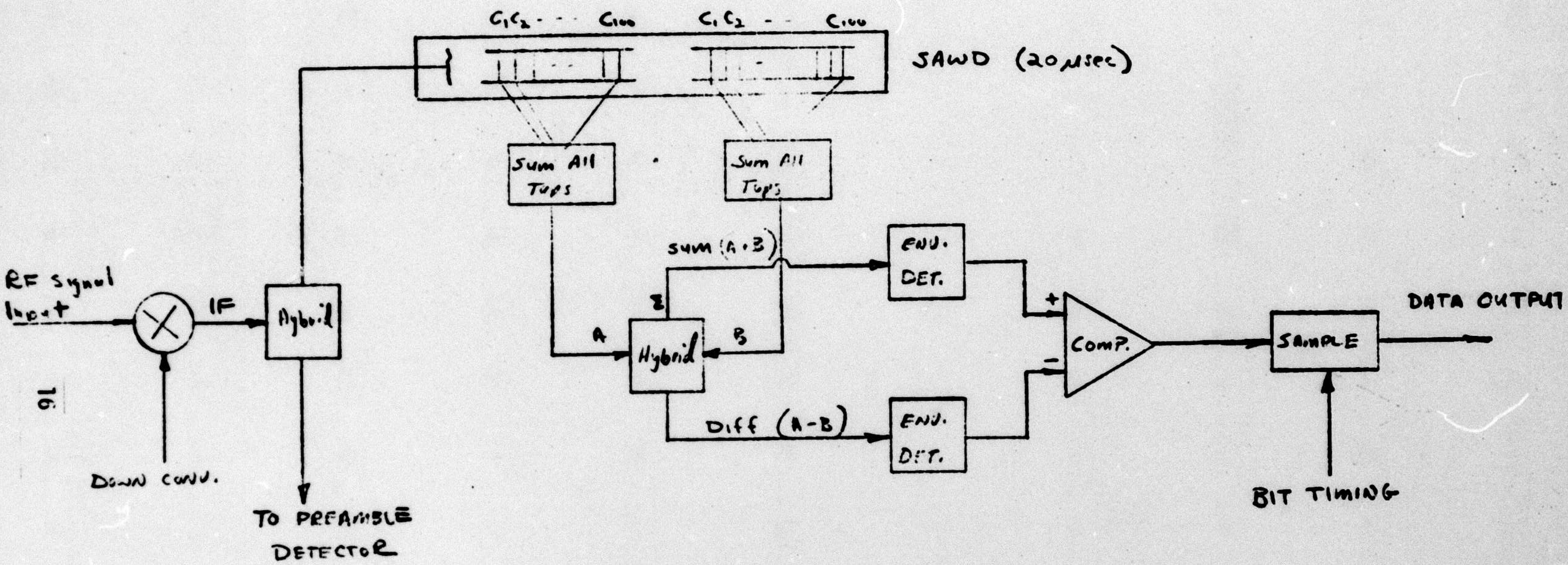
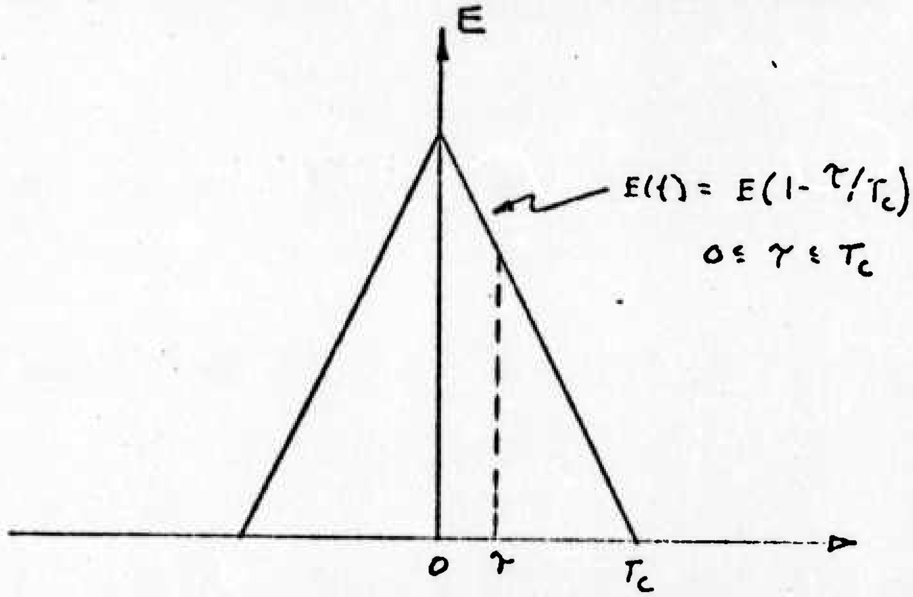


FIGURE 4. DIFFERENTIAL COHERENT PSK DEMODULATOR



SIGNAL AT OUTPUT OF ENVELOPE DETECTOR

LOSS FACTOR $= [1 - \tau/T_c]^2$

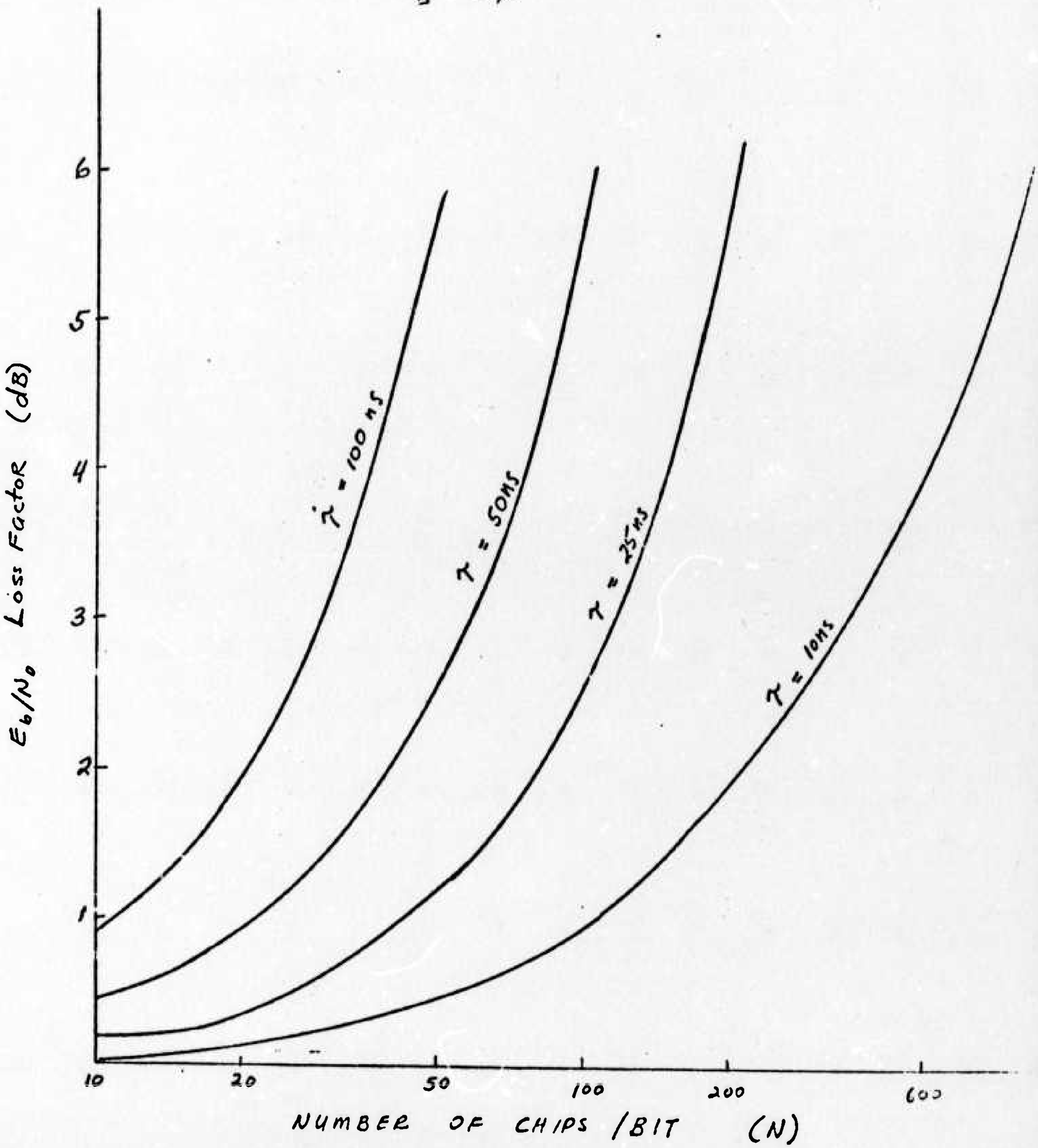
τ/T_c	LOSS FACTOR (dB)
0	0
.05	.45
.10	.92
.20	1.95
.30	3.10
.50	6.0

FIGURE 5. EFFECT OF TIMING OFFSETS TO MAXIMUM LIKELIHOOD DETECTION

FIGURE 6 E_b/N_o Performance Loss Due to Timing Offsets (τ)

MAXIMUM LIKELIHOOD DECODING

$T_s = 10\mu s$



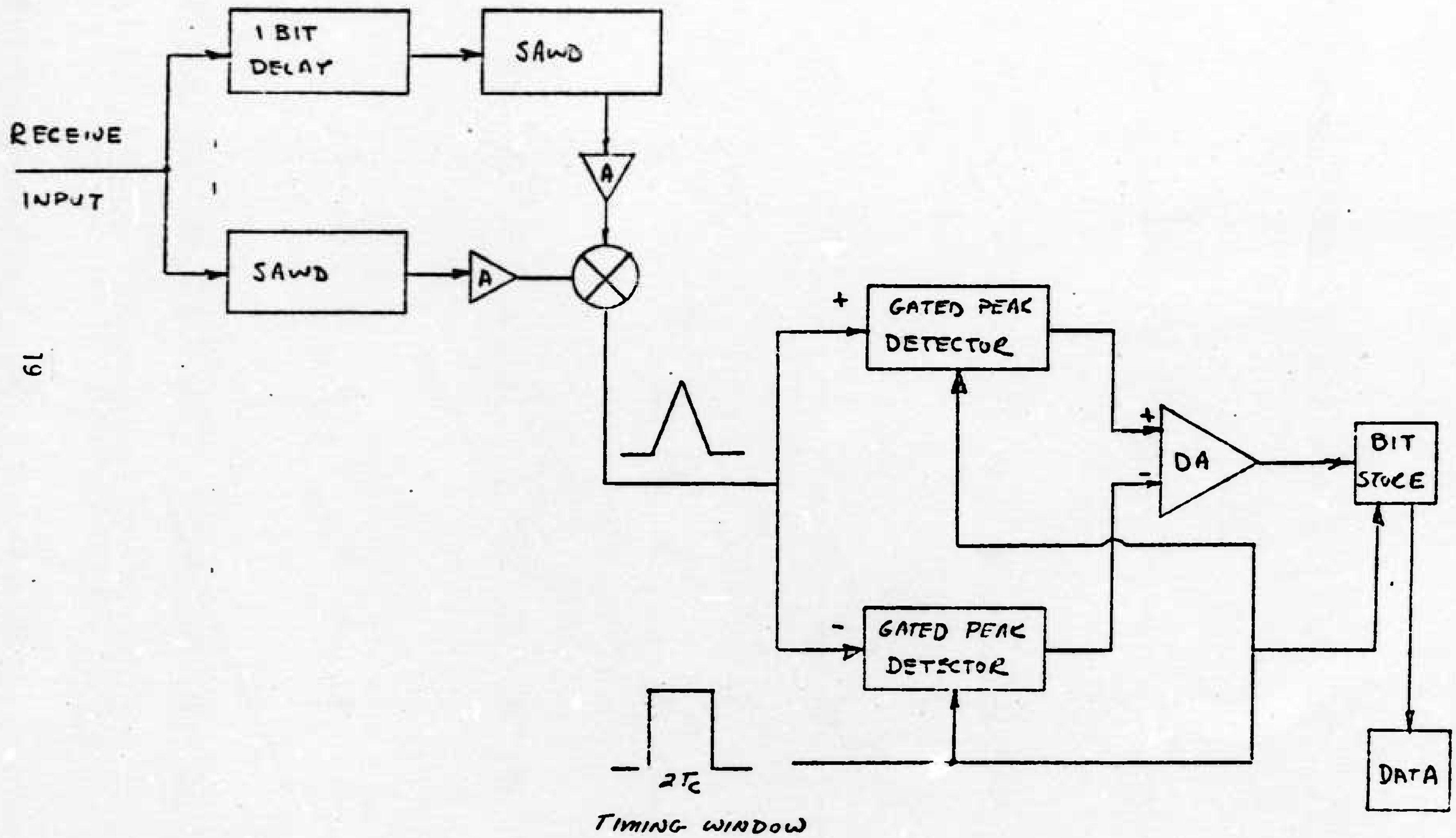


FIGURE 7 WINDOW PEAK SAMPLE & STORE DETECTOR

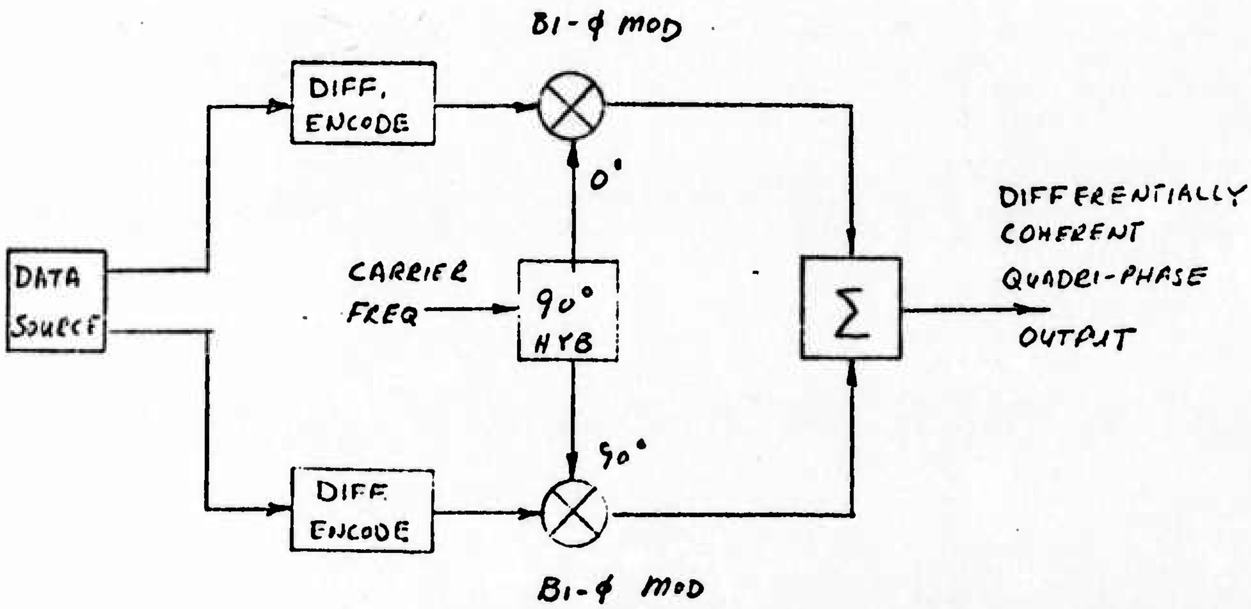
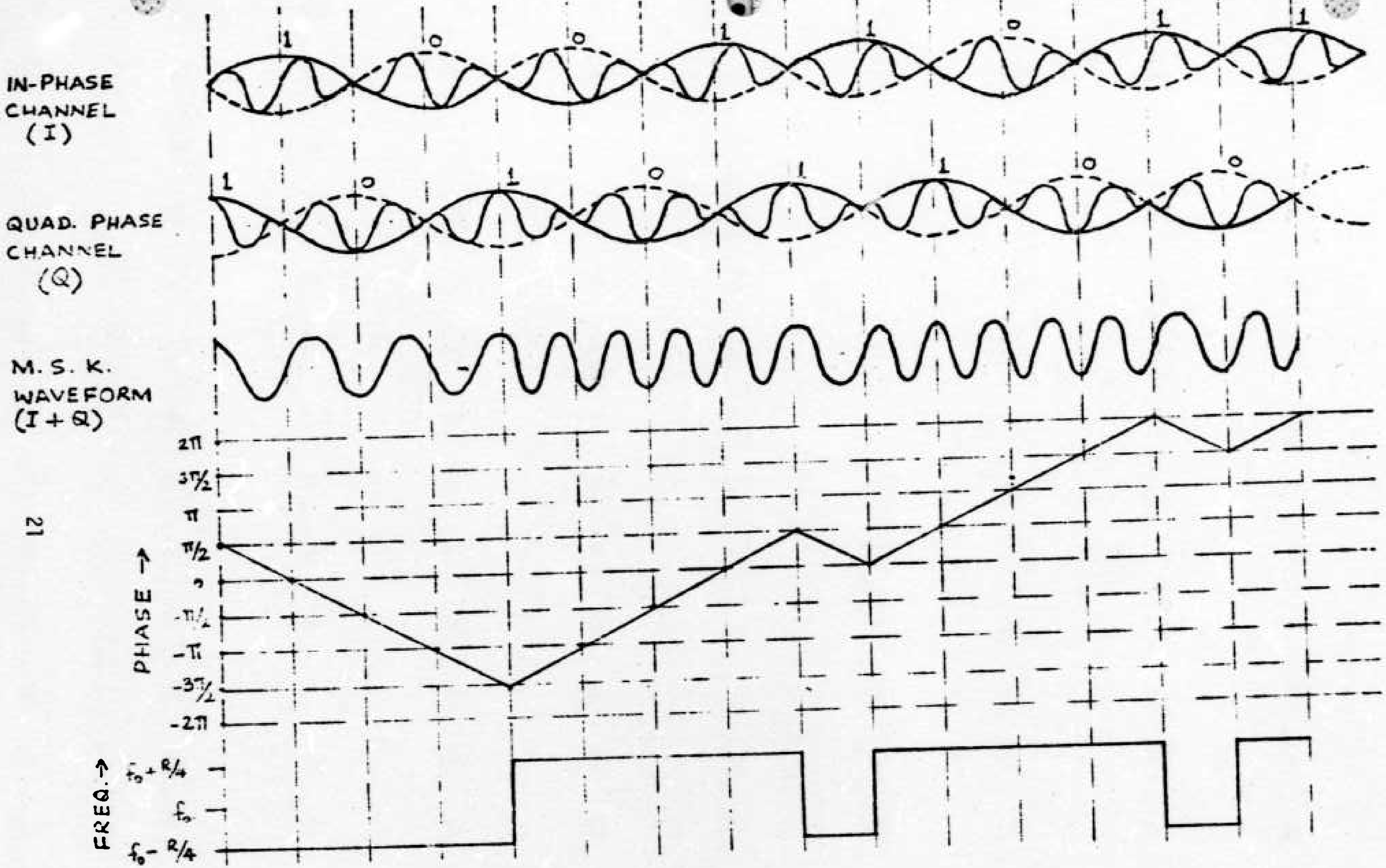


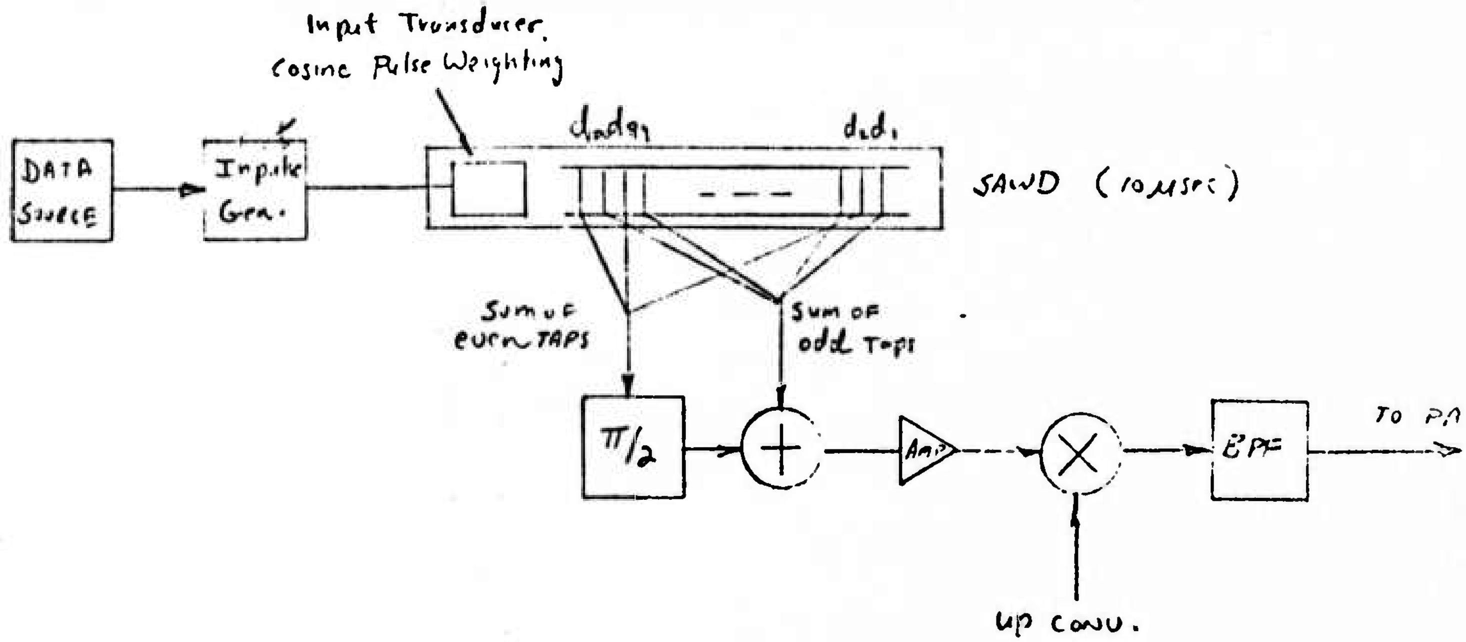
FIGURE 8 DC QUADRI-PHASE MODULATOR



IN-PHASE CHANNEL (I)
QUAD. PHASE CHANNEL (Q)
M. S. K. WAVEFORM (I + Q)
PHASE →
2π
3π/2
π
π/2
0
-π/2
-π
-3π/2
-2π
FREQ. →
f₀ + R/4
f₀
f₀ − R/4
1
0
0
1
1
0
1
1
1
0
1
0
1
1
0
0

FIGURE 9

M. S. K.



DATA BIT	OUTPUT CHIP SEQUENCE
1	$s_1 = d_{100}\, d_{99} \cdots d_1$
0	$s_2 = -d_{100}\, -d_{99} \cdots -d_1$

$R(s_1, s_2) = -1$

FIGURE 10 COHERENT MSK MODULATOR #1

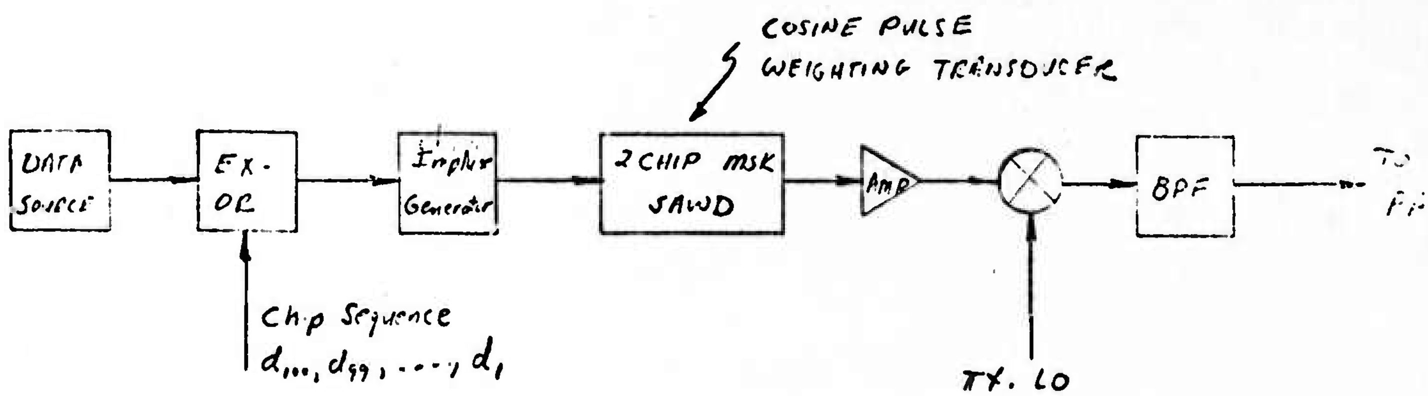


FIGURE 11 COHERENT MSK MODULATOR #2

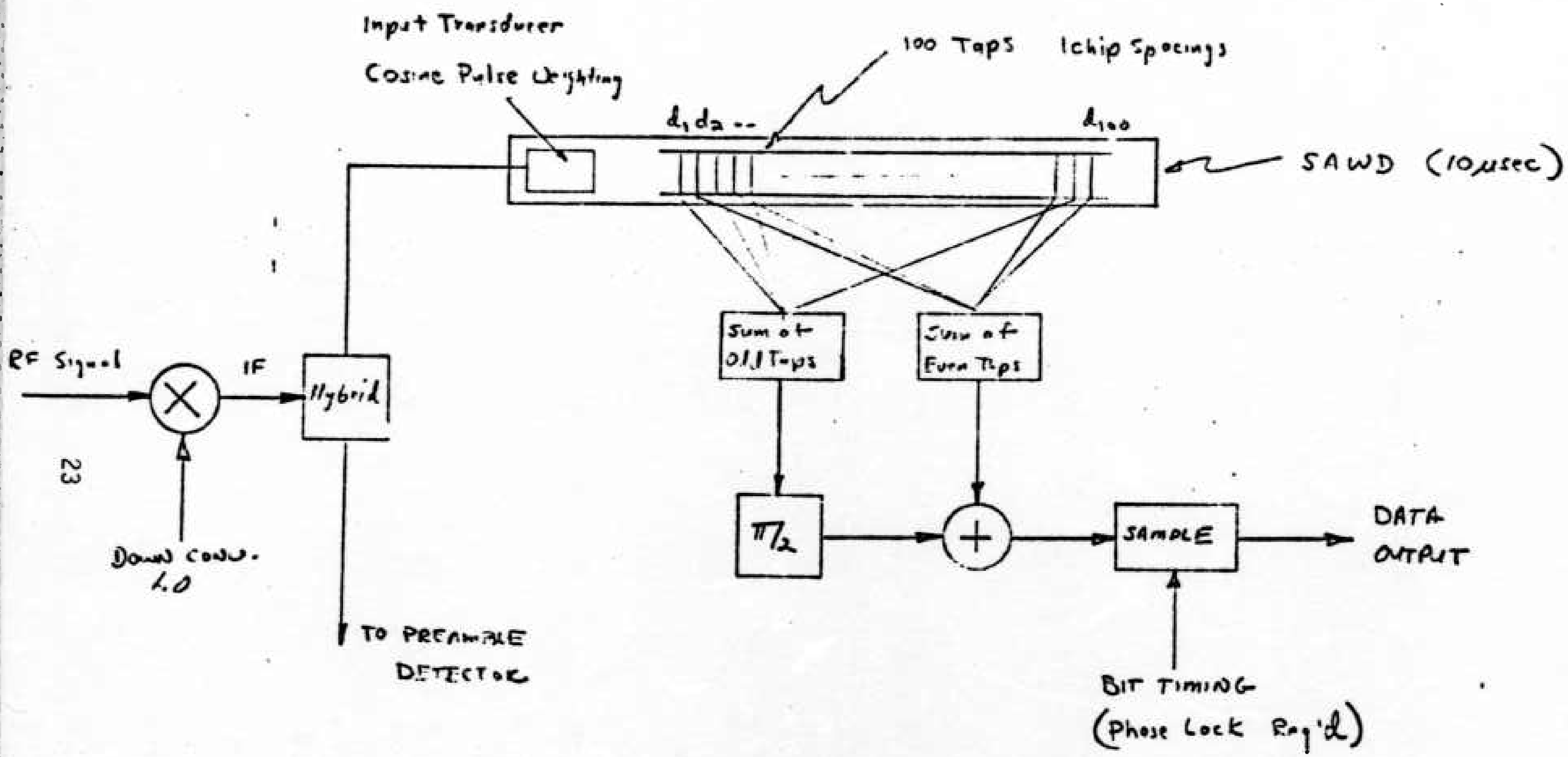


FIGURE 12 COHERENT MSK DEMODULATOR

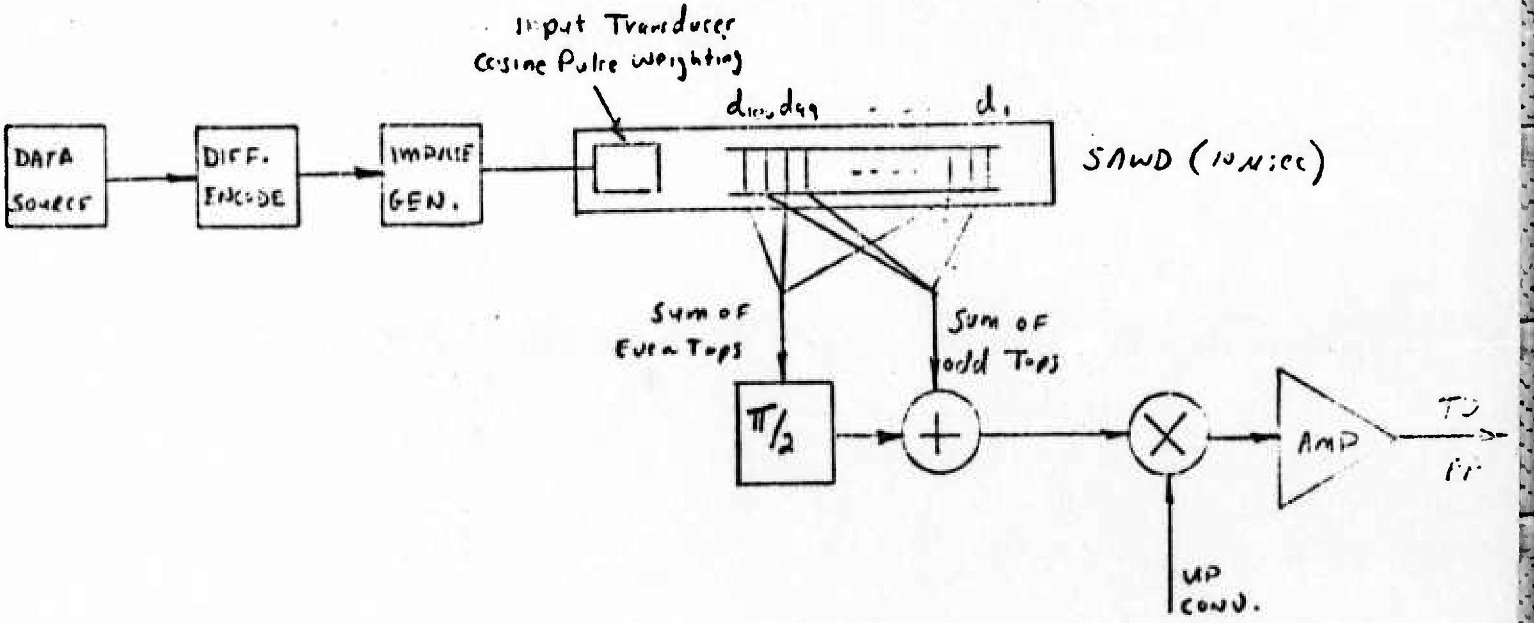


FIGURE 13 DIFFERENTIALLY COHERENT MSK MODULATOR

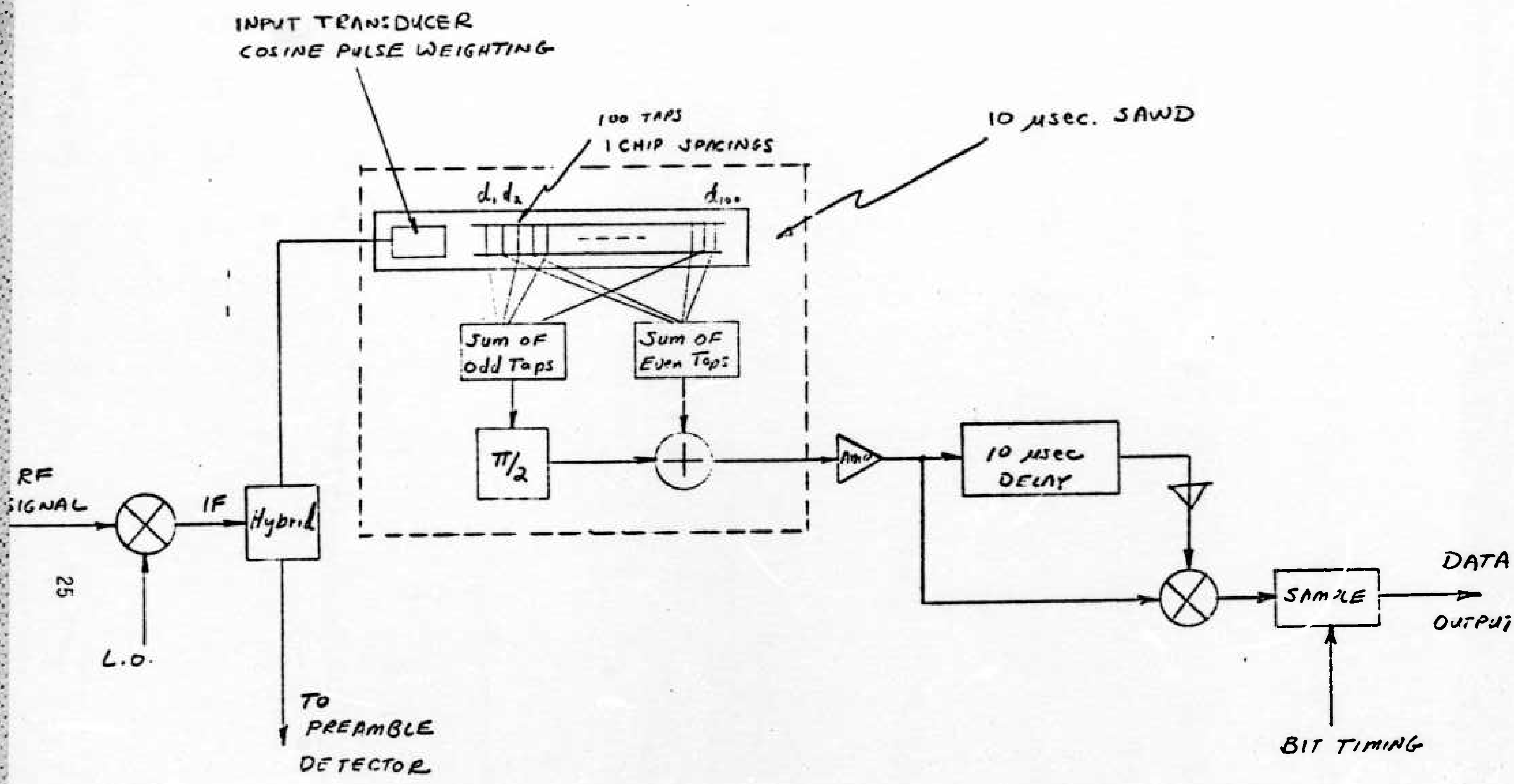


FIGURE 14A DIFFERENTIALLY COHERENT (DC) MSK DEMODULATOR

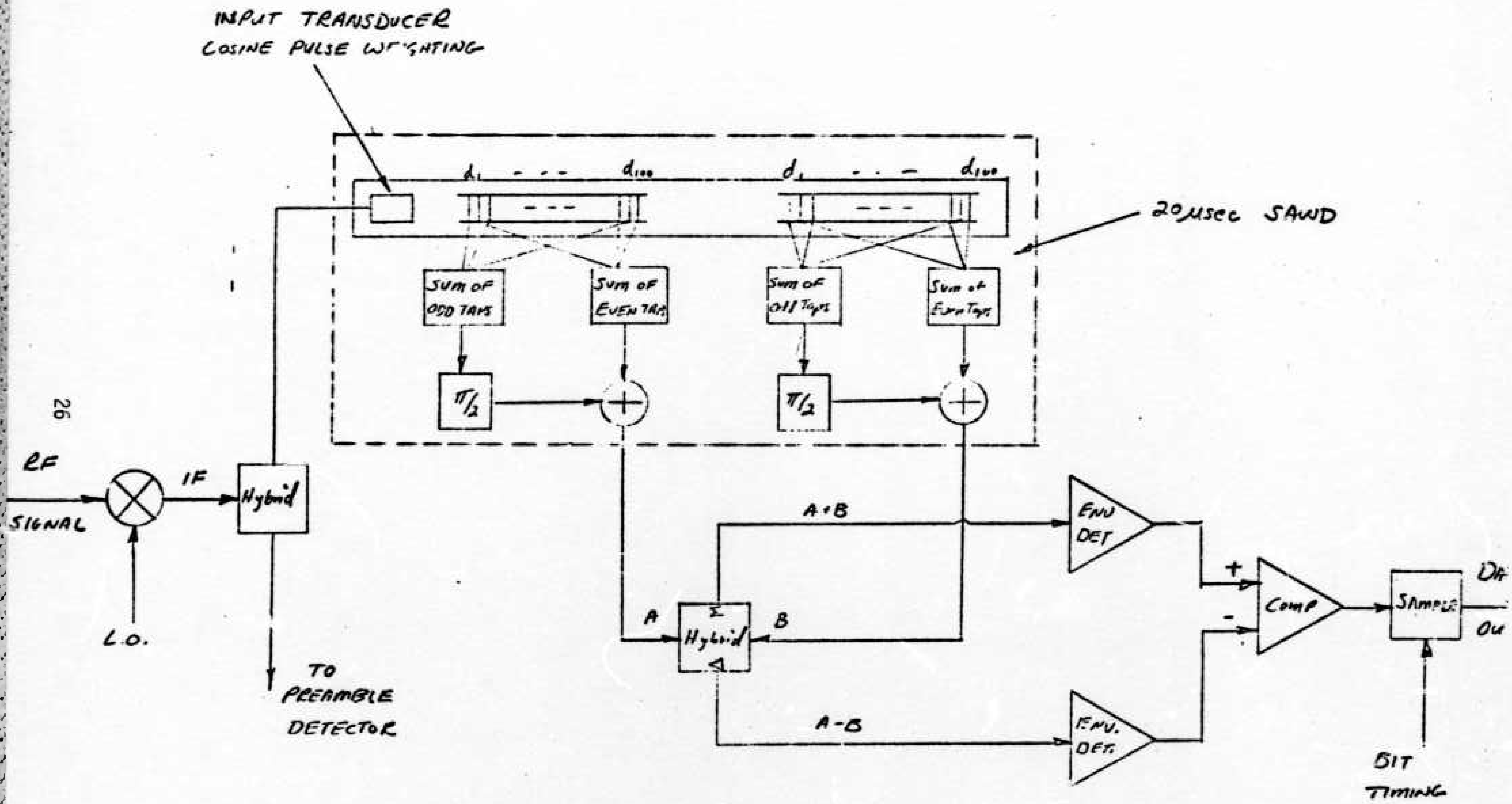
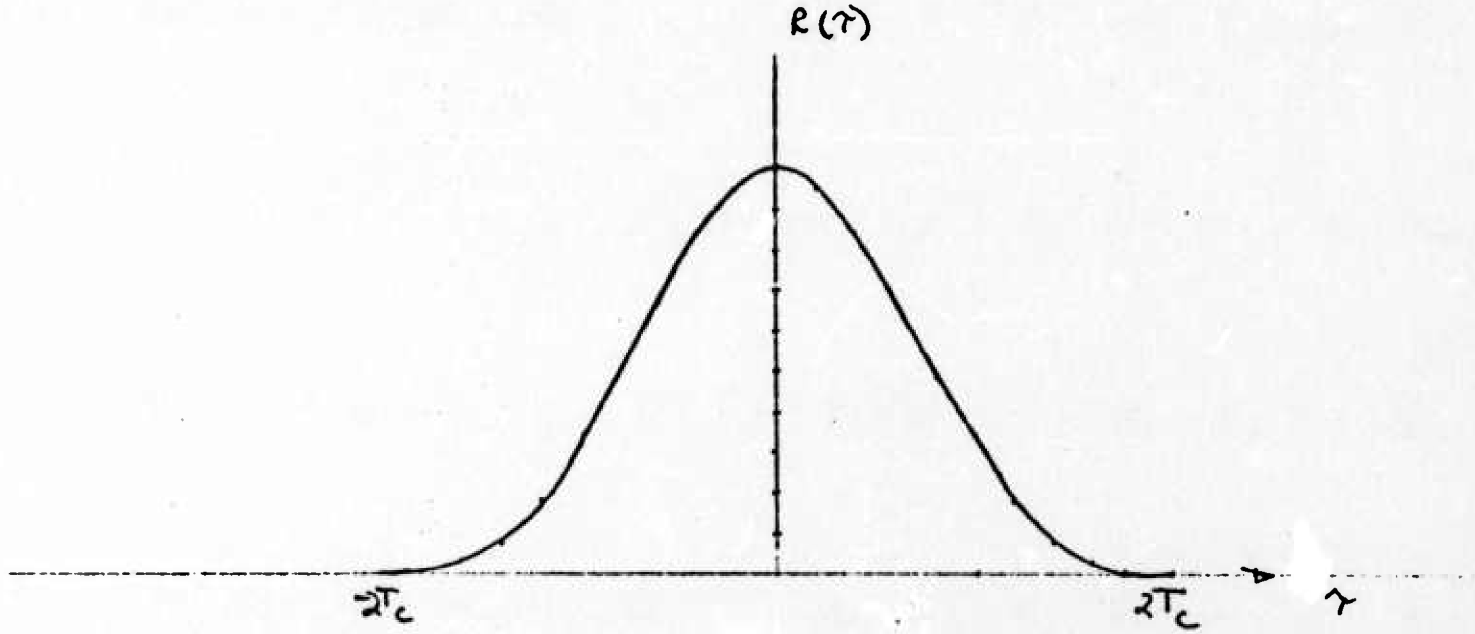


FIGURE 14B DIFFERENTIALLY COHERENT (DC) MSK DEMODULATOR

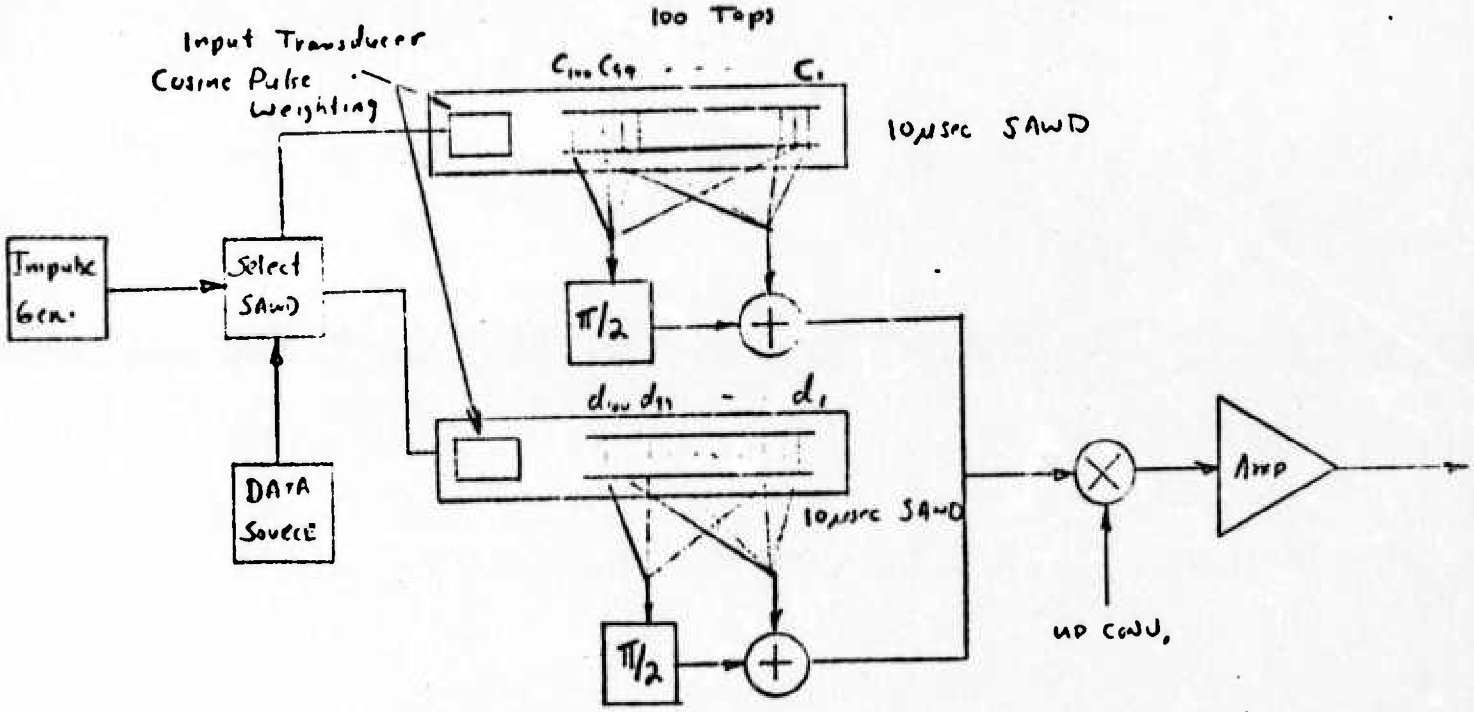
$$R(\tau) = \mathrm{Cosp}\,\frac{\pi t}{2T_c} * \mathrm{Cosp}\,\frac{\pi t}{2T_c}$$



SIGNAL AT OUTPUT OF MSK DETECTOR

$\tau/2T_c$	LOSS FACTOR (dB)	
0	0	IDEAL BIT TIMING
.05	.104	
.10	.41	
.20	1.58	
.30	3.49	
.40	6.23	
.50	9.95	

FIGURE 15 EFFECT OF TIMING OFFSETS FOR SAMPLING DETECTION OF MSK SIGNALLING



DATA BIT	OUTPUT CHIP SEQUENCE
1	$S_1 : C_{100}\, C_{99} \cdots C_1$
0	$S_2 : d_{100}\, d_{99} \cdots d_1$

$R(S_1 S_2) \approx 0$

FIGURE 16 PO MSK MODULATOR #1

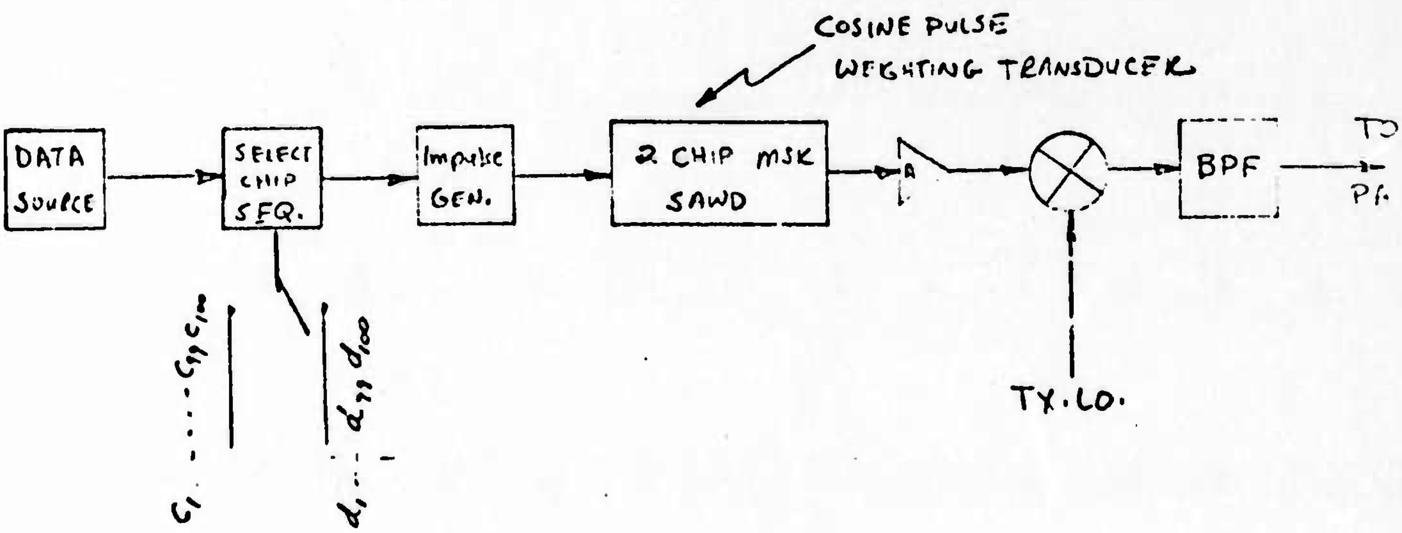


FIGURE 17 PO MSK MODULATOR #2

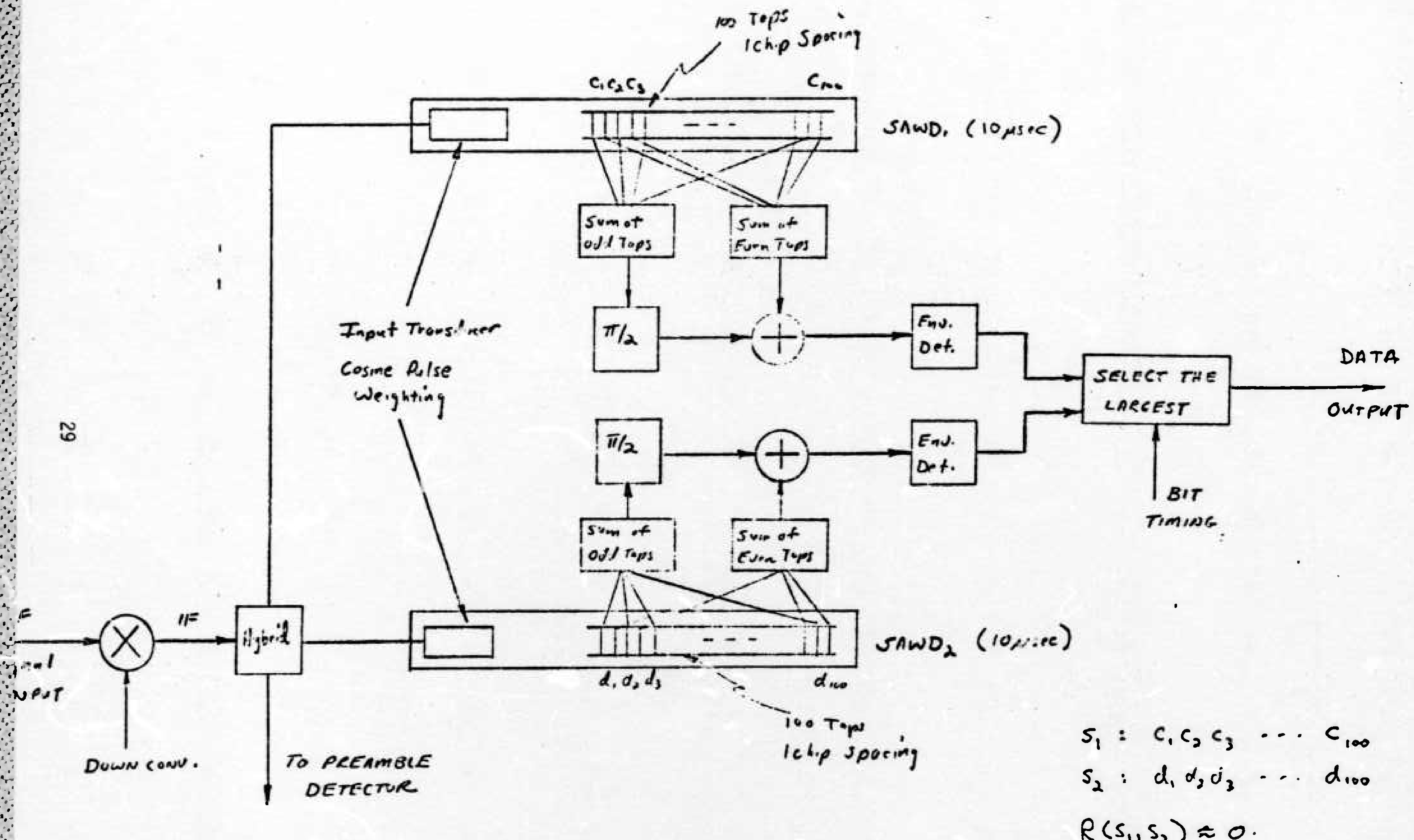
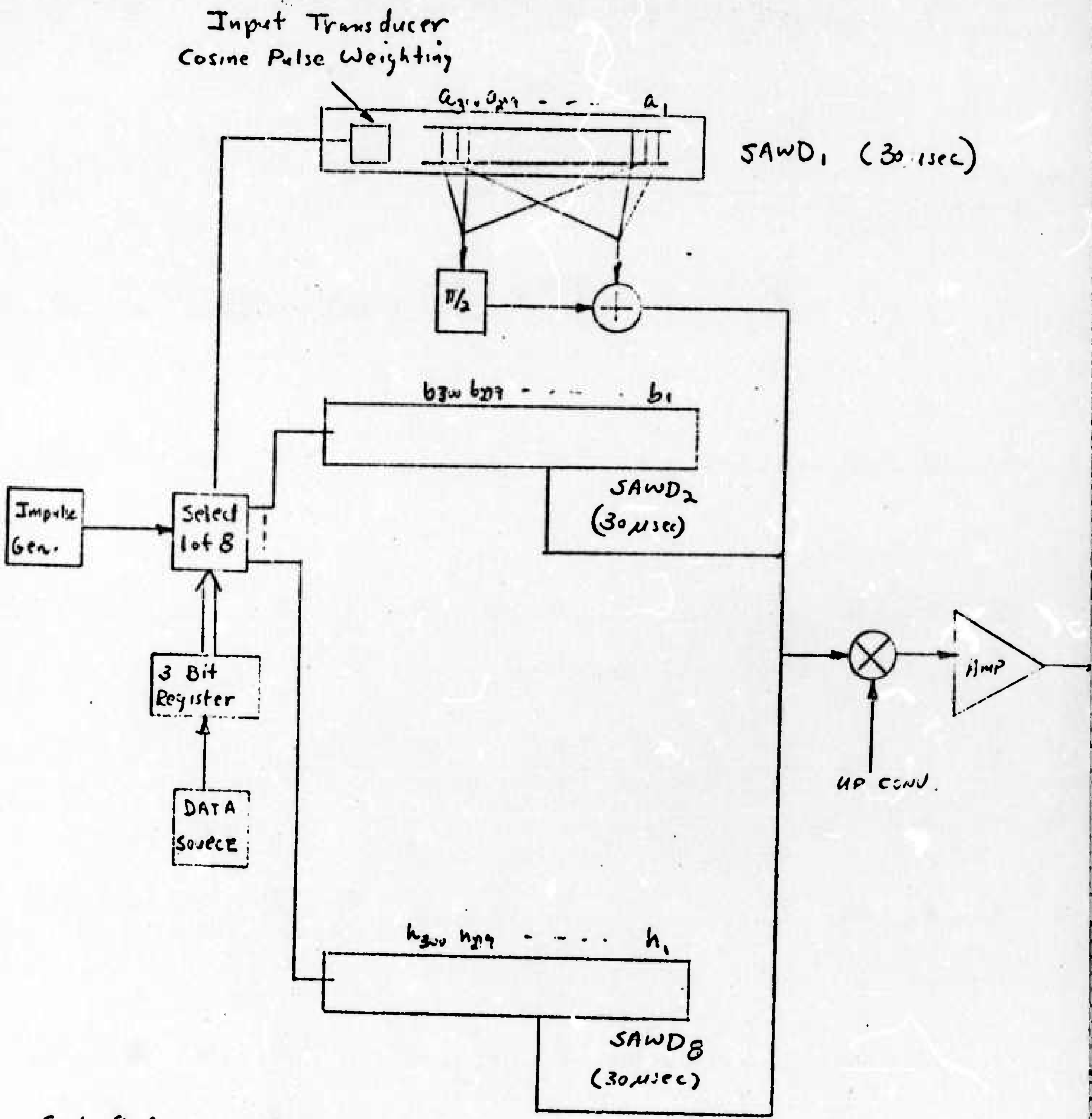


FIGURE 18 PO (Pseudo Orthogonal) MSK DEMODULATOR



$s_1 : a_1 a_2 \cdots a_{300}$

$s_2 : b_1 b_2 \cdots b_{300}$

$\vdots$

$s_8 : h_1 h_2 \cdots h_{300}$

$$R(s_i s_j) = \delta_{ij}$$

where $\delta_{ij} = \begin{cases} 0 & \text{if } i \neq j \\ 1 & \text{if } i = j \end{cases}$

FIGURE 17 8'ARY MSK MODULATOR

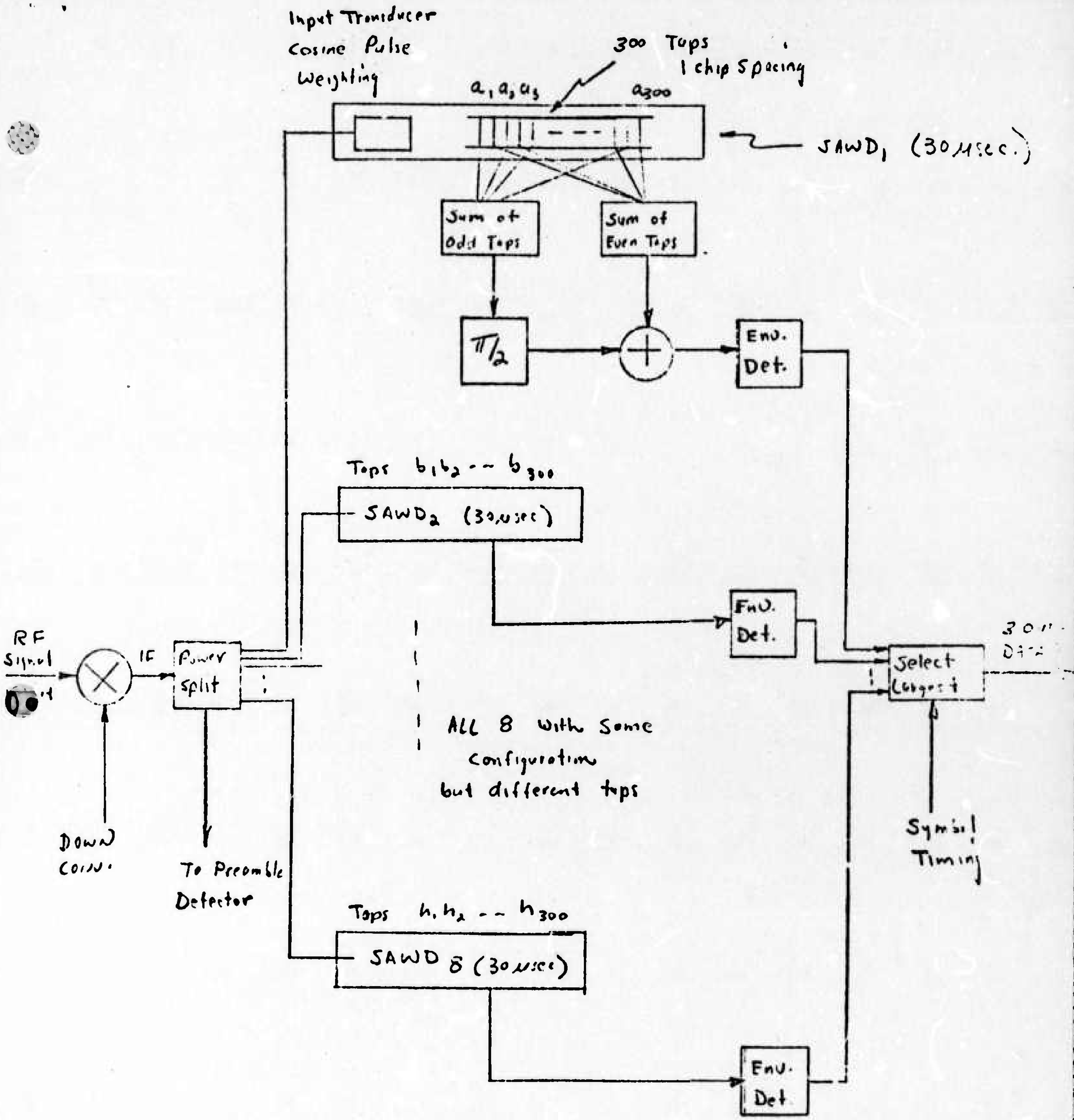


FIGURE 20 8'ARY MSK Demodulator

$S_1 : a_1 a_2 \cdots a_{300}$

$S_2 : b_1 b_2 \cdots b_{300}$

$\vdots$

$S_8 : h_1 h_2 \cdots h_{300}$

$R(S_i S_j) = \delta_{ij}$

where $\delta_{ij} = \begin{cases} 1 & \text{when } i = j \\ 0 & \end{cases}$

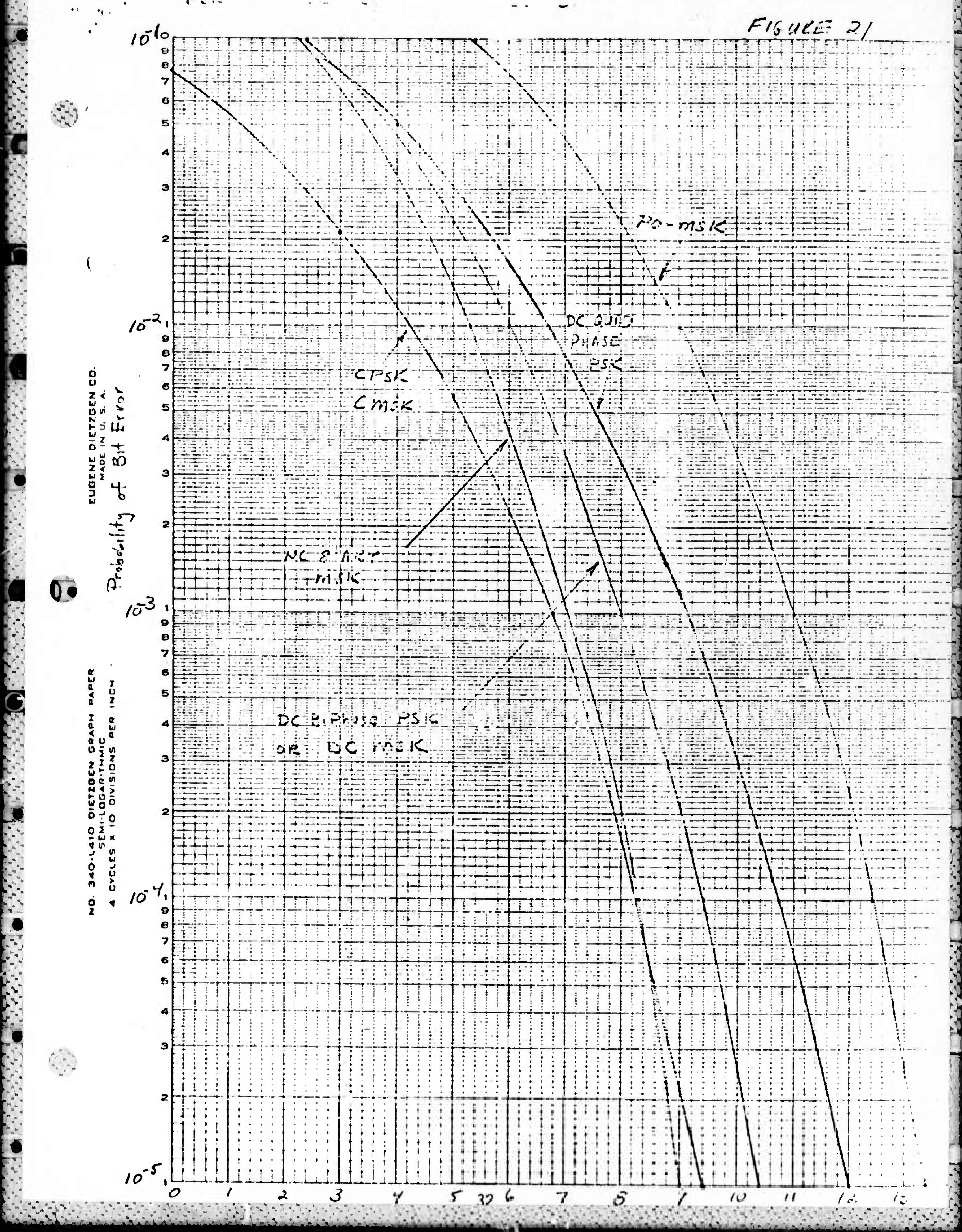
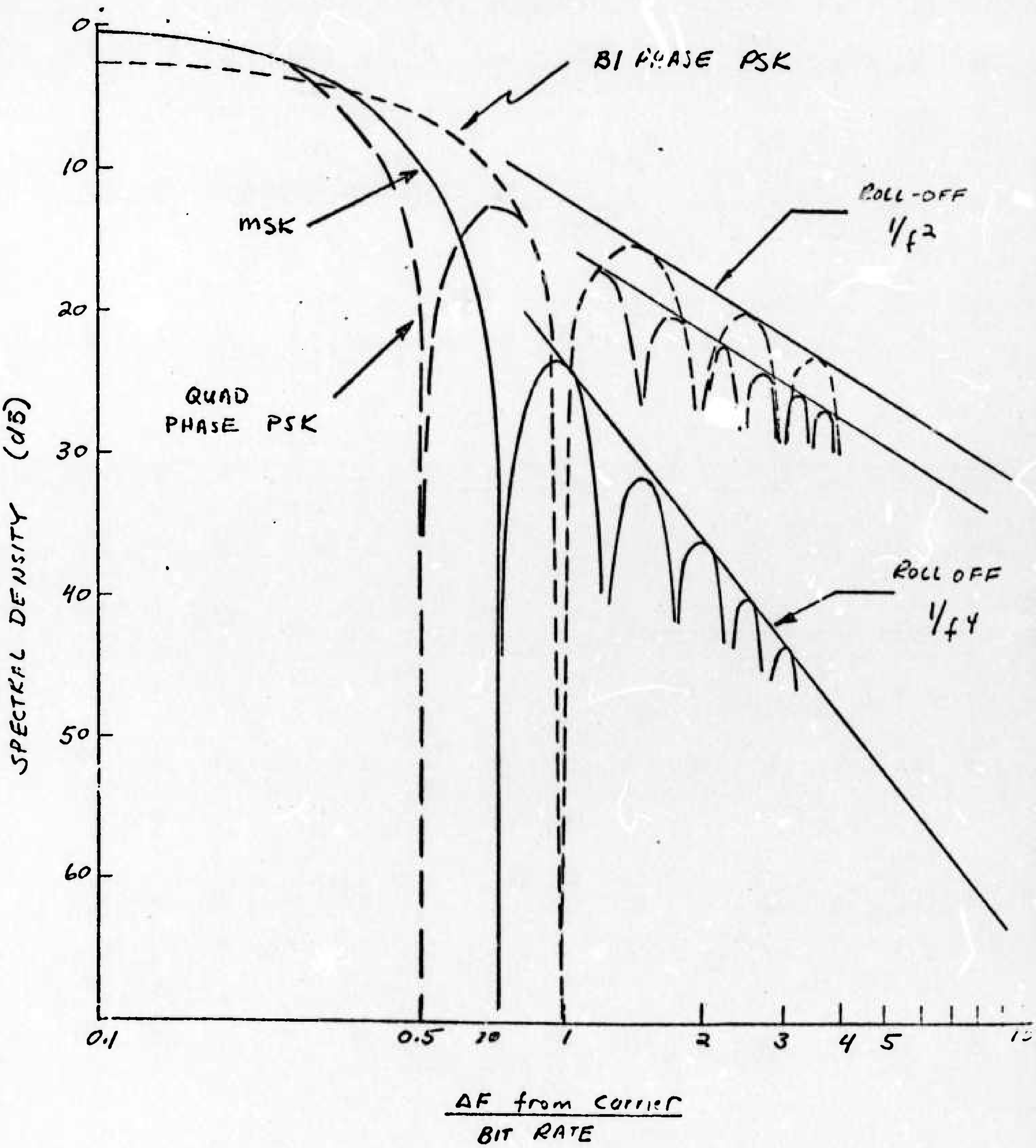


FIGURE 21
Probability of Bit Error
PO-MSK
DC QUAD PHASE PSK
CPSK
CMSK
NC & ART MSK
DC BiPhase PSK
OR DC MSK
EUGENE DIETZGEN CO.
MADE IN U. S. A.
NO. 340-L410 DIETZGEN GRAPH PAPER
SEMI-LOGARITHMIC
4 CYCLES X 10 DIVISIONS PER INCH

FIGURE 22 RELATIVE SPECTRAL DENSITY FOR MSK, QUAD PHASE PSK AND BI-PHASE PSK



Appendix C.2

Synchronization Preambles

1.0 INTRODUCTION

Transmission over the UHF links in the ARPA network will require packet synchronization at the receive and repeater terminals. By the utilization of surface acoustic wave devices (SAWD's), chip synchronization is not required. There are several approaches to the packet synchronization problem which can be utilized. These approaches are discussed in varying degrees of depth below, from which the recommended approach is selected.

Several assumptions are made in the evaluation of the candidate preamble schemes. Some of these assumptions are:

1. The modulation will be minimum shift keying (MSK).
2. Code spread spectrum pulse width of $T_C = T_B/N$, where N is the number of chips per bit, while allowing the radiated power to be constant.
3. Minimum data rate of the order of 100 kbps.
4. Bit error probability of 10^{-5}.
5. The processing gain will be in the vicinity of 10 to 20 dB.

Each of the approaches considered assumes that a bit is spread into 127 chips which make up a 127 chip maximum length sequence. Four preamble approaches are considered and compared. These approaches fall into two classifications: (1) Barker codes and (2) Repeat codes.

2.0 PREAMBLE SPECIFICATIONS

To begin a discussion of preamble detection, one must establish some preamble specifications. If we assume that a packet consists of 1,000 bits of data maximum, then a bit error probability of 10^{-5} results in a packet error rate of 10^{-2} without error detection and correction. Since error detection with automatic request for retransmission will be used rather than error correction, it would not be desirable that the packet error probability or retransmission probability exceed 10^{-2}. The preamble should have a detection probability at least as good and probably better than the packet detection probability. Therefore, the design goal for the probability of preamble detection will be 99.9%.

A frequently used relationship between false alarm probability and average time between false alarms is given by:

$$P_{FA} = \frac{1}{T_{FA}B}$$

where P_{FA} is the false alarm probability.

T_{FA} is the average time between false alarms.

B is the information bandwidth = 100 kHz.

Assume that a packet consists of 1,000 bits. Thus, a packet duration is 10 milliseconds. If we assume one false alarm for every 1,000 packet times, i.e., every 10 seconds, then $P_{FA} = 10^{-6}$. This will be the false alarm probability design goal.

For differentially coherent PSK on MSK modulation a 10.4 dB signal-to-noise ratio (E_b/N_o) is required to achieve a bit error rate (BER) of 10^{-5}. For this probability of bit error, 1% of the 1,000 bit packets will be in error. However, a reasonable approach is to assume a 50% probability of packet error in order to come up with the worst case preamble design. The 1,000 bit packet has a 50% probability of error for a BER = 7×10^{-4} corresponding to E_b/N_o = 8.2 dB. A 200 bit packet has a 50% chance of being correctly received for BER = 4×10^{-3} which corresponds to 6.5 dB. Since the preamble length is the same for either packet size, the design goal will be to provide a preamble having a 99.9% probability of detection in a worst case E_b/N_o environment of 6.5 dB. This margin can also be applied to some system degradations such as timing errors, frequency offset, doppler, and sidelobes.

3.0 PREAMBLE ANALYSIS

3.1 Coherent Combining

The required preamble length and E_b/N_o to achieve the probability of detection of .999 and probability of false alarm of 10^{-6} is a function of the detection technique. Coherent combining of the preamble yields the best results. The probability of missing the preamble versus the signal-to-noise ratio for a bit is plotted in Figure 1 with the number of bits in the preamble as a parameter.

Also, the false preamble probability os 10^{-6}. The results show that 8 bits are required to achieve the design goals. The analysis deriving these results is given in Appendix A.

The coherent combining of the bits at the output of the SAWD is difficult. The combining accuracy is a function of the IF frequency and must be less than 1 nanosecond. The delay can be achieved with a surface wave device matched to the MSK SAWD.

3.2 Non-Coherent Combining

If the preamble bits are combined non-coherently, additional bits are required to achieve the same design goals. This is demonstrated with Figure 2 where the number of preamble bits required for a missed preamble probability of 10^{-3} is plotted versus the signal-to-noise ratio. The curves are shown for two values of probability of false alarm 10^{-3} and 10^{-6}. To achieve the design goals stated above requires a minimum of 13 bits of preamble. The derivation of these curves is given in Appendix B. In order for the analysis to be tractable some simplifying assumptions were made. For low values of BT the results are a few tenths of a dB pessimistic.

3.3 Post Detection

A detection decision is made on each bit for the third approach. To exceed the threshold for detecting the preamble requires that M out of N bits are received correctly. In one case considered, a Barker code is used to achieve both bit timing and the sync preamble. The 13 bit Barker code is followed by another 13 bit Barker code where each bit of the second code is reversed in sign. For 26 bits of preamble 21 or more must agree with the stored reference at the receiver to achieve the design goals. The results are plotted in Figure 3. The derivation of these curves is given in Appendix C.

4.0 PREAMBLE APPROACHES

4.1 Introduction

Four different preamble approaches are discussed in this section. Some of the advantages and disadvantages of each approach will be given. There are many variations and implementation approaches which can be suggested. In the interest of getting at the underlying principles, just a few of the possibilities are shown.

For each of the preamble approaches considered each bit consists of 126 or 127 chips, the exact number depending upon the code used to select the chips. The basic preamble principles hold independent of the number of assumed chips per bit.

4.2 Barker Preamble With Coherent Detection (Coherent Combining)

One of the most classic approaches is the use of the Barker sequences for preamble detection. This family of sequences is characterized by:

$$R(k) = \begin{cases} N & k = 0 \\ \pm\ 1,0 & k \neq 0 \end{cases}$$

where R (k) is the autocorrelation function and N is the number of bits comprising the sequence. The size of the family is restricted since there are no more than nine known Barker sequences. The Barker code coming closest to satisfying the requirements stated above is of length N = 7. The next largest Barker sequence is of length N = 11. For N = 7 its sequence is given by:

+++--+-

This sequence gives a ratio of 16.9 dB between the peak and the sidelobes. A plot of the autocorrelation of the Barker sequence illustrating this is given in Figure 4. Each bit of the Barker sequence consists of 127 chips from a 127 chip maximum length shift register. If the Barker bit is positive, the 127 chip sequence is transmitted. If the Barker bit is negative, each chip of the 127 chip sequence is inverted. A possible preamble detector implementation is shown in Figure 5. Each SAWD is matched to one bit of the Barker code. The first SAWD has the cosine pulse inter-digital transducer in addition to the 127 taps corresponding to the 127 chips. For further details on possible SAWD implementations of MSK detectors, see Appendix C.1. The outputs of the 7 SAWD's are summed coherently, amplified and compared to a threshold. A level sensing circuit establishes the input signal level and normalizes the output of the video amplifier so that a fixed threshold V_T can be used. When this threshold is exceeded, the presence of the preamble is indicated. A pulse is triggered at the time the threshold V_T is first exceeded. The pulse is terminated when the signal fells below the threshold. The center of this pulse is then estimated which establishes the bit timing for the remainder of the packet. The preamble accomplishes two functions: (1) It establishes bit timing and, (2) it indicates the start of the message. A discussion of the accuracy to which the bit timing can be estimated is given in Temporary Note #33. The following expression was derived for the time of arrival variance for a practical center pulse estimator:

$$\sigma_T^2 = \frac{\gamma^2}{2\ (2\ E/N_0)}$$

where E/N_0 is the signal to noise ratio and γ in the chip duration. For γ = 79 nanoseconds and E_b/N_0 = 6.5 dB we get σ_T = 18.6 ns.

From the autocorrelation function of the preamble (Figure 4) it can be seen that multipath which is delayed by 100 ns relative to the received signal will be rejected. Since coherent MSK detection is used in this example there is a 180° phase ambiguity of the chips and bits. Therefore, the received cross-correlation pulse at the output of the SAWD will be either positive or negative. This presents no problem as long as we are not detecting data since the output of the envelope detector is always positive.

As noted above, the peak-to-sidelobe ratio is 16.9 dB (7:1). The sidelobes, in effect, add to noise, thereby reducing the effective signal/noise ratio by 1/7th (1.34 dB).

One practical shortcoming is in the implementation of this approach as it requires seven 10μsec SAWD's in cascade, each of which has from 10 to 30 dB insertion loss. Longer SAWD's can be used with more taps on each SAWD. The practical limitation with today's technology is about 40μsec.

With the use of the 7 bit Barker code described above, the design goals are missed only by 0.5 dB (the design goal required 8 bits). However, the next shortest Barker code is 11 bits long.

Another disadvantage is that the effects of doppler shift and frequency errors are cumulative through the entire preamble. Thus, the coherent technique is much more sensitive to such offsets than a non-coherent combining technique.

One feature of this technique is that the first two SAWD's could be used for detecting differentially coded MSK data signals. Also, the derivation of bit timing and the end of preamble or start of message indication do not have to be derived separately.

4.3 Barker Preamble With Differentially Coherent Detection (Coherent Combining)

Another Barker preamble approach is shown in Figure 6. This approach requires only two SAWD's while yielding performance close to that of the optimum approach discussed above. Differential MSK modulation and detection are employed. The details of the modulation technique are discussed more completely in Appendix C.1 on modulation approaches.

The MSK SAWD's are 1 bit time long or 10 μsec. In essence they are filters matched to the spread signal on a bit basis. They remove the pseudo-noise spreading. A baseband preamble is obtained at the output of the comparator. Although a second analog matched filter matched to the 7 bit Barker sequence would be optimal, a close approximation can be obtained by limiting the baseband signal as shown and sampling at a high rate relative to the bit rate. Ten samples per chip would be realistic. This 8890 (10 x 127 x 7) length sequence of samples is then correlated in a digital matched filter with the stored sequence. It may be possible to implement this operation with a high speed ROM. The summer consists of counting the pairs of bits which are alike. When the digital counter exceeds a predetermined threshold value the preamble presence is established. Bit timing can also be derived in a manner similar to that discussed above.

The advantage of this approach is the reduction in the number of SAWD's. Another advantage is that since the bits are encoded differentially the bit detection circuitry can be used to detect subsequent message data bits without the need for separate data detection circuitry.

A disadvantage is that there is some loss of performance. A detailed analysis of the degradations due to the single bit quantization has not been done but based on experience it is estimated that this degradation may be in the vicinity of 1 to 2 dB depending upon the number of samples per bit. Added to this is the 1.34 dB degradation due to sidelobes. An additional 2 dB of performance can be gained by extending the Barker preamble to 11 bits.

Another disadvantage is the rate of sampling required in order to adequately capture the correlation peaks and achieve a good estimate of bit timing. Also, the digital filter of 8890 samples operating at 100 MHz is quite complex, hardware-wise, to implement.

4.4 Repeat Preamble With End of Preamble Detector (Coherent Combining)

The third candidate preamble technique is shown in Figure 7. The preamble consisting of n chips from a maximum length sequence is repeated 7 times. The MSK SAWD detects one bit at a time. The SAWD is matched to the n chip sequence. The output of each detected bit is coherently added to the previous outputs of detected bits. At point A in the figure the envelope of $R(\tau)$ builds up in time as shown in Figure 8. A pulse is triggered at the time V_T, is first exceeded. The pulse is terminated when the signal falls below the threshold. The center of this pulse is then estimated which becomes the bit timing pulse for the remainder of the packet.

The second part of the preamble is the four bit Barker sequence. Its puspose is to establish the end of the preamble and the beginning of the message. Each bit of the Barker code consists of n chips from a maximum length PN sequence. A logic "one" takes the sequences as is and a logic "zero" inverts the sequence. A 40μsec SAWD may be used to match to the Barker sequence. An alternative is to use four 10μsec SAWDS similar to that shown in Figure 5. A threshold V_{T2} is used to indicate the presence or absence of the code at the sample time. If the output exceeds the threshold V_{T2} at sample time, then the end of preamble is indicated and data detection commences.

This approach is relatively straightforward. It requires a few extra preamble bits because the detection process is divided into two parts.

4.5 M Out of N Preamble Detector (Post Detection Combining)

An M out of N preamble detector, where N is the number of bits in the preamble and M is the number of bits which must match to detect the preamble, is shown in Figure 9A. The preamble length N is 26 bits. The preamble consists of a 13 bit Barker code followed by an inverted 13 bit Barker code. The Barker code and its inverse is given by.

+++++--++-+-+   Barker

-----++--+-+- $\overline{\text{Barker}}$

The MSK detection process can be implemented as shown using a SAWD. The details of this implementation approach are discussed in the Modulation Waveform paper, Appendix C.1.

In this preamble technique the bit integration occurs after the bit decisions are made. At the detector output, point A of Figure 8, the waveforms are correlation pulses of 4 chips duration (316ms) occuring at the bit rate (10μsec intervals). The pulses are either positive or negative. A positive pulse exceeding the threshold yields an output pulse from the upper comparator (C). This pulse is shifted asynchronously into the register. An initial estimate of bit timing is also derived from this pulse which is used to clock the register at subsequent bit intervals. If the pulse at the next bit sampling interval is positive and exceeds the threshold, it is shifted into the register. If the pulse is negative an output is obtained from the lower comparator. At the bit sampling interval this pulse is shifted into the register as a logic "0". The matched filter consists of a shift register and taps with each tap weighted by either a "one" or a "zero" in accord with the Barker code. One digital implementation approach is shown in Figure 9B. The two sums compute the number of matches and mismatches between the incoming data and the locally stored code sequences. The difference between the two outputs is compared against a threshold. When the threshold is exceeded an end of preamble (EOP) is obtained. If the register contains all zeroes and the 26 bit $\overline{\text{Barker}}$, Barker code is received, the difference between the two summers of the digital matched filter (point P on Figure 9) follows the correlation function shown in Figure 10. The correlations are separated by the bit interval. Without any errors the output is always negative or zero until the preamble lines up exactly in the register. At this time the output is 26 units.

The expected performance obtained with this preamble approach is calculated in Appendix C. The results are plotted in Figure 3. The probability of a false alarm and the probability of a false preamble are plotted as a function of the threshold. In this analysis two thresholds are involved. The first one is used for the detection of the bits and the second for the detection of the preamble. The probability of a false alarm is not a function of the signal-to-noise ratio. The probability of a missed preamble is a function of the signal-to-noise ratio and is plotted for an SNR of 7.4 dB and 6.5 dB. If the preamble is set to 21, the probability of a false alarm is 10^{-7} and the probability of a missed preamble is 10^{-3} for a 6.5 dB SNR. This meets our design objectives.

One of the disadvantages of this approach is the length of the required preamble when compared with the approaches discussed previously.

An advantage is the relatively simple implementation requirement. The basic speed of the detector is at the bit rate with the exception of a counter in the center of pulse estimator which will operate from 5 to 10 times the chip rate. The basic SAWD detector is used for detecting both the preamble bits and the subsequent data bits. The bit timing sampling pulse is derived simultaneously with the receipt of the preamble. A continuously improved bit timing estimate during the preamble enables a continuous improvement in the performance of the detected preamble bits.

4.6 Alternate Bit Timing Recovery Approach

Another way of acquiring bit synchronization which can be applied to any of the above preamble approaches is by using a conventional analog phase-lock-loop. A VCO is phase-locked to the incoming envelope detected correlation peaks which occur at the bit rate. The correlation pulses sample the phase of the local 100 kHz reference and hold the sample value. This sample value is filtered and the output is the control voltage for the VCO. After the VCO is locked to the correlation peaks, its timing is transferred to a counter derived from a stable clock. A method of implementing this technique is shown in Figure 11.

For this phase lock-loop, Hoffman's** analysis for transient response is shown in the table below. A loop bandwidth (ω_n) of $2\pi 16$ kHz is assumed for a phase-lock loop with 100 kHz reference frequency. This assumes AGC has been set and no noise is present.

** "Phase-Lock Techniques" by F. M. Gardner, p. 34.

For noise, no analysis is available to calculate transient time. For S/N 10 dB, doubling the number of bits required seems appropriate; i.e., up to 22 bits might be required. AGC is assumed before phase locking. Of course, the phase-locking acquisition can occur before AGC is finally settled out.

Ten (10) ns of phase error corresponds to .4dB degradation point for ideal sampling with timing offset for maximum likelihood detection. One-hundred (100) ns corresponds to $2T_C$ windows peak and store detection worst case where .5 dB degradation is seen. A 10 ns phase error requirement also places a stringent requirement on DC phase drift of the phase detector and counter circuits used. Strobing a counter with 10 ns increments requires clocks and counters operating in excess of 100 MHz.

Ten (10) ns requirements on timing is possible, but difficult, either with phase-locking or with SAWD's matched for preamble timing and detection. Therefore, if this approach is to be used, the window detecting method is recommended.

5.0 COMPARISON OF PREAMBLE APPROACHES

Figure 12 contains some parameters which are important for deciding which preamble approach is most desirable for the ARPA Packet Radio Network. The four preamble approaches are compared on the basis of performance and equipment complexity. Based on the probability design goals, the length of the preamble is different for each approach operating in the same signal-to-noise environment. The best performance is denoted by 1 and the poorest performance by 4. One should be cautioned that there may be other system factors which were not considered in this paper that may influence the selection of the length of the preamble.

Several approaches are possible for deriving bit timing. One approach uses a bit timing loop as discussed in 4.6. Another uses a time of arrival estimator. The objective is to estimate the time of occurrence of the correlation peak. Potentially the bit timing loop yields superior tracking performance. Both techniques can be applied to any of the preamble approaches discussed above. However, the longer preambles have the advantage of yielding more accurate bit timing estimates. On the basis of these arguments, the M out of N preamble technique can yield the best bit timing accuracy at the end of the preamble.

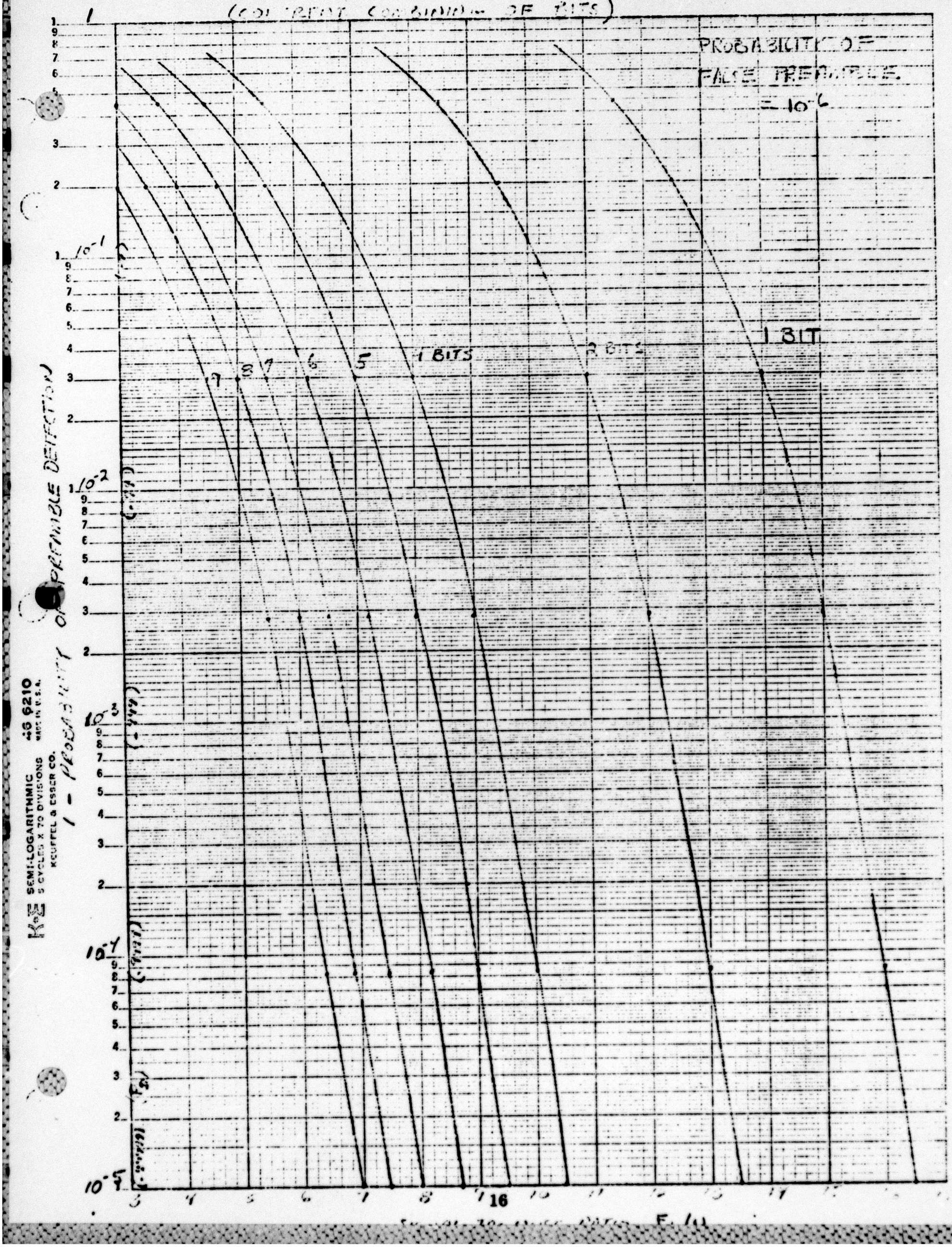
Multipath rejection is related to the width of the autocorrelation pulse and the sidelobe energy. Since the autocorrelation pulse width is identical in all approaches, only the sidelobes are of concern in this analysis. The sidelobe energy is greatest for the repeat preamble technique. The peak to sidelobe ratio is larger for the M out of N preamble detector than it is for the two coherent 7 bit Barker preamble approaches. Therefore, the multipath rejection is greater for the M out of N technique.

Doppler and frequency offset tolerance is related to the length of the SAWD and whether successive bits are combined coherently or non-coherently. The M out of N preamble technique gives the greatest tolerance because each bit is detected and then successive bits are accumulated. Thus, the phase error is accumulated over only a single bit (10μsec). One approach to solving the problem should the frequency error be excessive over a 7 bit interval is to estimate the frequency error and compensate for it at the injection frequency of the local oscillator. This adds to hardware complexity.

The equipment complexity comparison is subjective and was related largely on the basis of the number of SAWD's required and their design requirements.

One significant feature of several of the approaches discussed is the hardware commonality with the message demodulator. In the case of the second and fourth approaches listed in Figure 12, the MSK demodulators are identical for both the preamble and data. Some additional hardware is necessary for detecting the preamble.

On the basis of this tradeoff study, it is clear that the M out of N detector preamble with the 13 bit Barker followed by 13 bit inverted Barker is the recommended approach. One of the basic assumptions of this analysis is that the six tradeoff parameters considered are the most critical and that they are of equal importance.


PROBABILITY OF
FALSE PREAMBLE
= 10^-6
1 BIT
2 BITS
4 BITS
5
6
7
8
9
1 - PROBABILITY OF PREAMBLE DETECTION
K+E SEMI-LOGARITHMIC 46 6210
5 CYCLES X 70 DIVISIONS
KEUFFEL & ESSER CO.
MADE IN U.S.A.

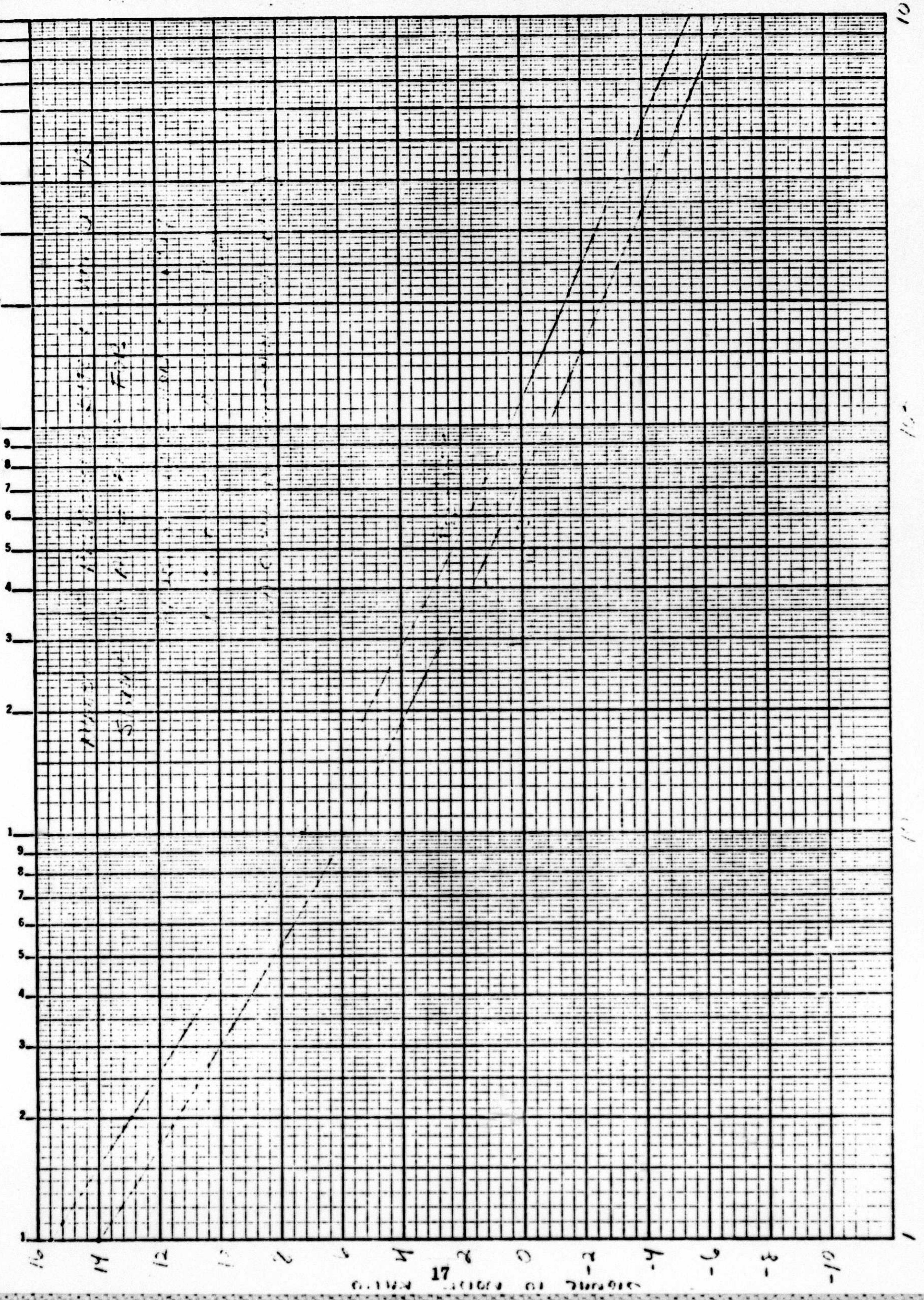
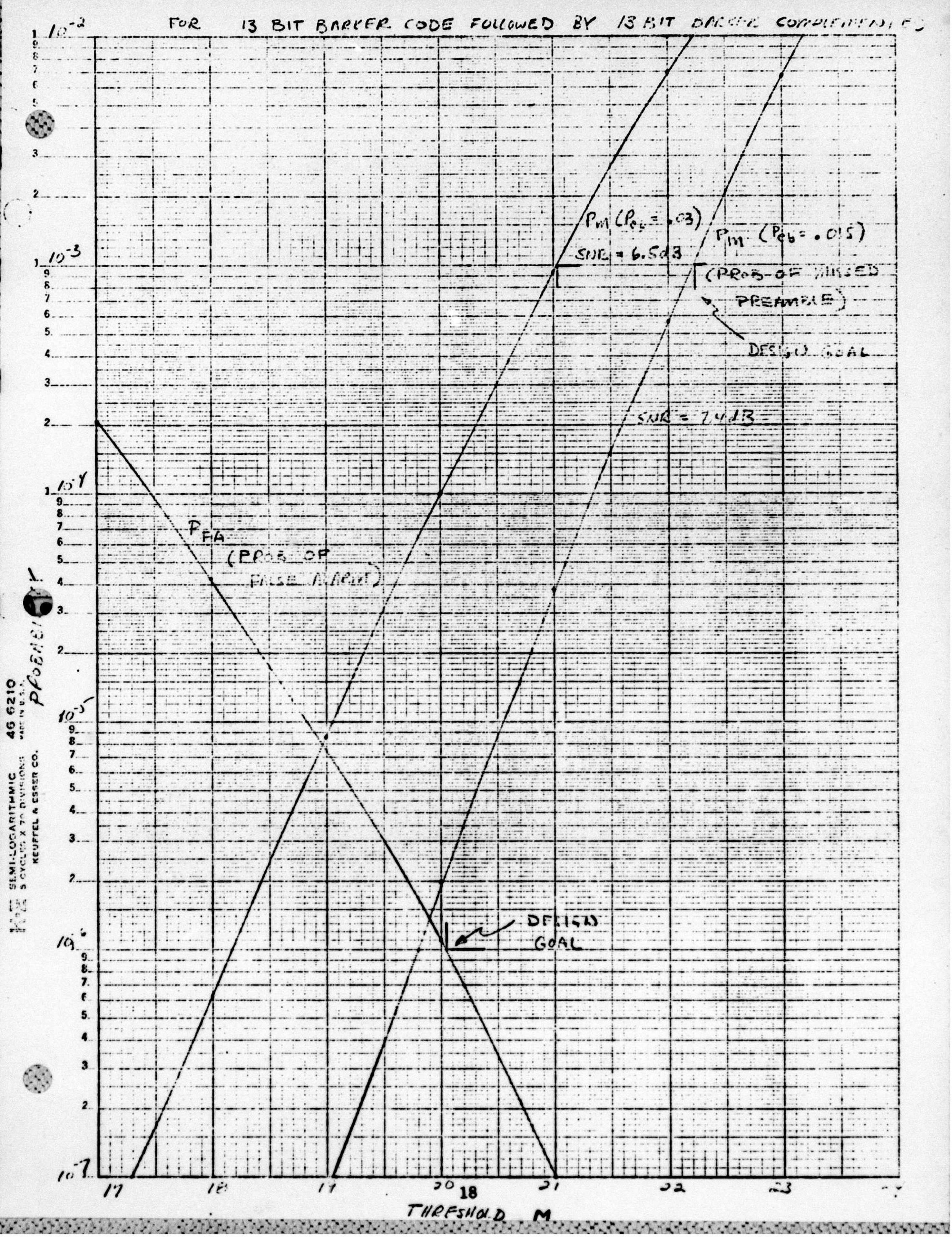

K&E SEMI-LOGARITHMIC 3 CYCLES X 70 DIVISIONS
46 5493 MADE IN U.S.A.
KEUFFEL & ESSER CO.

FIGURE 2


FOR 13 BIT BARKER CODE FOLLOWED BY 13 BIT BARKER COMPLEMENTED
P_M (P_{eb} = .03)
SNR = 6.5dB
P_M (P_{eb} = .015)
(PROB OF MISSED PREAMBLE)
DESIGN GOAL
SNR = 7.4dB
P_{FA}
(PROB OF FALSE ALARM)
DESIGN GOAL
PROBABILITY
THRESHOLD M
SEMI-LOGARITHMIC 46 6210
5 CYCLES X 70 DIVISIONS MADE IN U.S.A.
KEUFFEL & ESSER CO.

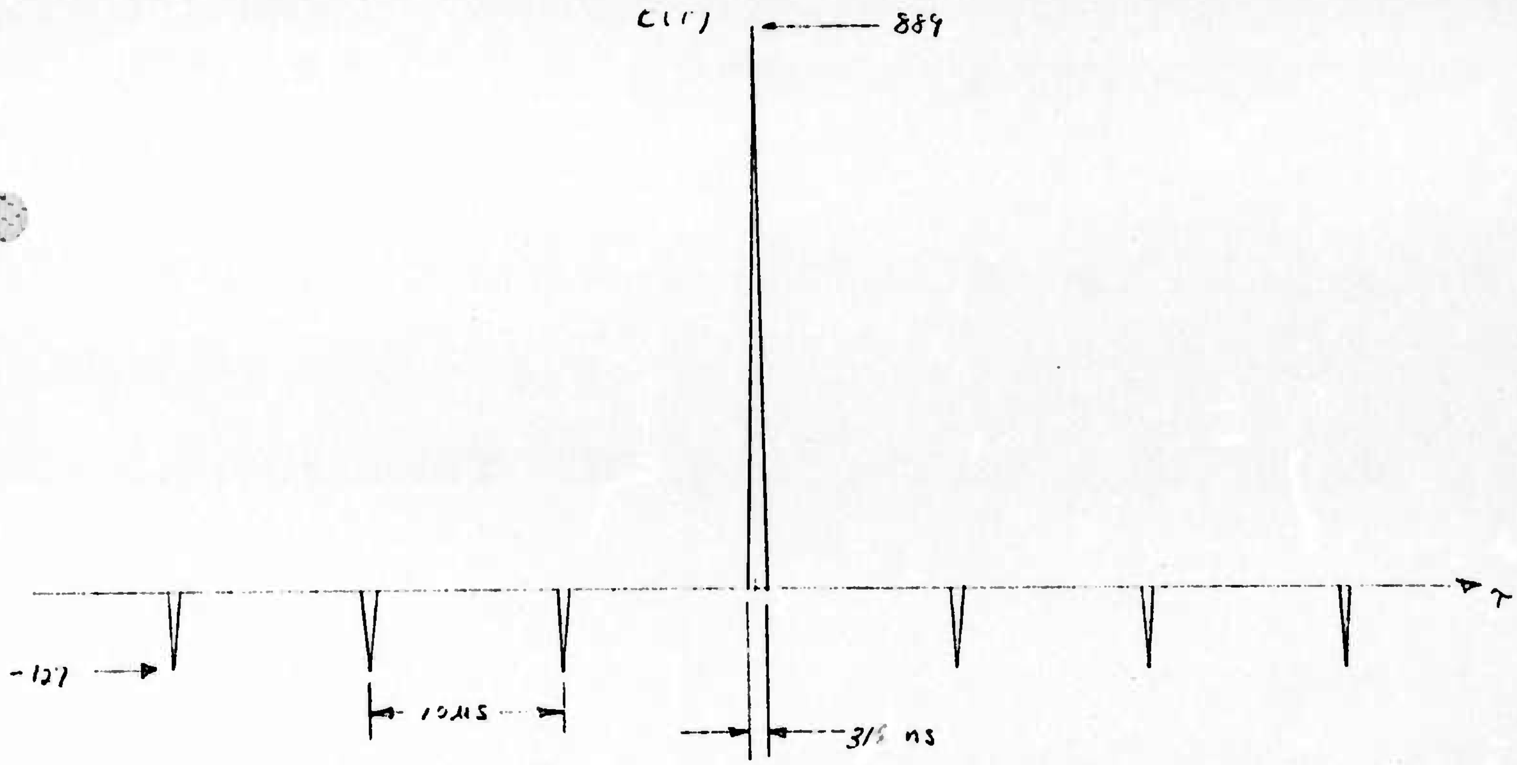


FIGURE 4 AUTOCORRELATION OF 7 BIT BARKER SEQUENCE WHERE EACH BIT CONSISTS OF 127 CHIPS OF A MAXIMUM LENGTH SEQUENCE AT 100 KBPS RATE

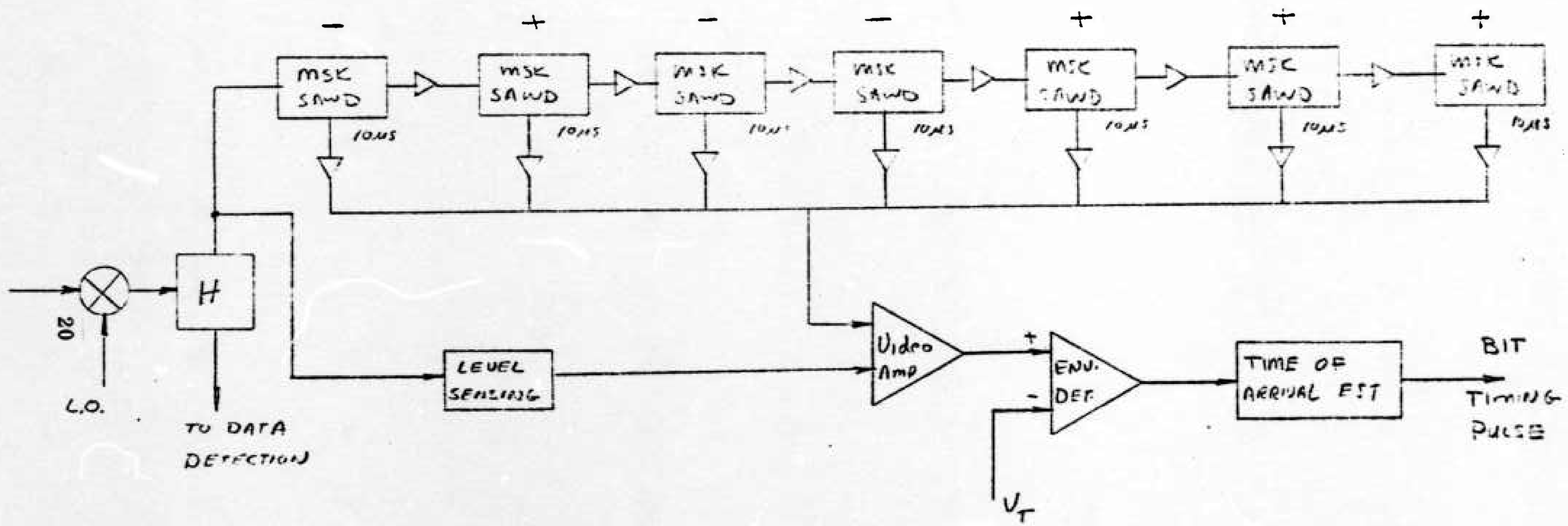


FIGURE 5 7 BIT BARKER PREAMBLE & BIT SYNC DETECTOR

COHERENT COMBINING DETECTION

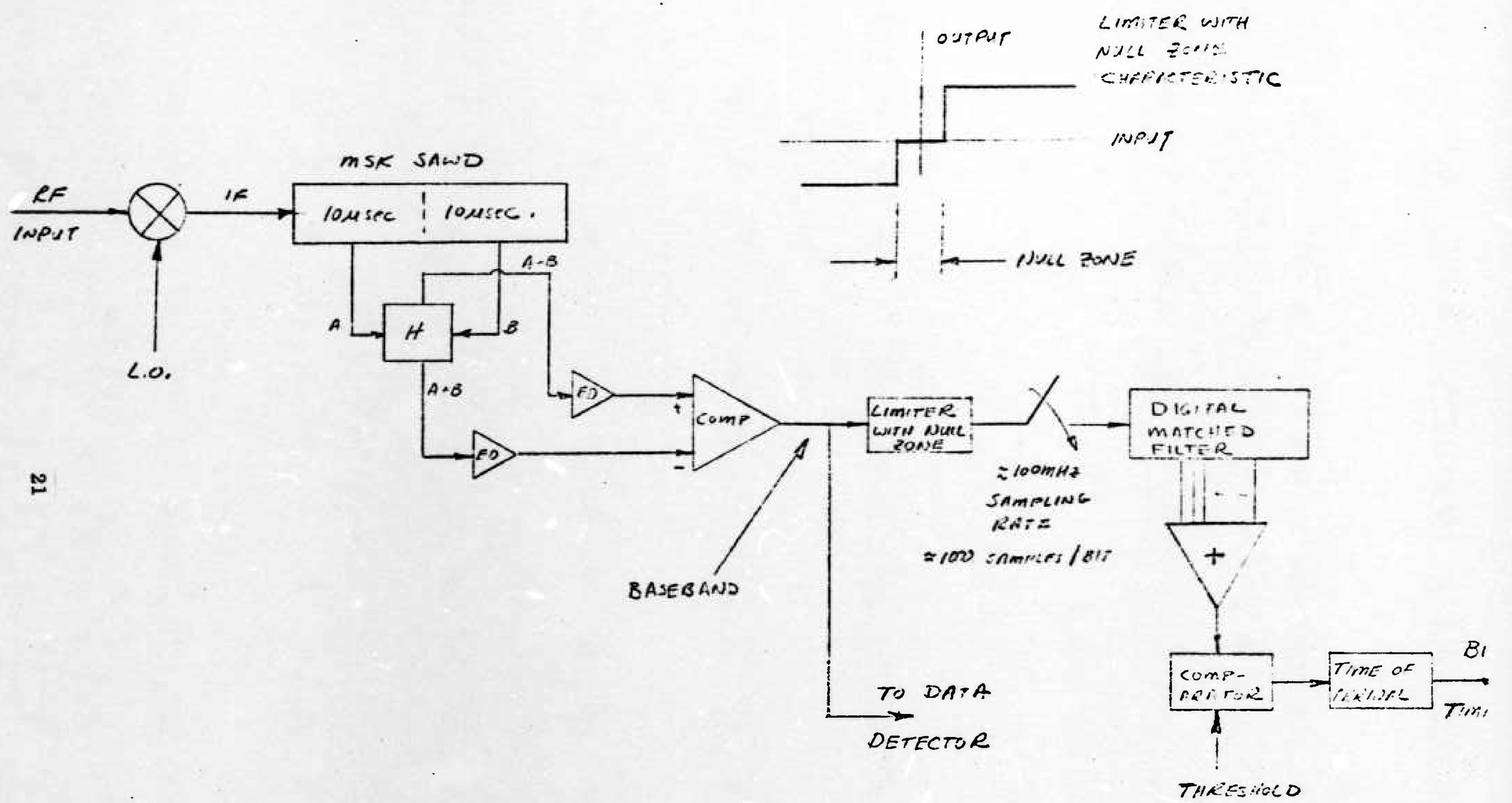


FIGURE 6 7 BIT BARKER PREAMBLE & BIT SYNC DETECTOR

DIFF. COHERENT COMBINING DETECTION

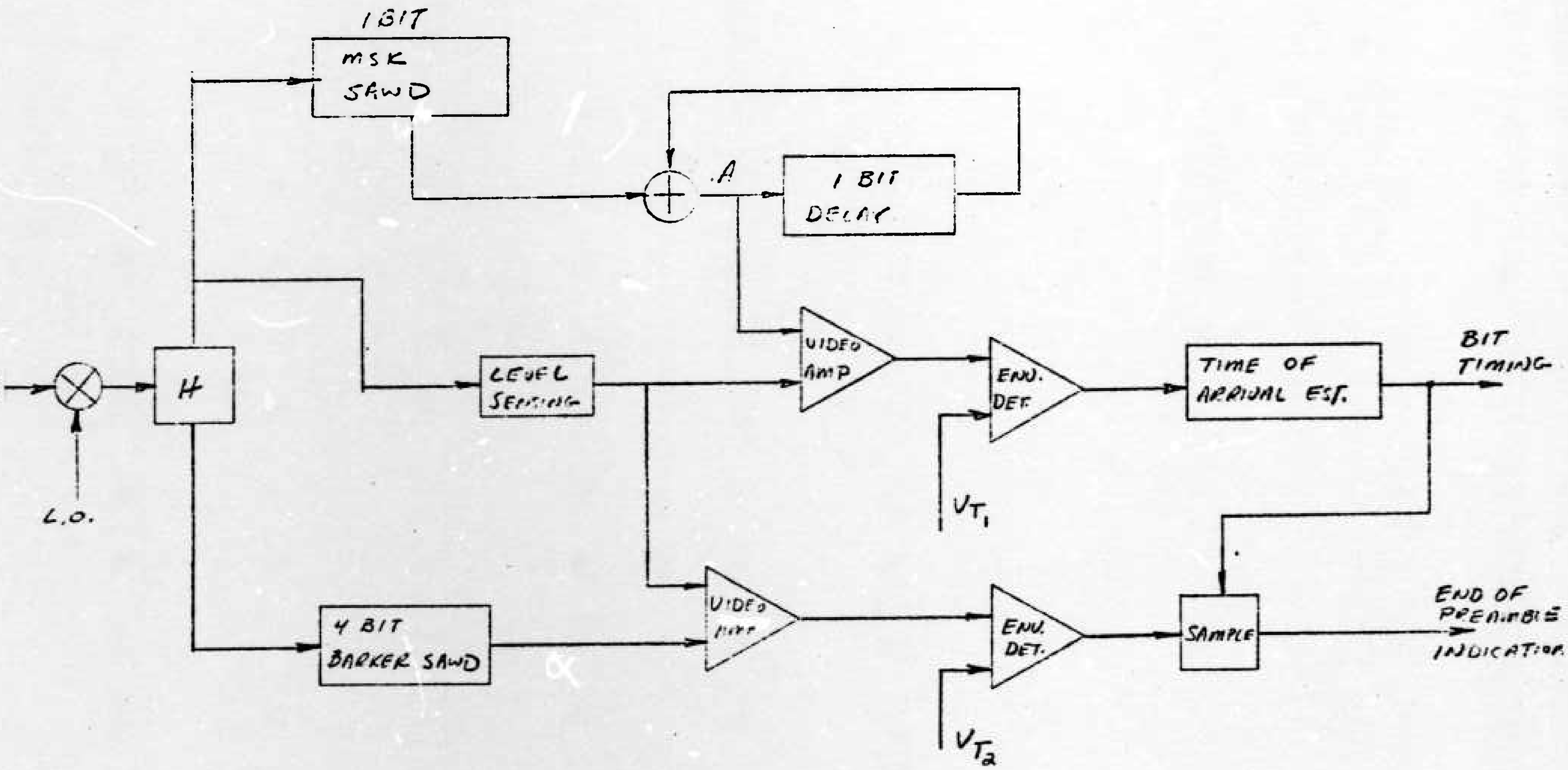


FIGURE 7 REPEAT PREAMBLE WITH END OF PREAMBLE DETECTOR

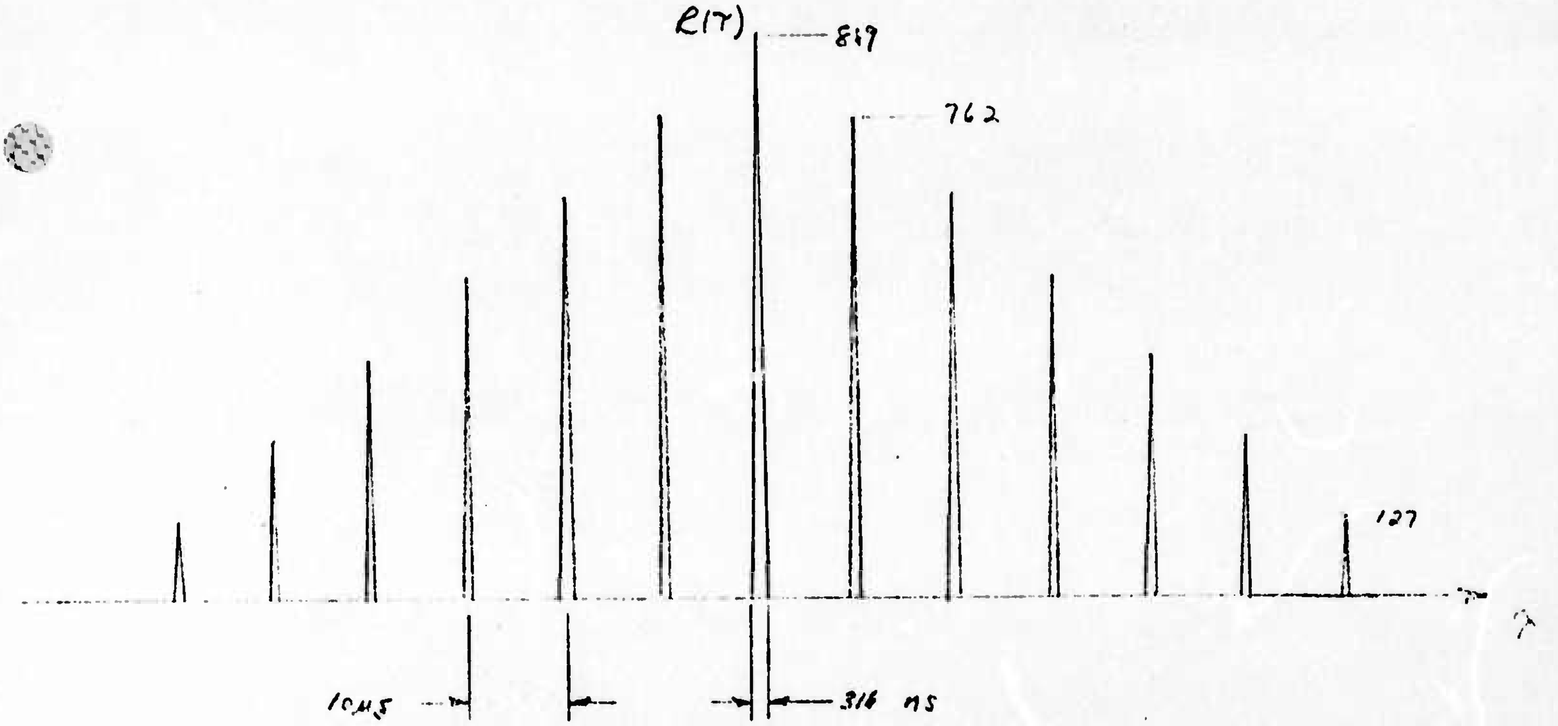
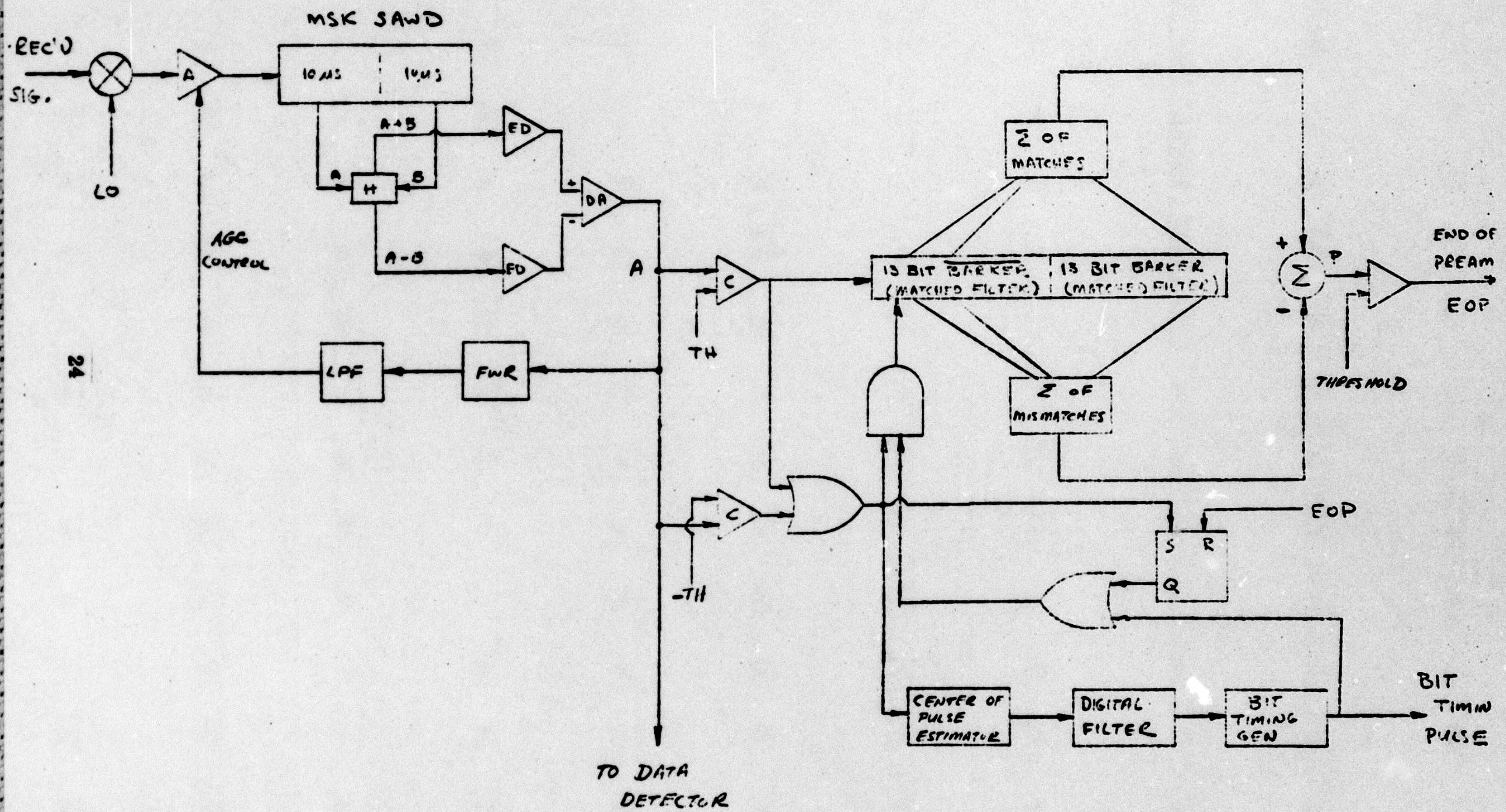
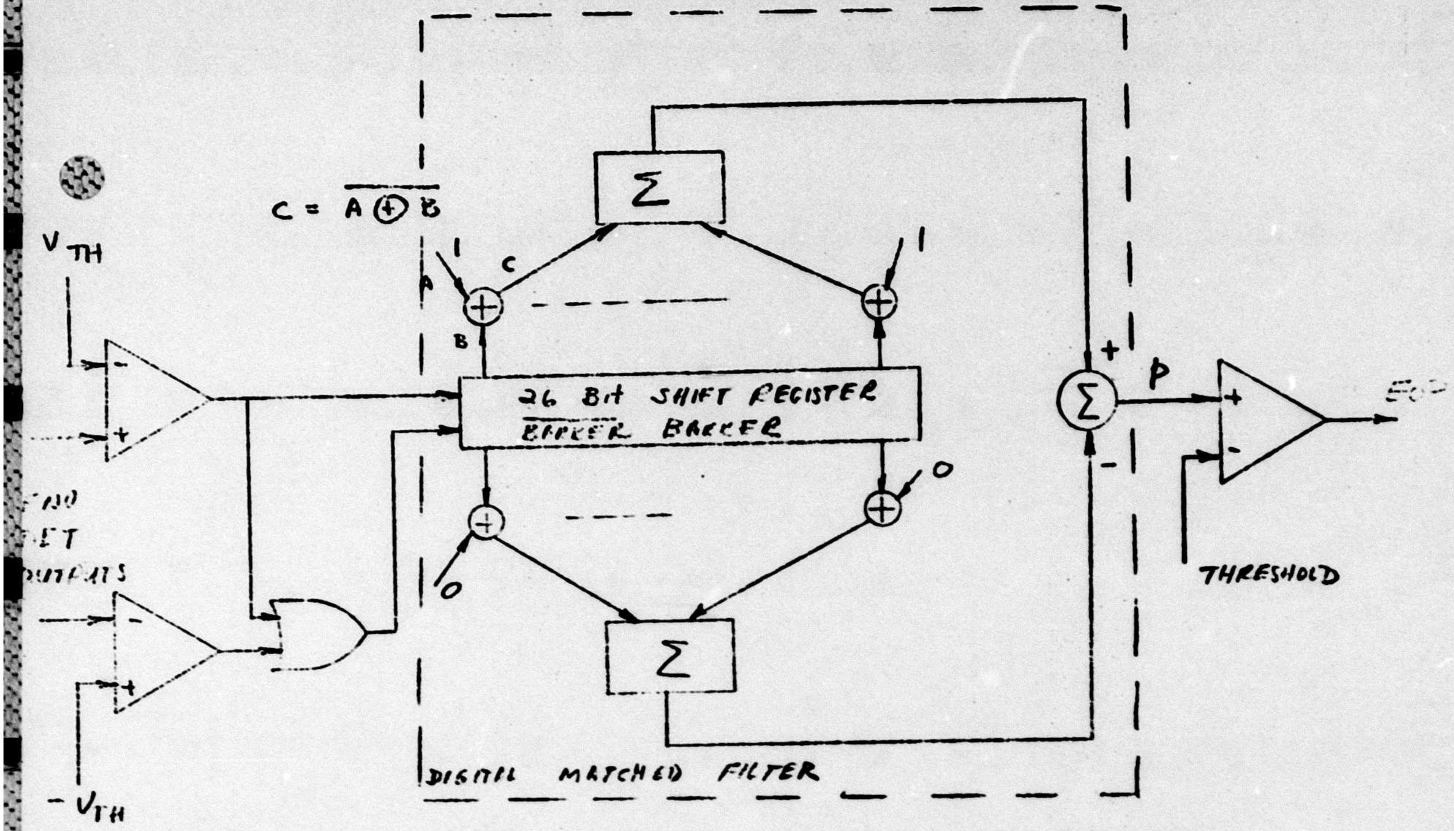


FIGURE 8 AUTO CORRELATION OF 7 "ONES" WHERE EACH BIT CONSISTS OF 127 CHIPS OF A MAXIMUM LENGTH SEQUENCE AT 100 KBPS.

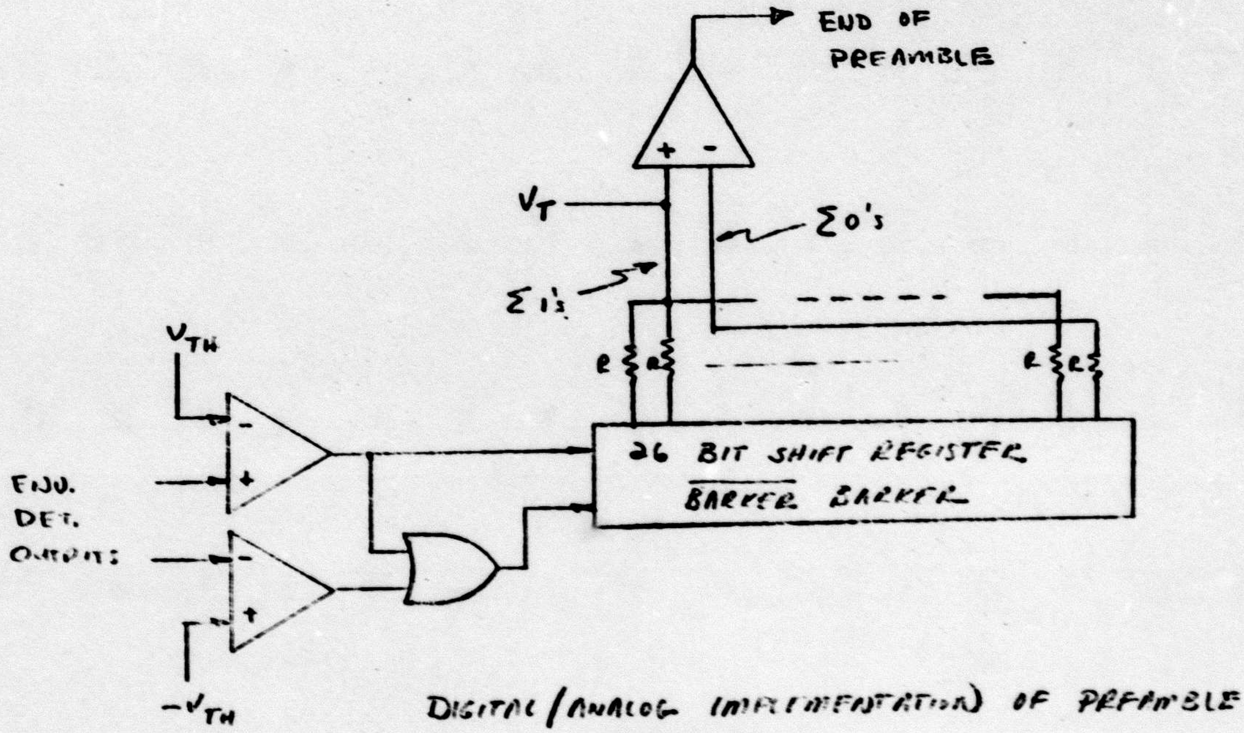
FIGURE 9A 13 BIT BARKER [illegible] PREAMBLE & BIT SYNC DETECTOR

(NONCOHERENT DETECTION)





DIGITAL IMPLEMENTATION OF PREAMBLE FILTER



DIGITAL/(ANALOG IMPLEMENTATION) OF PREAMBLE FILTER

FIGURE 9B

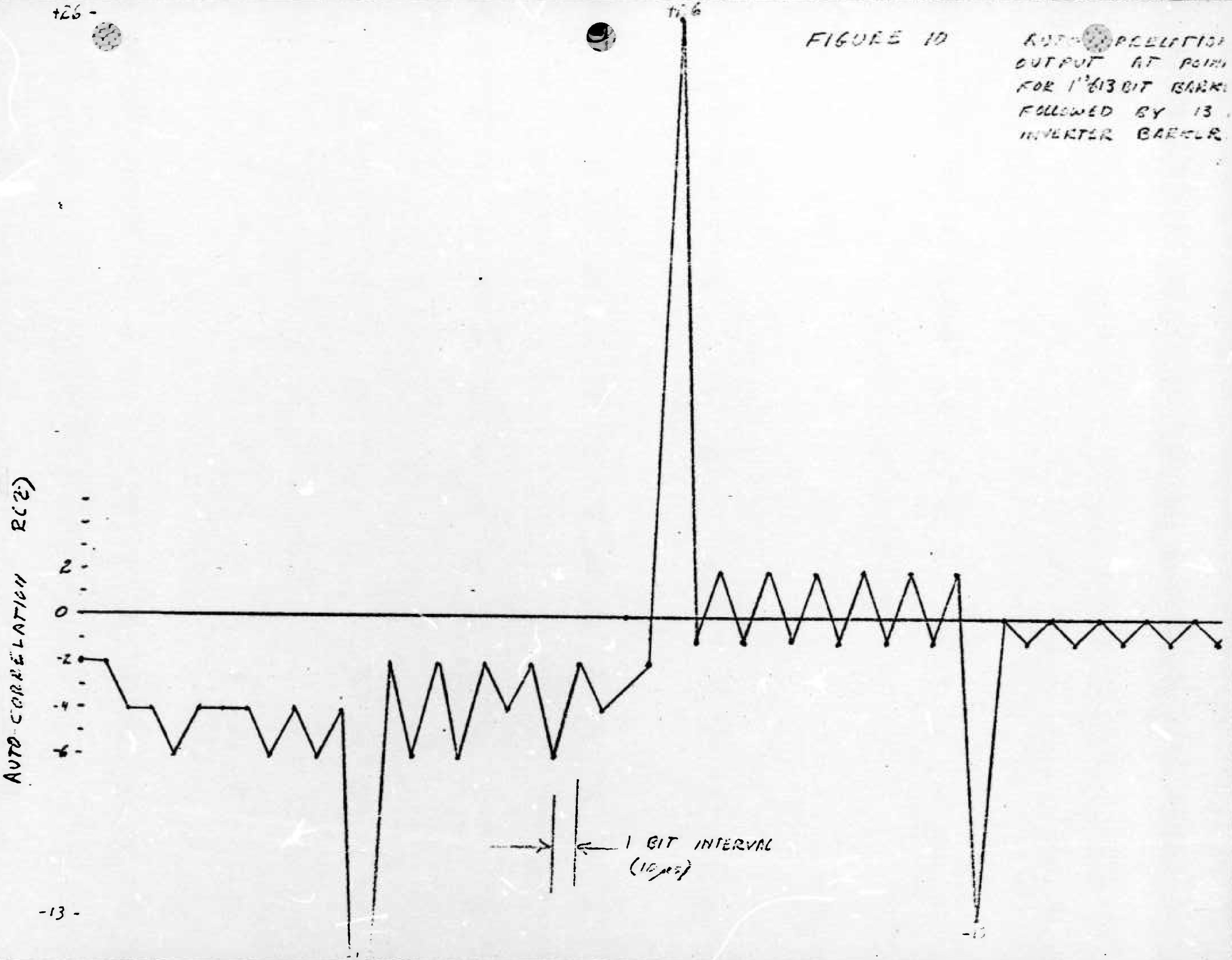


FIGURE 10 AUTO-CORRELATION OUTPUT AT POINT FOR 1st 13 BIT BARKER FOLLOWED BY 13 INVERTED BARKER

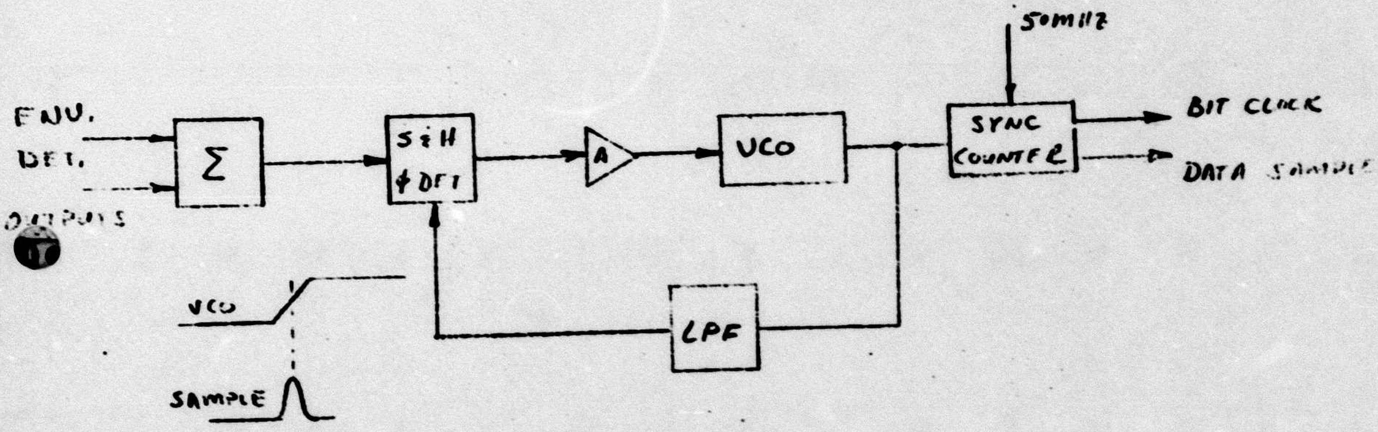


FIGURE 11 FUNCTIONAL DIAGRAM OF BIT SYNC LOOP

TRADEOFF PARAMETERS / PREAMBLE DETECTION TYPE	LENGTH OF PREAMBLE FOR FIXED P_D & P_{FA}	BIT TIMING ACCURACY	MULTIPATH REJECTION	DOPPLER AND FREQ. OFFSET TOLERANCE	EQUIPMENT COMPLEXITY	COMMONALITY WITH MESSAGE DEMODULATOR
BARKER PREAMBLE WITH COHERENT DETECTION	1	3	2	3	4	3
BARKER PREAMBLE WITH DIFF. COHERENT DETECTION	2	4	2	3	2	1
REPEAT PREAMBLE DETECTOR	3	2	3	2	3	2
M OUT OF N PREAMBLE DETECTOR	4	1	1	1	1	1

EVALUATION SCALE: 1 MOST DESIRABLE
TO
5 LEAST DESIRABLE

FIGURE 12

Appendix A

This appendix gives a summary of the analysis used to derive the detection probabilities of Figure 1. For a perfect envelope detector the probability of detection is

$$P_D = Q\left(\sqrt{\frac{2P_S}{\sigma^2}}, \frac{T}{\sigma}\right)$$

which is the Marcum Q function where P_S is the power in the carrier sinewave at the detector input, T is the bit length and σ^2 is the noise power. The derivation of this is shown in Appendix B. The improvement in signal-to-noise ratio when n pulses are integrated with ideal predetection integration is n times the signal-to-noise ratio of that for a single pulse. Therefore, 10 log n less signal-to-noise ratio is required in dB for n pulses combined coherently before detection.

Appendix B

A model of the preamble detection circuitry is shown in Figure B1. The local code is cross-correlated with the incoming signal in a SAWD. The SAWD performs the function of a modulator and an integrator. The effective bandwidth of the SAWD is $B = 1/T_B$ where T_B is the integration time. The incoming signal itself was spread to a bandwidth $W = 1/T_c$, where T_c is the code chip duration. After the integration for time $T = NT_B$, where N is the number of bits which are coherently added, the signal is envelope detected and applied to a threshold device for a synchronization decision.

The time required to achieve bit synchronization (i.e., for the detected output to exceed the threshold) is a function of the signal-to-noise ratio, the desired detection probability, P_D and the false alarm probability, P_{FA}. The probability densities of the noise and the signal plus noise required to calculate these probabilities depend upon N. For N = 1, the probability density functions (P.d.f.) at the envelope detector output are Rayleigh and Rician, respectively. For $N \geq 8$ the probability density functions can be approximated by Gaussian densities for both noise and signal plus noise.

Consider first the case for N = 1. If the detector is assumed to be a perfect envelope detector the p.d.f. of the output noise in the absence of a signal (i.e., in the absence of sync) is Rayleigh. Hence, the probability of false alarm is:

$$P_{FA} = \int_T^{\infty} \frac{v}{\sigma^2} \exp\left(-\frac{v^2}{2\sigma^2}\right) dv \qquad (1)$$

$$= e^{-T^2/2\sigma^2}$$

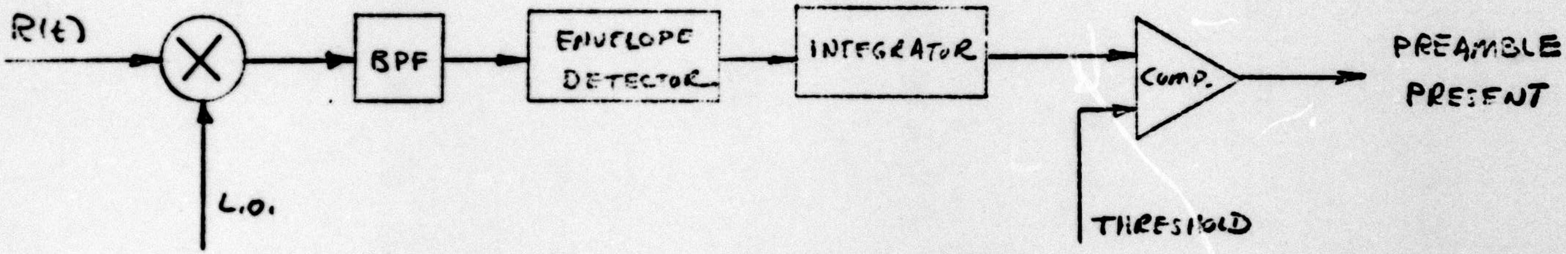


FIGURE B1 PREAMBLE DETECTION MODEL

when T = threshold voltage

σ^2 = $N_0 B$ = input noise power

for P_{FA} = 10^{-6}, we find that T/σ = 5.26 and

P_{FA} = 10^{-3}, we find that T/σ = 3.72

The p.d.f. of the detector output when one is within the two chip synchronization window is Rician. Hence, the probability of detection is

$$P_D = \int_T^\infty \frac{V}{\sigma^2} \exp\left(\frac{V^2 + 2Ps}{-2\,\sigma^2}\right) I_o\left(\frac{V\ 2Ps}{\sigma^2}\right) dv \qquad (2)$$

where Ps is the power in the carrier sinewave at the detector input. This is the Marcum Q function.

$$P_D = Q\left(\sqrt{\frac{2Ps}{\sigma^2}}\,,\ \frac{T}{\sigma}\right) \qquad (3)$$

for $P_{FA} = 10^{-3}$ so that T/σ = 3.72, we find that P_D = 0.999

for P_s/σ^2 = 22.5

Hence, the predetection SNR is:

SNR = 13.5 dB, $P_{FA} = 10^{-3}$, P_D = 0.999

For large BT products, i.e., for large N, we may use the central limit theorem to approximate the p.d.f. as Gaussian, both in and out of synchronism. Consider first the out-of-sync case and assume the detector is a square law device. The envelope detector is considerably more difficult to handle analytically; however, the results are very similar. The expected value of the detector output signal is:

$$E\{n\} = 2BT\,No$$

and the variance is

$$E\left\{n^2\right\} = \sigma_n^2 = (2BTNo)^2/BT = 4BTNo^2$$

The false alarm probability is therefore,

$$P_{FA} = 1/2\left[1 - \text{erf}\left(\frac{T - E\{n\}}{\sqrt{2}\,\sigma_n}\right)\right]$$

and for $P_{FA} = 10^{-3}$ we have $\dfrac{T - E\{n\}}{\sqrt{2}\,\sigma n} = 2.19$

If the codes are within sync (i.e., 2 chip window) the mean and variance of the detector output is:

$$E\left\{V\right\} = 2\,BTNo\left[1 + \frac{Ps}{BNo}\right]$$

and $E\left\{V^2\right\} = \sigma_s^2 = (2BTNo)^2\left[1 + \dfrac{2Ps}{BNo}\right]/BT$

from which we get $P_D = 1/2\left[1 + \text{erf}\left(\dfrac{E\{V\} - T}{\sqrt{2}\,\sigma_s}\right)\right]$

and for $P_D = .999$ we have $\dfrac{E\{V\} - T = 2.19}{\sqrt{2}\;\sigma_s}$

Now we are interested in solving for Ps/BNo = SNRi as a function of BT. We get

$$BT = \left(\frac{3.1 + 3.1\sqrt{1 + 2\,SNR_i}}{SNR_i}\right)^2$$

From this expression we can plot the results as given in Figure 2 for $P_{FA} = 10^{-3}$. A similar equation can be derived for different values of P_{FA}. The results are also plotted in Figure 2 for $P_{FA} = 10^{-6}$.

APPENDIX C

This appendix gives a summary of the analysis used to derive the preamble detection probabilities of Figure 3. In this approach a detection decision is made for each bit. To exceed the threshold for detecting a preamble requires M out of N bits being received correctly. If we assume that N "ones" are transmitted, then M-N errors are allowed in the received sequence. Therefore, the probability of detecting the preamble is

$$P_D = 1 - \sum_{i=1}^{M-N} \binom{N}{i} P_{eb}{}^{i} (1-P_{eb})^{N-i}$$

where P_{eb} is the probability of a bit error. For a 7.4 dB SNR and the threshold set for a probability of a false alarm per bit, P_{FA_b} = .3 per bit, the probability of a bit error is .015. With a SNR = 6.5 dB and the same P_{FA_b}, the probability of a bit error is .03. Without signal on the input a false alarm occurs with the occurrence of M or more "ones." Therefore, the false alarm probability can be written as

$$P_{FA} = \sum_{i=M-N}^{N} \binom{N}{i} P_{FA_b}{}^{i} (1-P_{FA_b})^{N-i}$$

Utilizing these two equations the results were plotted in Figure 3 for N = 26 and various values of M.

Appendix C. 3

Impact of Channel Options on Repeater Design

Appendix C.3

Impact of Channel Options on Repeater Design

INTRODUCTION

This note presents an analysis of various techniques proposed for the channelization of the packet radio network. The channelization methods considered do not include all possible, but represent a cross-section of many of the ideas of the group. The criteria of analysis is equipment complexity. The channelization schemes considered are as follows:

A. Narrow band single channel
B. Narrow band selective single channel
C. Narrow band dual adaptive selective channels
D. Code spread spectrum single channel
E. Code spread spectrum selective single channel
F. Code spread dual adaptive selective channels
G. Code spread spectrum with dual data rates
H. N-parallel code spread spectrum channels

Several assumptions are made in the evaluation of the candidate channelization schemes. These assumptions are:

1. Radio channel will be somewhere in the 1 GHz to 2 GHz range with approximately 150 MHz available bandwidth.
2. The modulation will be minimum shift keying (MSK).
3. Minimum data rate of the order of 100 KB/S
4. Target bit error probability of 10^{-5}.

Candidate Descriptions

A. The narrowband single channel is a fixed 150 KHz BW channel that the entire network would operate on. This represents the simplest approach to the question of channelization.

B. The narrow band selective single is identical to the narrow band approach with the additional flexibility of being able to choose the center frequency of the channel. This selection would be based on apriori information regarding the spectral environment with which groups of network elements must coexist.

C. Narrow band dual selective channels is a channelization method whereby a network element can adaptively choose the narrow band channel it will use for transmission and reception based on the traffic level and presence of non-network spectrum users. The repeater network element handles repeater-repeater traffic and terminal-repeater traffic simultaneously. A new channel selection would take place whenever the packet transmission time delays, number of retransmissions become excessive or when a non-network user causes coexistence problems.

D. Code spread spectrum single channel spreads the 150 KHz channel by a factor of the order of 100 to reduce the spectral density of the transmitted signal. This results in a channel of 15 MHz bandwidth. The spectral density of 1/100 of the narrow band case for a given transmitter power.

E. The code spread spectrum selective single channel is identical to D with the additional flexibility of selecting the center frequency of the channel based on apriori information about the spectral environment of the network element.

F. The Code Spread Dual Adaptive selective channel scheme provides for adaptively changing a spread spectrum frequency channel to avoid traffic delays or coexistence problems. The approach is similar to scheme C which was for narrow band channels. The terminal to repeater frequency is adaptively selected independent of the repeater to repeater frequency selection and both channels operate simultaneously.

G. Code spread spectrum with dual data rates is similar to E; however, a higher data rate is used for repeater-repeater traffic than for terminal-repeater traffic. The repeater-repeater link data rate would be of the order of 500 KB/S obtained by maintaining the 20M chip rate but reducing the number chips per bit. This would take advantage of the difference in the characteristics of the media between repeater-terminal mobile links which are subject to higher path loss, multi-path and impulsive noise and repeater-repeater links which should approach free space path loss and known multipath. This difference could be as much as 40 db in path loss of equal spaced links.

H. The N-parallel code spread spectrum is a channelization technique where a network element can select N-parallel channels from an ensemble of channels within a 150 MHz band to operate on simultaneously. This allows for (1) greater traffic volumes between lower path-loss links, such as repeater-to-repeater, (2) selection of channels that do not have coexistence problems and (3) redundancy (greater spread) in message transmission for designated critical traffic.

Equipment Complexity

This section discusses equipment implementation and the relative complexities for each channelization scheme considered in this note. Table 1 shows features of the transmitter and receiver excluding microprocessor and interface functions.

The appendix contains block diagrams pertinent to this discussion. Figure 1 is a general block diagram applicable to all channelization schemes.

Figure 2 depicts the recommended method of MSK modulation using SAWD's as impulsed match filters of 2 chip (bit for NB cases) length. This method is chosen over an impulse "coded SAWD" since more energy can be impulsed into the matched filter. The third column of the chart shows the length of SAWD required and the fourth column is the required IF. Higher IF's are required for adaptive channel schemes.

Figure 3 depicts the RF head portion of the packet radio for all channelization schemes. The difference between schemes is the bandwidth of the output bandpass filter, which is limited to $\geq 2\%$ BW due to filter limitations for NB cases and 140 MHz BW for WB cases.

The transmitter/receive LO are assumed to be identical frequencies. Two different schemes are envisioned, one for fixed frequency single channel operation; the other for a frequency agility that is implemented by a digital frequency synthesizer. These are shown in Figures 4 and 5 respectively.

The IF, signal detection, bit synchronization and pre-amble detection circuits are all shown in Figure 6. The AGC amplifier requirements, signal detection (following SAWD's), bit sync and pre-amble are essentially the same for all channelization schemes considered. As noted on the diagram, the SAWD's and SAWD implementation is dependent on channelization and, of course, dependent on modulation type.

Following is a discussion of equipment implementation for each channelization option.

A. Narrow Band Single Channel

This traditional channelization scheme is the simplest to design and implement. No code generator in the modulator is required. Standard 70 MHz IF's are chosen to use currently available circuits. Only hardware problem is implementation of 150 KHz bandwidth filter at 70 MHz. This is a 0.2% bandwidth. An LC filter cannot be built to meet this requirement, but a crystal filter can be designed to meet this requirement without too much difficulty.

B. Narrow Band Selective Channel

Since a frequency agile system is required, a synthesizer with 1 MHz steps would have to be designed. Also, RF antenna bandwidth and final BPS must be 140 MHz bandwidth to permit frequency channeling. IF's must be in the region of 200 MHz, since IF should be somewhat higher than RF bandwidth to preclude spurious emission in the antenna. Because of the narrow bandwidth (0.1%) at 200 MHz, LC filters are not practical and crystal filters are not feasible, a second IF (such as 10.7 MHz) is required to realize this bandwidth requirement.

C. Narrow Band Dual Adaptive Selective Channel

The requirements for this channelization scheme are the same as B, the NB selective channel set-up, plus additional isolation requirements. Single univarsal modules are required for terminals where one frequency is dynamically allocated. For repeaters, two modules are assumed. This essentially creates full duplex operation with repeater/terminal traffic on one channel and repeater/repeater traffic on another channel. Since these two modules are physically close together, a high degree of isolation is required from one transmitter to the adjacent receiver. If the antennas are 5 feet apart, then there would be only 40 db of propagation loss. This means that with 10 dbi of antenna gain on both transmit and receive antennas and the transmitter P.A. output of 40 dbm, then the input to the receiver preamp would be 100 MW with no filtering. For this reason, a very narrowband tunable (for adaptive selective channelization) filter is required. There are no tunable filters such as Yig filters which have 150 KHz bandwidth. The antennas might be arranged in such a way that additional isolation is obtained. For example, vertical polarized antennas could be stacked to provide additional isolation. However, this channelization method has no reasonable solution to full duplex operation.

D. Code Spread (WB) Single Channel

This scheme is similar to scheme A (NB single channel) except a spread spectrum using code spreading techniques is used.

A code generator must be used. This code generator uses conventional TTL, and must generate one of two codes dependent on the data bit value. A 30 MHz bandwidth filter is required, both at the RF in/output and at the IF. Unlike the NB case, a crystal filter design is not necessary. The receiver requires 100 chip coded SAWD matched filters.

E. Code Spread (WB) Selective Channel

The technique combines the features of B and D, NB selective channel and WB spread spectrum. Therefore, equipment complexities are commensurate; i.e., a wider RF bandwidth (140 MHz), a higher IF, coded SAWD's and a synthesizer are required. These particular requirements have been discussed in earlier sections.

F. Code Spread (WB) Dual Adaptive Selective Channels

This system embodies C and D configurations; i.e., selective, frequency adaptive channelizations with spread spectrum.

This channelization concept essentially requires all of the additional equipment circuitry discussed in all the previous sections; i.e., code generators, RF and IF bandwidths, coded detection SAWD's, synthesizers and isolation. The problem of isolation for full duplex operation can be solved with Yig filters, since the bandwidth is 100 times wider for the spread over the non-spread. However, these filters will require continuous power for tuning and heating (temperature compensation) of the order of 15 watts for the two filters. In addition, there will be as much as 5 db of insertion loss through the filter.

G. Code Spread Dual Rate Selective Channel

This channelization procedure is identical to E, except that dual rates (100 & 500 KBPS) are used. Spectrum occupancy and spectral density is identical in both rates since the chip rate (20 MCPS) is assumed constant and the spread factor Nc chips per bit reduced. If higher rates are assumed for repeater/repeater operation only, and since this operation is more than likely line-of-sight, higher SNR result and can sustain higher data rates without degradation.

Compared to E, WB Selective channeling, this technique would allow the same MSK modulator code generator to generate four codes (two at each rate) instead of two. This is the only transmitter circuitry over scheme E that is required, except for possible interface modification with the microprocessor.

The receiver would remain unchanged from that of receiver E up to the SAWD's. Here, two more SAWD's would be required at the higher rate. The remaining IF and signal processing circuitry would be duplicated, the second to handle the higher rate. This would include signal detection, pre-amble detection and AGC. If sample and hold type of phase detection (harmonic mode phase detection) is used in the bit synchronization circuitry, then the bit sync circuitry need not be reproduced.

H. N-Parallel Code Spread Channels

The channelization techniques would essentially be N modulators with - 2 watt PA's with a common power combiner to the antenna, where each modulator would be of the code spread type. Similarly, the receiver would use a common pre-amp followed by a power divider feeding essentially N independent down converters, IF's and demodulator circuits.

Special techniques could be used in the frequency generation circuit to generate N local oscillators to simplifying hardware requirements. Considering a linear power amplifier for the common amplification of these N channels, it should be noted that the peak power capability of the amplifier must be N^2 times the peak power for any one channel. This seriously limits the signal power for a given channel and reduces the range of operation. For example, a 10 watt peak power amplifier with five channels could handle only 400 MW on each channel for full linear operation. The efficiency of this P.A. would be approximately one-fourth that of a class C P.A.

CONCLUSION

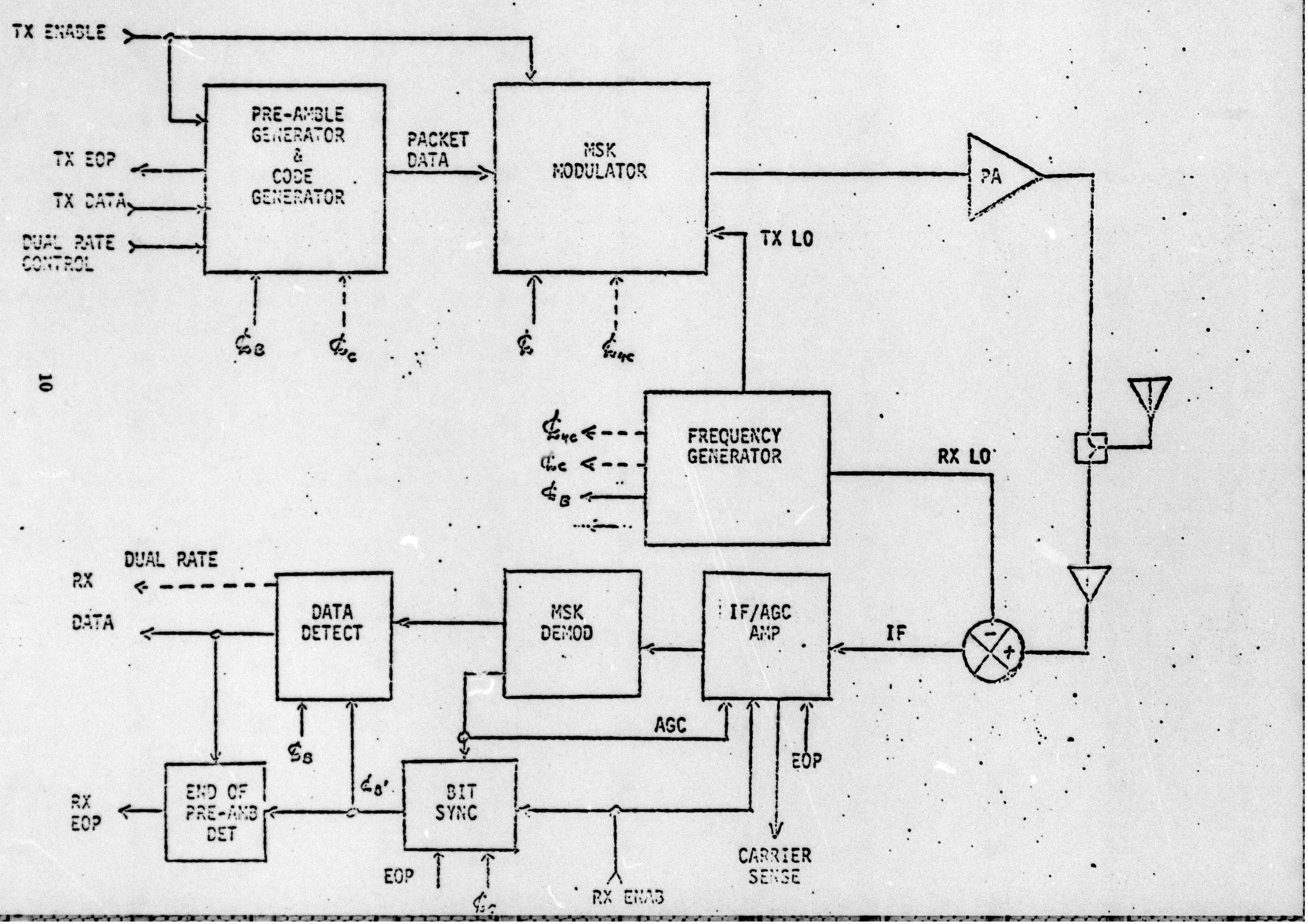
A summary of the previous sections is shown in the following table for equipment complexity.

A rating of 1 to 10 is used, 1 being most desireable or least equipment complexity. This rating is not in absolute terms, but only in relative terms or qualitative levels; i.e., a 2 rating has more equipment complexity than a 1 rating, but how much more complex is not implied.

TABLE OF CHANNELIZATION OPTIONS CRITERIA COMPARISONS

	EQUIPMENT COMPLEXITY
A. Narrowband Single Channel	(1)
B. Narrowband Selective Channel	(3)
C. Narrowband Dual Selective Channels	(10)
D. Code Sparead Single Channel	(2)
E. Code Spread Selective Channel	(4)
F. Code Spread Dual Adaptive Selective Channels	(9)
G. Code Spread Dual Data Rates Selective Channel	(6)
H. N-Parallel Code Spread	(8)

( ) Ratings to 10 - 1 being least complex
10 being most complex



TX ENABLE
TX EOP
TX DATA
DUAL RATE CONTROL
PRE-AMBLE GENERATOR & CODE GENERATOR
PACKET DATA
MSK MODULATOR
PA
TX LO
FREQUENCY GENERATOR
RX LO
IF
IF/AGC AMP
MSK DEMOD
DATA DETECT
DUAL RATE
RX
DATA
AGC
EOP
CARRIER SENSE
BIT SYNC
END OF PRE-AMB DET
RX EOP
RX ENAB

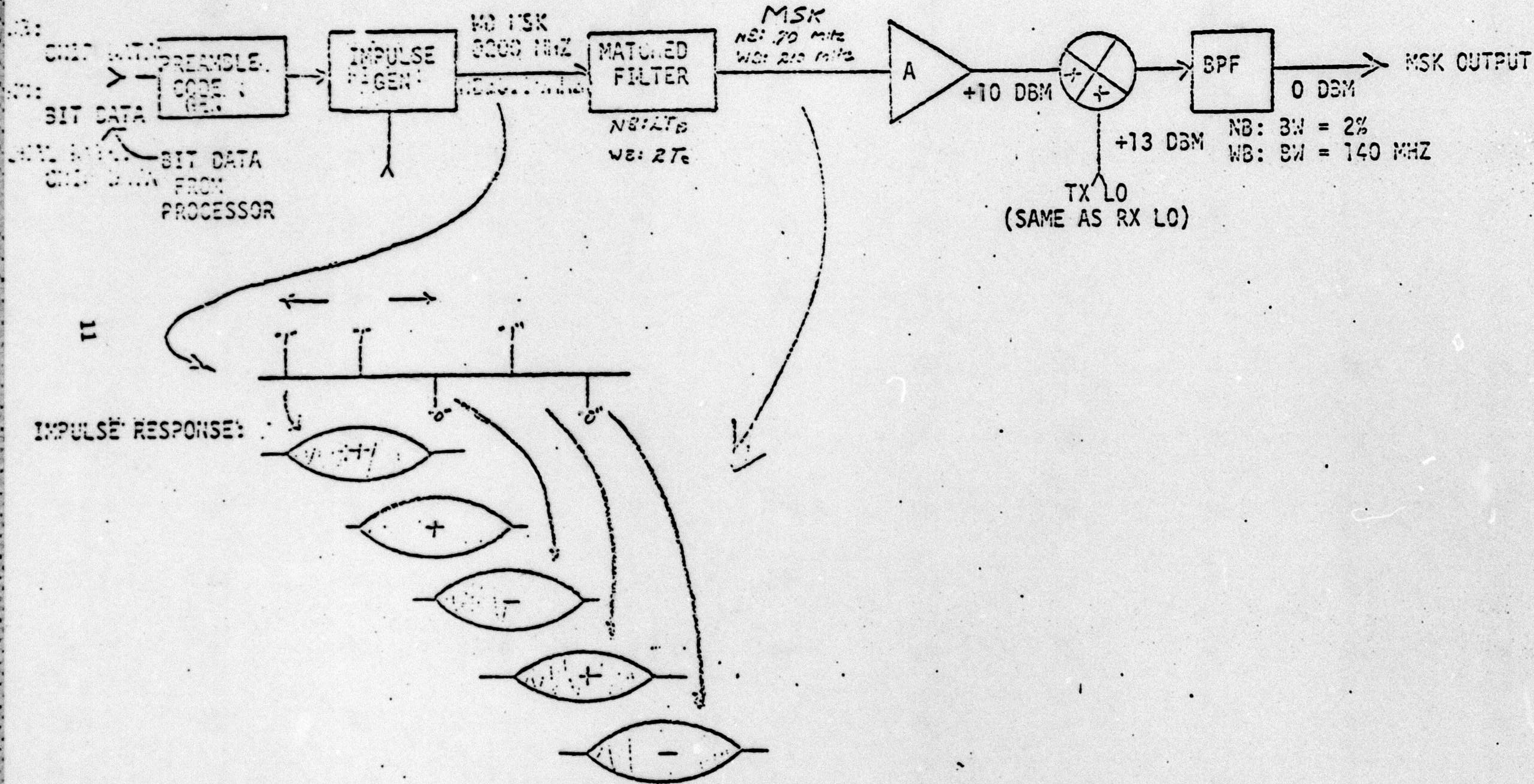


FIGURE 2: MSK MODULATOR

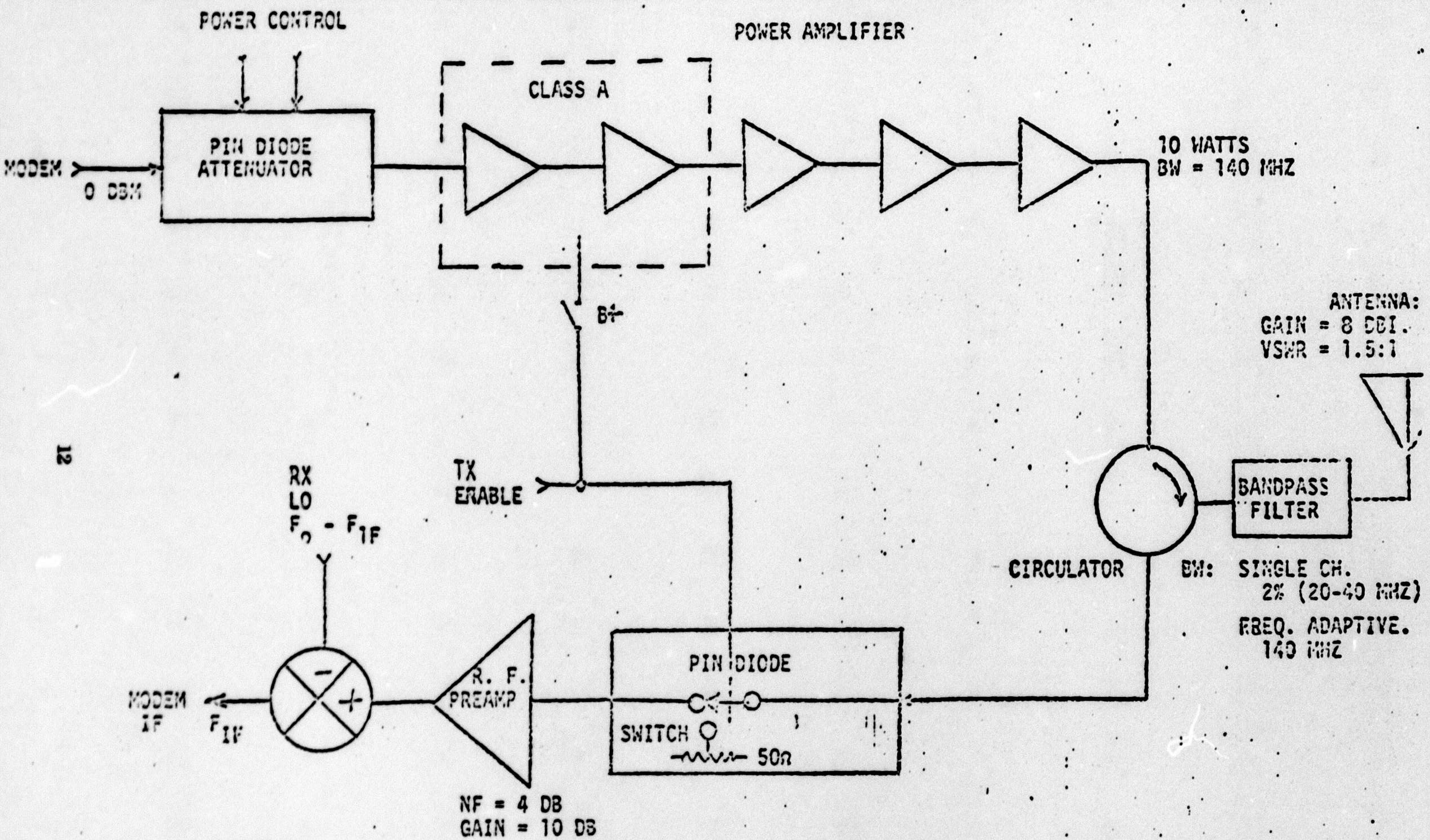


FIGURE 3: R.F. SYSTEM

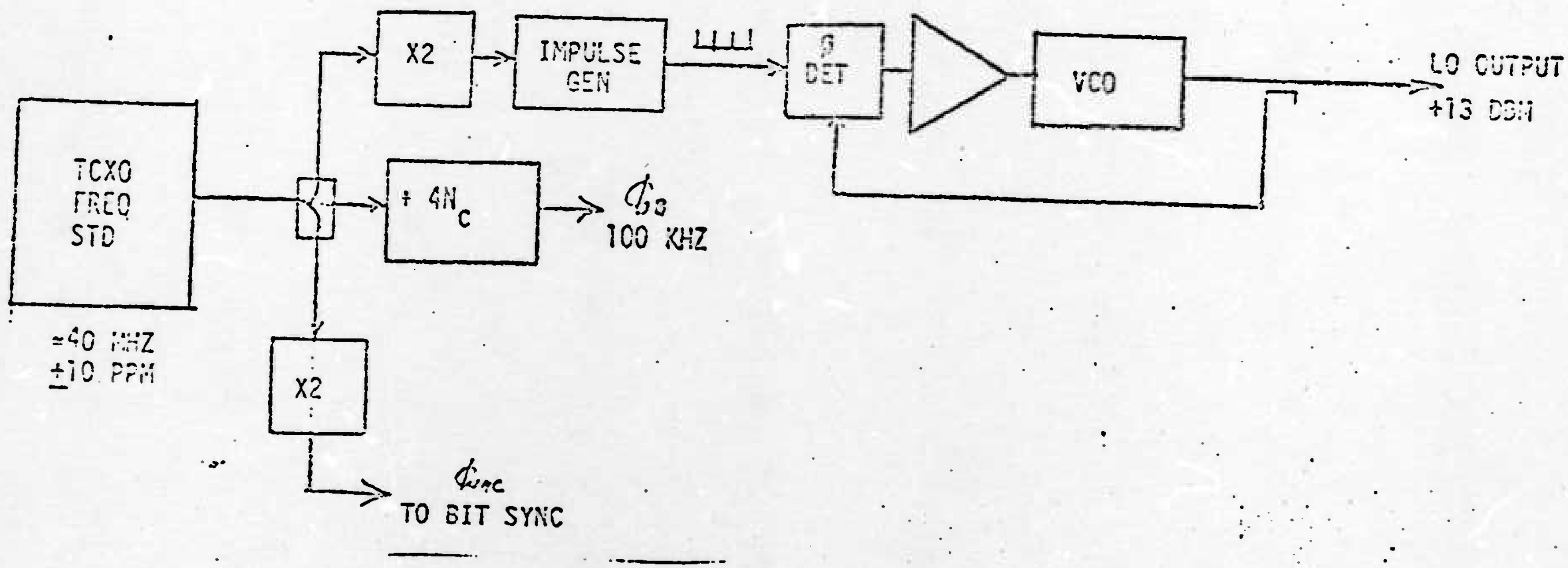


FIGURE 4: FREQUENCY GENERATION
NB & WB SINGLE CHANNEL

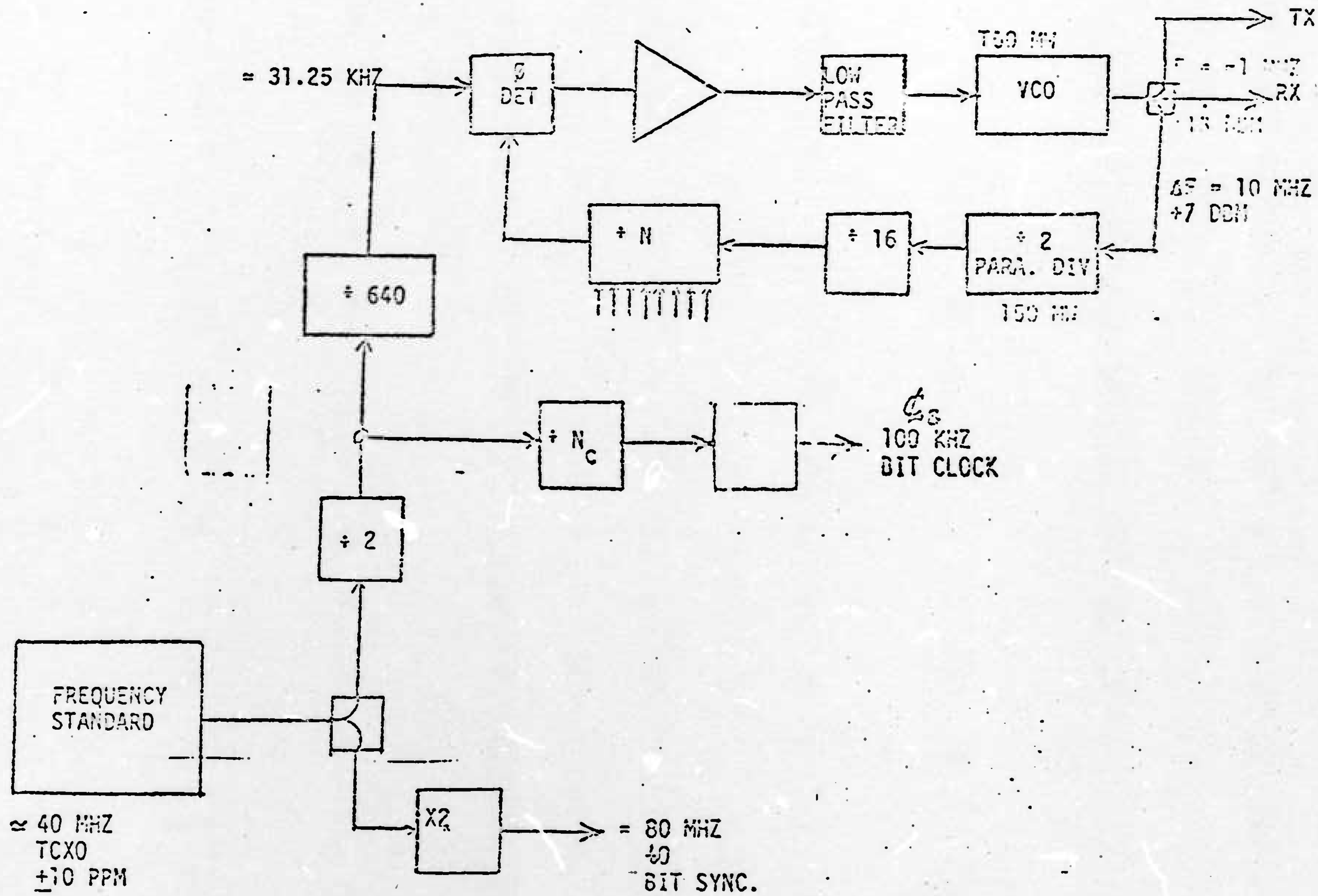
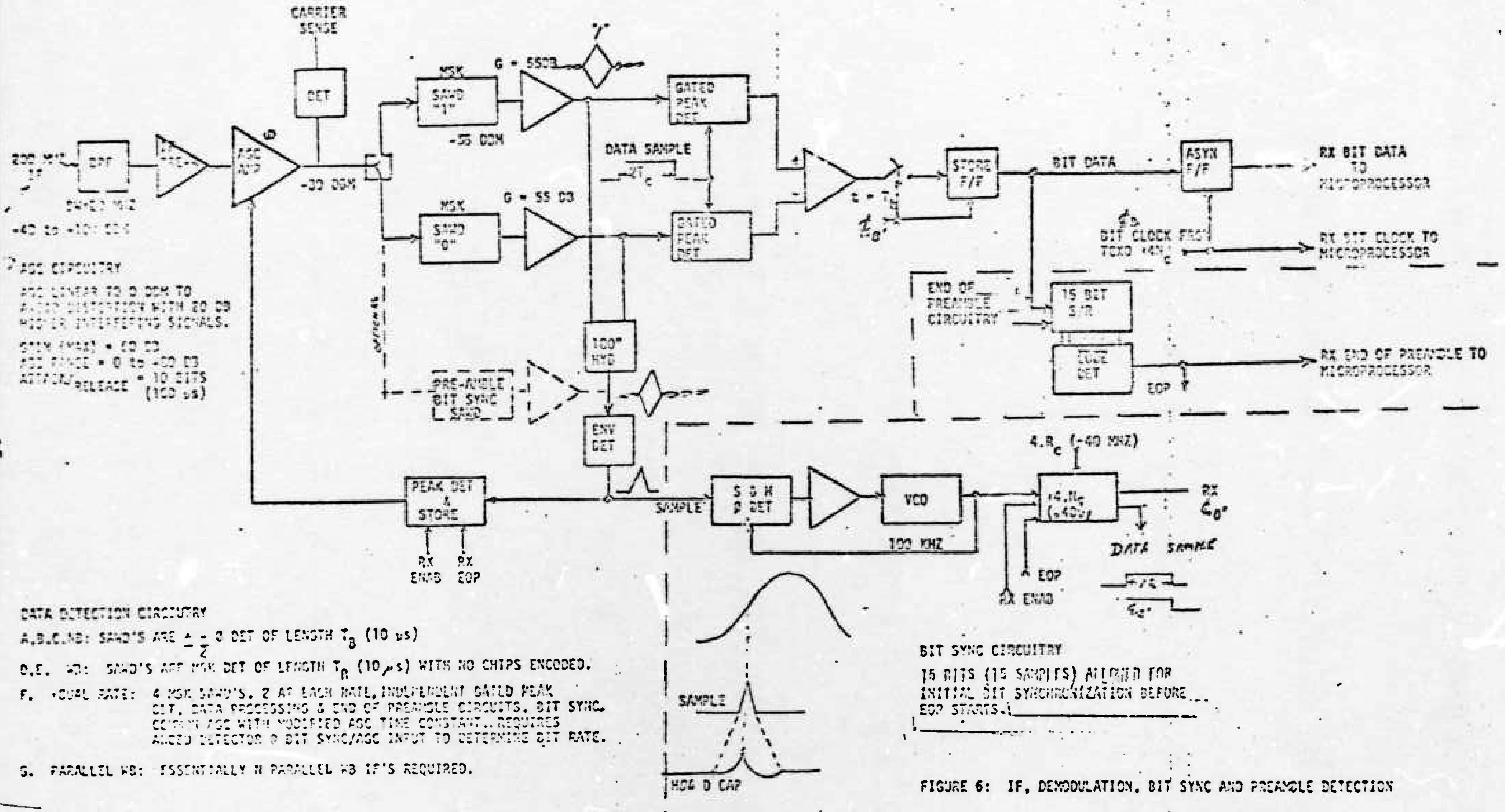


FIGURE 5: FREQUENCY SYNTHESIZER, 1 MHZ STEPS
NB & WB FREQUENCY ADAPTIVE CHANNELS AND
DUAL RATE



DATA DETECTION CIRCUITRY

A,B,C,NB: SAWD'S ARE $\pm\frac{\pi}{2}$ DET OF LENGTH T_B (10 µs)

D,E. WB: SAWD'S ARE MSK DET OF LENGTH T_B (10 µs) WITH NO CHIPS ENCODED.

F. DUAL RATE: 4 MSK SAWD'S, 2 AT EACH RATE, INDEPENDENT GATED PEAK DET, DATA PROCESSING & END OF PREAMBLE CIRCUITS. BIT SYNC. COMMON AGC WITH MODIFIED AGC TIME CONSTANT. REQUIRES ADDED DETECTOR @ BIT SYNC/AGC INPUT TO DETERMINE BIT RATE.

G. PARALLEL WB: ESSENTIALLY N PARALLEL WB IF'S REQUIRED.

BIT SYNC CIRCUITRY

15 BITS (15 SAMPLES) ALLOWED FOR INITIAL BIT SYNCHRONIZATION BEFORE EOP STARTS.

FIGURE 6: IF, DEMODULATION, BIT SYNC AND PREAMBLE DETECTION

SIZE	CODE IDENT	DWG NO.
B	13499	
SCALE	REV	SHEET

191

Appendix C.4

Power Budget Analysis

1.0 INTRODUCTION

The power budget for the repeater design is presented in this note. The analysis is similar to Note #4 by Cory; however, this analysis assumes parameters which more accurately reflect the design of the experimental system. Additionally, the question of dynamic range of signal levels is addressed for a given repeater spacing and minimum distance between terminals. The objective is to provide some insight into the allocation of this dynamic range between a transmit power control and the receiver AGC. Sensitivity analysis of this dynamic range to repeater spacing and minimum terminal spacing is included.

2.0 ASSUMED SYSTEM PARAMETERS

The following assumptions are made about system parameters for the power budget calculations:

1. Carrier frequency range is 1700 MHz to 1850 MHz.
2. Channel Characteristics:
   Path loss is corrected according to Okumura, et. al. [1] for repeater-terminal links where the repeater antenna

is assumed to be at 45M height and the terminal antenna at 3M. Free space loss is assumed between repeaters with both antennas at the 45M height. Non-additive white gaussian noise is assumed.

3. Data rate is 100 KB/S for repeater-terminal links and 500 KB/S for repeater-repeater links.
4. The modulation is differential coherent minimum shift keying (DCMSK). For this modulation method, the ratio of hertz per bit is 1.5. This includes 95% of the signal energy.
5. Code spread of 100 for terminal-repeater links and 20 for repeater-repeater links.
6. Noise figure of receiver is 9 db.
7. Antenna gain above isotropic is 0 db for the terminal and 9 db for repeater.
8. The transmitter power amplifier output is 10 watts.
9. The maximum probability of a bit error is 1×10^{-5}.

3.0 THE POWER BUDGET EQUATION

The basic relation for determining the carrier to noise power relation (C/N) is the following equation:

$$C/N \text{ (in db)} = ERP - \text{Path loss} + G_{ra} - 10 \log KT_oB - NF \qquad (1)$$

Where: ERP = Effective radiated power in dbm
G_{ra} = Receive antenna gain
KT_o = Noise spectral density
B = Occupied bandwidth
N.F. = Noise figure of receiver in db

The occupied bandwidth for the assumed system parameters is:

$$B = (\text{Spread Factor}) \times (\text{Data Rate}) \times (\text{Hz/Bit Factor}) = 15 \text{ MHz} \quad (2)$$

for repeater to repeater and repeater to terminal links.

The relation between signal to noise and bit error performance for DCMSK is given in Figure 1. It should be noted that signal to noise is expressed in E_b/N_o where E_b is the energy per bit. Hence, one must convert C/N to E_b/N_o. This can be accomplished by multiplying C/N by T (the bit interval) and E_b/N_o by 1/B to yield:

$$\frac{C \cdot T}{N} = \frac{E_b}{N_o \cdot B} \quad (3)$$

so that

$$E_b/N_o = C/N \cdot TB \quad (4)$$

The factor TB can be viewed as the product of the processing gain and the Hz/bit factor. The noise power, N, also has this factor and, therefore, E_b/N_o is independent of the code spread and Hz/bit. The relation for E_b/N_o is then:

$$E_b/N_o = \text{ERP} - \text{Path loss} + G_{ra} - 10 \log KT_oB^1 - \text{N.F.} \quad (5)$$

Where:

$$B^1 = (\text{Data Rate})$$

4.0 TRANSMISSION LOSS

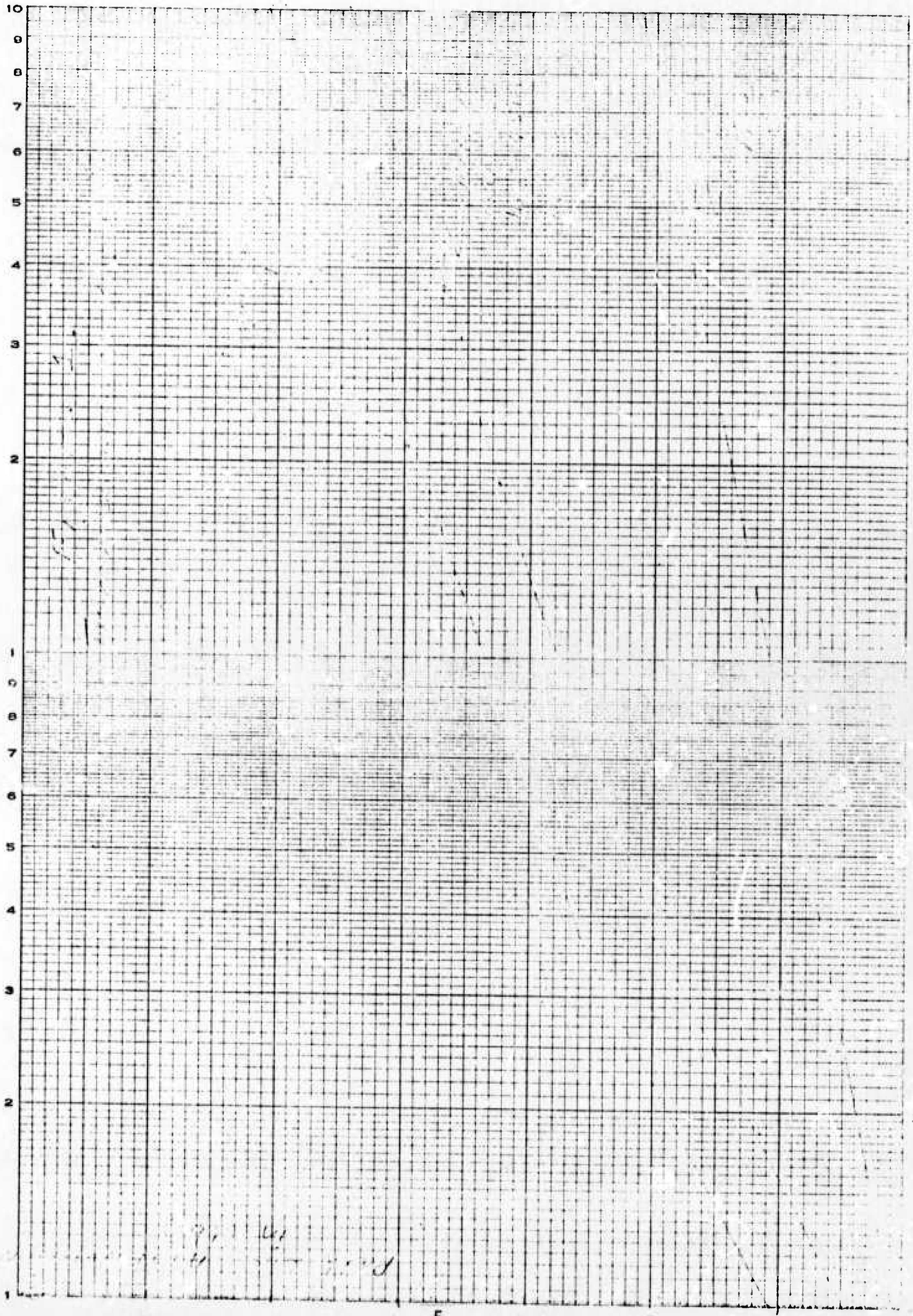
The free space loss equation must be altered for the mobile conditions of repeater-terminal links. Okumura provides statistical data for this correction. For the assumed antenna heights, the median path loss per Okumura at 1 mile is approximately 34 db greater than the free space loss. Transmission loss, Okumura corrected, varies with distance at approximately 40 db/decade, compared with free space loss of 20 db/decade. Figure 2 shows the transmission loss for free space versus distance, the median Okumura loss, and the 99% Okumura loss. The 99% line provides insight into the variance of Okumura statistics and is interpreted to be the value for which the probability of the loss at any time being less than or equal to this value is .99.

5.0 REQUIRED SIGNAL TO NOISE RATIO

The minimum E_b/N_o for P_e of 10^{-5} or less can be obtained from Figure 1. For DCMSK, E_b/N_o must be greater than or equal to 10.4 db.

6.0 HYPOTHESIZED REPEATER SPACING

In order to determine the dynamic range of the signal level, one must make assumptions about the repeater spacing and the minimum distance between terminals. The maximum distance between a repeater and terminal using Equation (5) and E_b/N_o of 10.4 db is 1.4 miles for Okumura 99% transmission loss. The spacing between repeaters will be assumed to be equal to this maximum distance to assure full coverage of a given area and provide a minimum of two repeaters being able to cover any terminal location. Figure 3 shows this ideal geometry of repeater spacing ignoring geographical constraints such as mountains, etc. The minimum


10
9
8
7
6
5
4
3
2
1
9
8
7
6
5
4
3
2
1
5

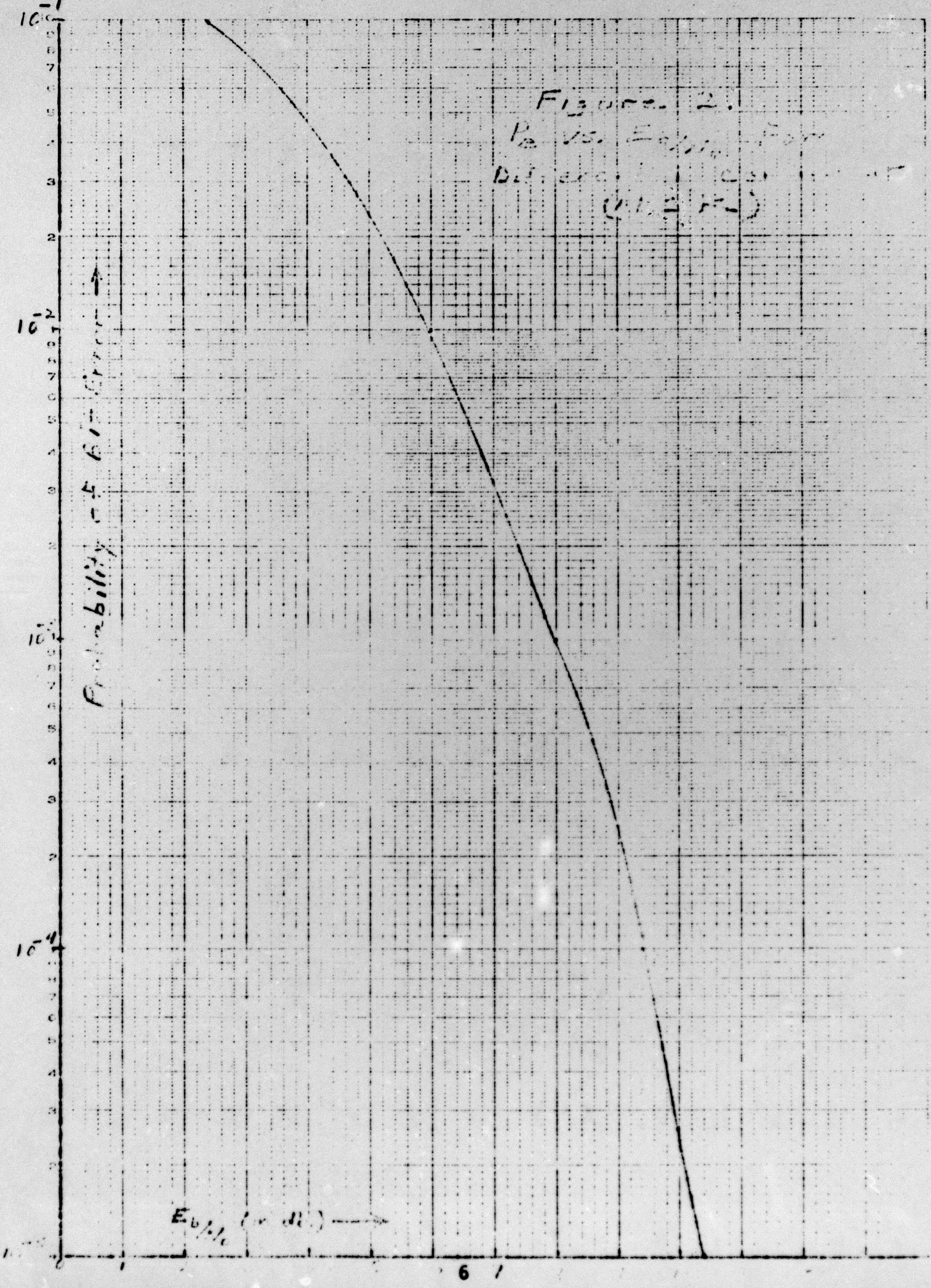
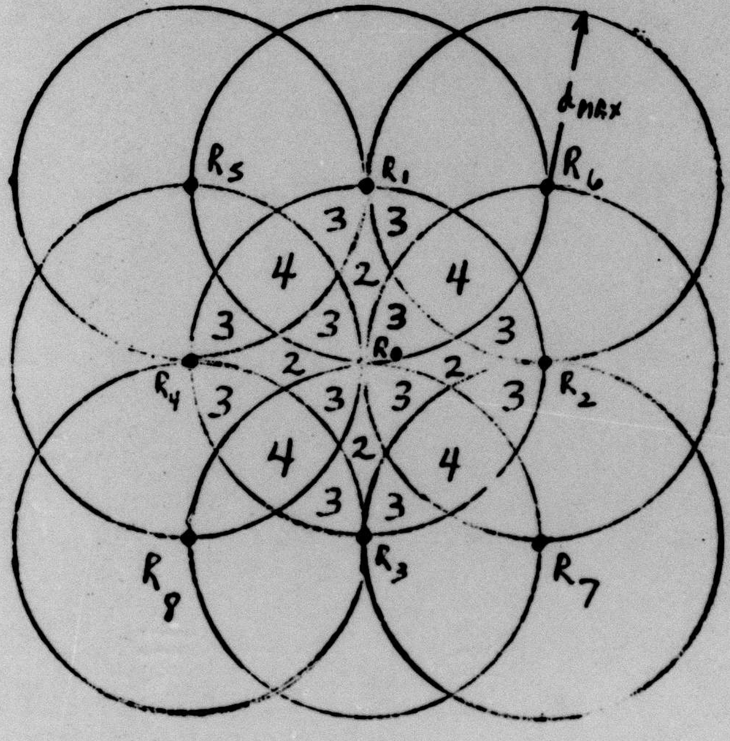


Figure 2.
Pe vs. Eb/No For
Probability of Bit Error
Eb/No (in dB)



note:

1. d_{max} – maximum Terminal-Repeater link distance

2. Numbers in R_0 circle of coverage indicate no. of repeaters covering that area.

Figure 3. Hypothesized Repeater Spacing

distance between terminals is assumed to be 50 feet.

The threshold level of C/N for E_bN_o = 10.4 db is - 11.3 db which means that the signal + noise power is approximately equal to KT_oB or - 102.5 dbm. The maximum received signal power will occur between terminals spaced the minimum distance (50 ft.). In this case, the path loss will be obtained from the free space transmission loss equation. From Equation (1) and the assumed system parameters, the signal power is - 18 dbm. Therefore, the dynamic range of signal energy is 84 db. This dynamic range decreases as the minimum distance between terminals increases at a rate of 6 db/octave.

If terminals are uniformly distributed radially from a repeater to the maximum distance (1.4 miles), then the average packet from a terminal would originate at .7 mile range. A competing packet from an adjacent repeater would be 33 db above this average terminal packet. The total dynamic range obtained from near to far conditions is independent of repeater spacing since terminal to terminal minimum distance signal level is much greater than repeater-repeater signal levels. It is noted that terminal-terminal packet transfers will not be feasible but represent interfering signals which can force the linear portion of receiver into non-linear operation. One logical allocation of this overall dynamic range between power control and receiver AGC would be to normalize repeater to repeater signal levels to the average terminal packet signal level through power control. This would then require the power control dynamic range to be 33 db for the hypothetical repeater spacing. This would vary with repeater spacing at a rate of 6 db/octave. For example, if the repeaters were spaced $\sqrt{2}$ times the maximum terminal-repeater distance, then the power control range would be 30 db.

The overall dynamic range for a minimum distance between terminals of 50 feet is 84 db, and for the repeater spacing shown in Figure 3, the power control range would be 33 db leaving a minimum dynamic range of the receiver AGC of 51 db. Additional dynamic range would be necessary to handle pathological cases where the power control was stepped past the normalized level.

7.0 CONCLUSIONS

The conclusions that can be drawn from this analysis based on the assumed system parameters are the following:

1. The maximum spacing between repeater and terminal is in the order of one mile.
2. The overall dynamic range of signal levels is independent of repeater spacing and is dependent only on minimum distance between terminals and varies at 6 db/octave.
3. The overall dynamic range for minimum terminal distance of 50 feet is 84 db.
4. An average terminal packet signal level at a repeater will be 33 db below an adjacent repeater packet level.
5. One solution to the large variation between repeater signal levels and terminal signal levels would be to normalize repeater signal level through power control.

The tradeoff in the suggested power control appears to be in the facts that repeaters would not be able to "hop over" repeaters on their way to a station, but repeater packets would not dominate over terminal to repeater traffic. A quantitative comparison as to impact on network system performance (throughput and time delay) is necessary to evaluate this power control scheme.

The path loss study carried on at the University of Hawaii, Note #70, indicated that the path loss vs. distance had the same slope as Okumura's data; however, it was approximately 15 db less in actual loss at 413 MHz. If this was the case, then the maximum repeater-terminal link would be approximately 3.0 miles.

REFERENCE

1. Okumura, et. al., Field Strength And Its Variability In VHF and UHF Land-Mobile Radio Service, Review of the Electrical Communication Laboratory, Vol. 15, numbers 9-10, September-October, 1968.

Appendix C.5

Power Sources for Repeaters and Terminals

Power Sources for Repeaters and Terminals

Abstract: Portable power for a repeater or terminal cannot be provided in small lightweight packages. It is recommended that (1) the equipment be designed to operate from 24VDC, (2) that the power source be outboard from the equipment and (3) the power source selected from a range of options depending upon deployment (AC-DC; silver-zinc/zinc-air/nickel-cadmium batteries; thermoelectric-battery hybrid).

INTRODUCTION

The terminals and repeaters are intended to be mobile equipment and as much as possible self-contained including the primary power source. This note reviews many of the available power source options.

The terminal requires the most mobility; thus, size and weight are the primary considerations. The operational useage can be described in terms of hours with a requirement to replace or recharge the primary power source frequently. The repeater must operate for weeks of unattended and somewhat unpredictable useage. Although the repeater must still be mobile, the size and weight can be sacrificed for life expectancy, reliability and minimum maintenance. The cost aspect impacts on the selection of the source and its capability to be discarded or rejuvenated.

Radio Equipment Loading

It is not the intent of this paper to describe in detail the voltages and current drain of the terminal and repeater. This will be derived separately. Estimates range between 15 and 60 watts. The major power surge is required for the 10W RF amplifier which, if left on, would consume approximately 42 watts from a 24VDC source. The PA is not very sensitive to supply voltage with the major effect being a drop in output power. Figure 1 shows some typical data of PA output power change as a function of supply voltage variation. Since regulation the high surge currents are on the 24VDC source, it would be desireable to have a primary power system that is 24VDC. If the voltage were lower, DC to DC converters would be required. This introduces more losses and requires the primary power source to be larger. The issues of power distribution in the equipment, dc to dc converters and efficiencies will be described in a separate paper when the equipment power requirements are defined.

The reason for mentioning the power at this time is to bias our thoughts towards a battery that has a flat voltage discharge characteristic and a voltage of 24VDC.

Summary of Power Sources Not Considered

Motor Generators: If a conventional gasoline engine is employed, the size, weight and heat generated are undesireable for the terminal configuration. For the repeater, the reliability of a device with many moving parts is low and the maintenance requirement high. Because of the energy crisis, there is interest in wind power generators. This is not considered a reliable solution at this time. A wind powered device would only be applicable as a charging system for batteries. There may be potential for specific repeater locations.

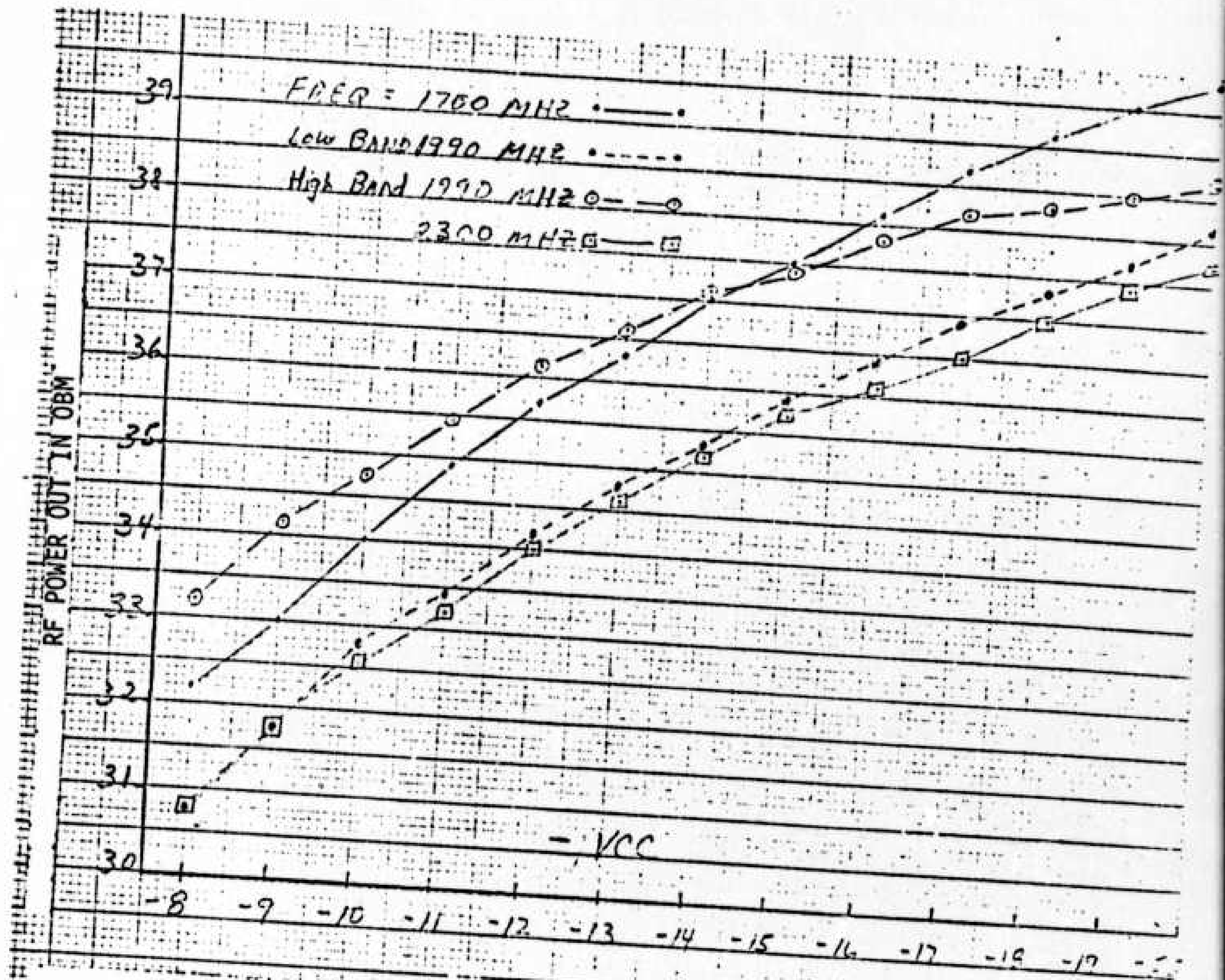


FIGURE 1 PA POWER OUTPUT CHANGE WITH SUPPLY VOLTAGE CHANGE

Fuel Cells: A review of several manufacturers indicates that the market is primarily oriented toward converting water to gas by using electricity rather than the reverse process. In general, the fuel cell systems have either turned out to be too complicated, too costly requiring large amounts of noble metals or dangerous due to fuel combinations. Fuel cells are not commercially available and there is no indication that there are advantages that warrant them to be placed on the market in the near future.

Solar Cells: For space flights, solar cells have proven to be a reliable source of power. Cells are commercially available that produce 1 watt of output for a 100 MW/CM^2 incident light in a package approximately 6 inches by 6 inches by 3/4 inch. For 50 watts of continuous power under optimum conditions, 13 sq. ft. of surface is required. This surface area presents a problem in windy environments. The power output is sensitive to shade, cloudy conditions, geographical location with respect to the sun's angle and seasonal length of the day. There promises to be significant research into solar cells in the next five years but for the near term, solar cells are not a viable option.

Power Sources Considered (General)

The traditional portable power source is battery. For the terminal, this appears to be the most logical choice. For a repeater, batteries supplemented with thermoelectric generators or solar cells are potential solutions.

Battery Characteristics

Some of the typical considerations in selecting a battery type are:

- Nominal voltage
- Current drain (average and surge)
- Operating cycle
- Life expectancy
- Recharging
- Service temperatures
- Size and weight

Only secondary cells (cells that can be recharged) are considered. Table 1 is a summary of several of some secondary cell parameters. The silver zinc cell has high energy density both in weight and size. Eagle Pitcher Industries builds a silver zinc 800AH, 24VDC (approx. 19KW Hr) battery for satellites. The battery dimensions are 11.3 in. by 13.5 in. by 9.0 in., weighs 170 pounds (costs approximately $17,000). The zinc-air battery is also a unique battery with high energy density. Recharging the cell consists of replacing the zinc plates. If the equipment need not be left on, the battery can be deactivated. This has potential for the terminal equipment. The self discharge characteristic of the cells is of primary concern for a repeater. Even if there is no load, the cell will drain itself. As seen from Table 1, only silver-zinc, silver cadmium or lead acid will hold a charge for six months or longer. The life expectancy is a measure of when the cell must be salvaged and rebuilt or disposed of. Variations of cell capacity with temperature is also a factor. Figure 2 shows the variation of some of the cells as a function of temperature. The figure shows the lead acid cell to suffer from loss of capacity at lower temperatures. If the lead acid cell is left in a discharged condition, it could suphate, requiring that the battery be salvaged. Another factor in selecting batteries or cells is the voltage variation as the cell discharges. Nickel cadmium and zinc air have very flat voltage characteristics. Figure 3 is a typical discharge characteristic of the batteries considered. The zinc air has a flat voltage during discharge which allows this type of battery to power circuitry directly without regulators. The lead acid battery voltage changes considerably during discharge and could not be a long term power source without a regulating device.

As a reference for anticipating battery size and weight for average power drain and the length of operation required, Figures 4 and 5 are presented. Figure 4 is aimed primarily at a hand held or suitcase terminal where Figure 5 is for long term operation as for a repeater. In both cases, the battery weight is described in terms of the maximum and minimum energy density devices. The size is also described in two scales representing the range of values depending upon selection of a battery. Note should be made of the battery weight in terms of tons. This corresponds to a lead acid battery system.

TABLE 1: SECONDARY BATTERY CELLS

TYPE	ABBR	OPEN CKT V	WORKING VOLTAGE	ENERGY DENSITY WHR/lb	ENERGY DENSITY WHR/Cu.In.	RECHARGE CYCLES	SELF DISCHG. @ 70°C	LIFE EXPECTANCY
Silver Zinc	Ag/Zn	1.6-1.8	1.3-1.6	25-95	1.3-6.8	500	1 yr	1 yr. wet/2-5 yrs. dr
Silver Cadmium	Ag/Cd	1.4	0.8-1.1	11-33	0.6-2.8	200-3K	6 mo.	1 to 5 yrs. wet/5 yrs
Nickel Cadmium	Ni/Cd	1.3	1.0-1.2	8-17	0.5-1.9	> 1K	3 mo.	> 5 yrs.
Zinc Air	Zn	1.4-1.5	1.2-1.3	25-100	3.0-8.0	50	36 hr.	36 hr.active/5 yr. sh
Lead Acid	Pb	2.12	1.5-2.0	10-15	0.7-1.4	> 1K	1 yr.	3-5 yrs. wet/5 yrs. d
Nickel Zinc	Ni/Zn	1.8	1.5-1.6	20	1.4	150	1 mo.	> 1 yr.

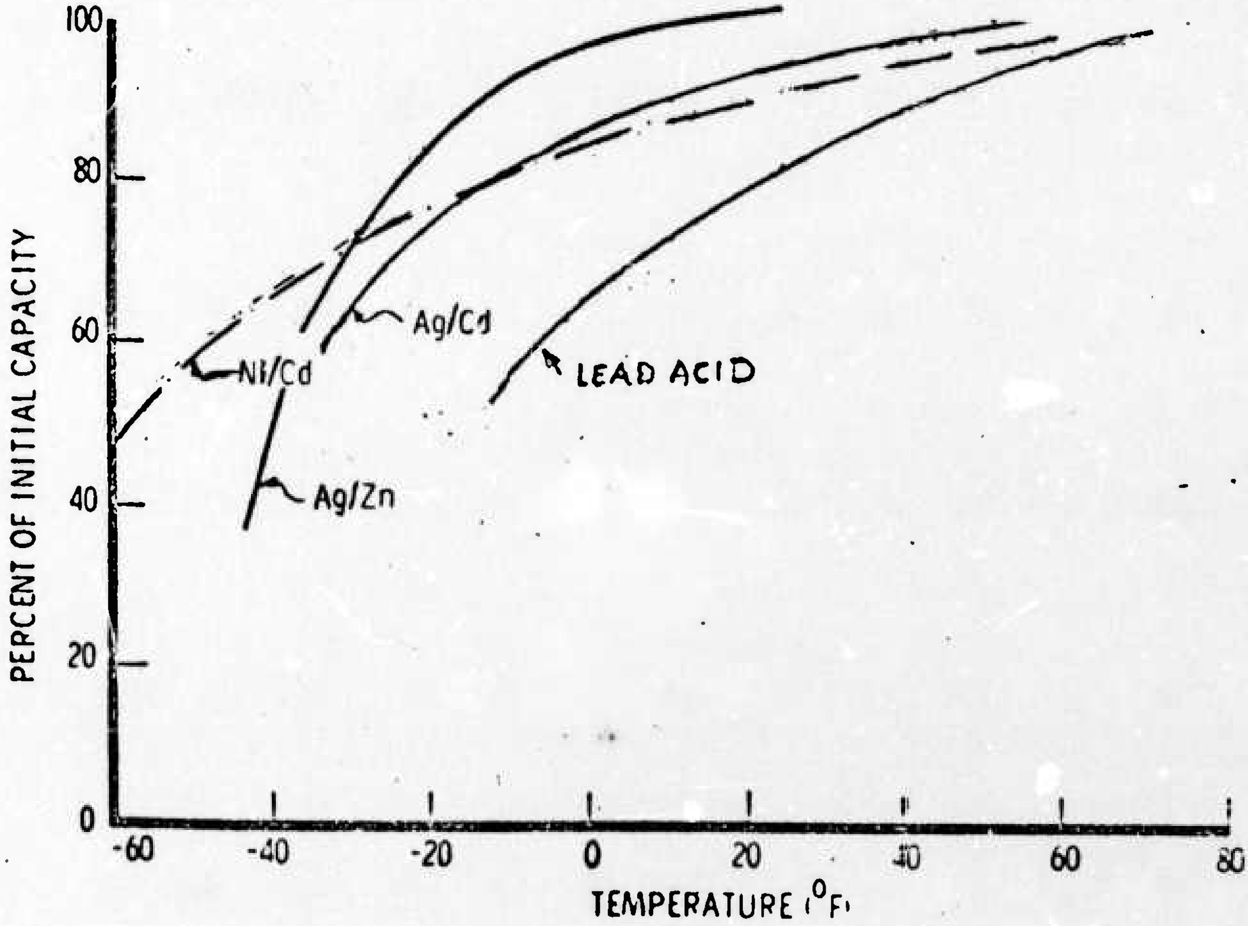
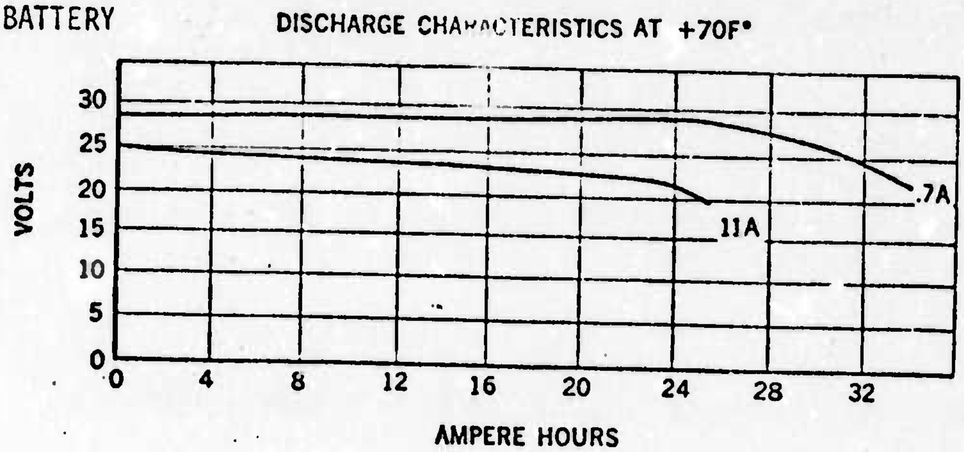
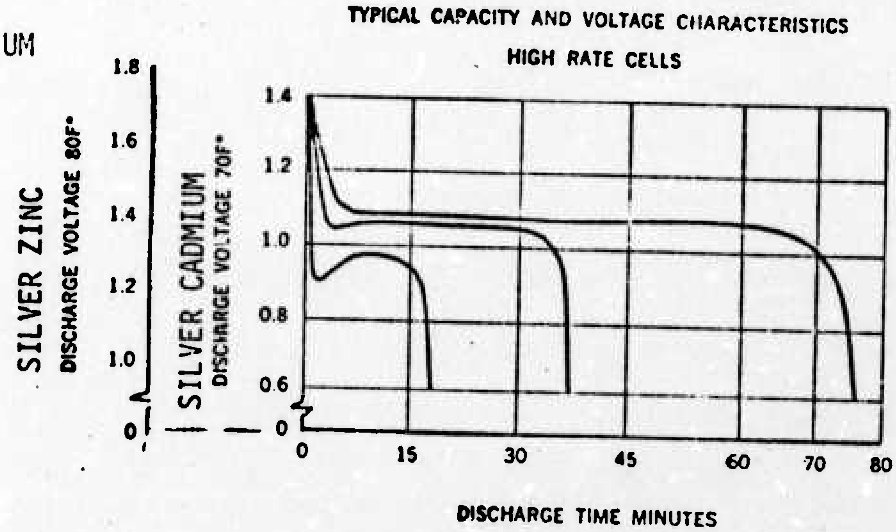


FIGURE 2 : CAPACITY RETENTION OF ELECTRO CHEMICAL CELLS DURING LOW DISCHARGE RATES AS A FUNCTION OF TEMPERATURE.

ZINC AIR BATTERY



SILVER ZINC
SILVER CADMIUM



NICKEL CADIUM

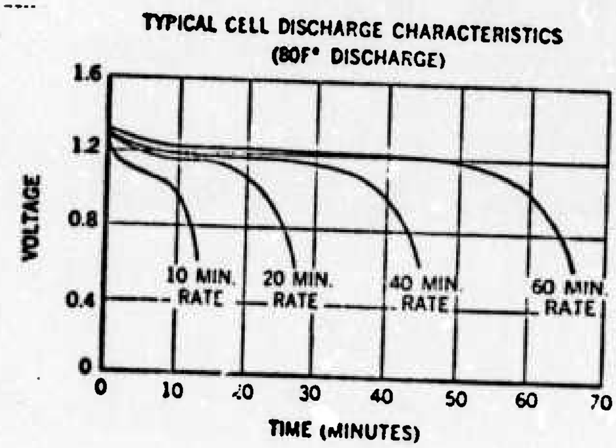
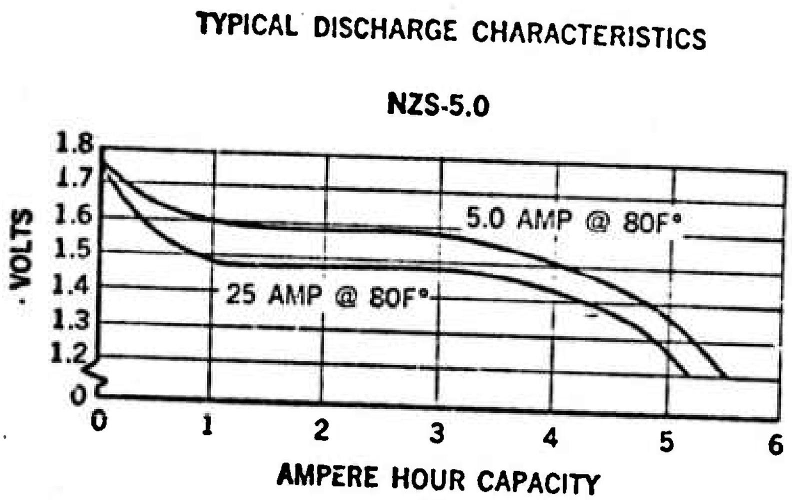


FIGURE 3: BATTERY DISCHARGE CHARACTERISTICS

NICKEL ZINC



LEAD ACID

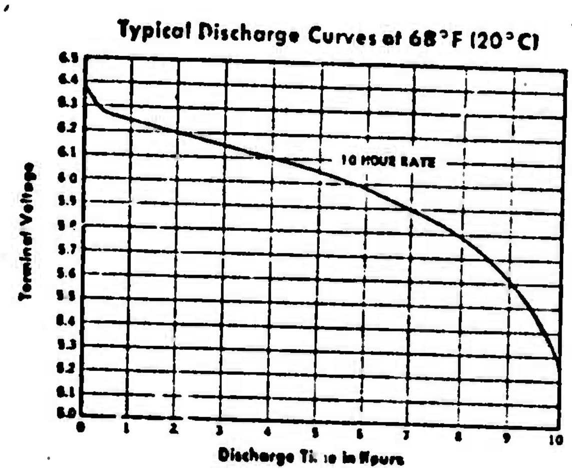


FIGURE 3 (CONT'D): BATTERY DISCHARGE CHARACTERISTICS

FIGURE 4

BATTERY CAPACITY - WEIGHT - SIZE

VS. DURATION OF OPERATION

(SHORT TERM)

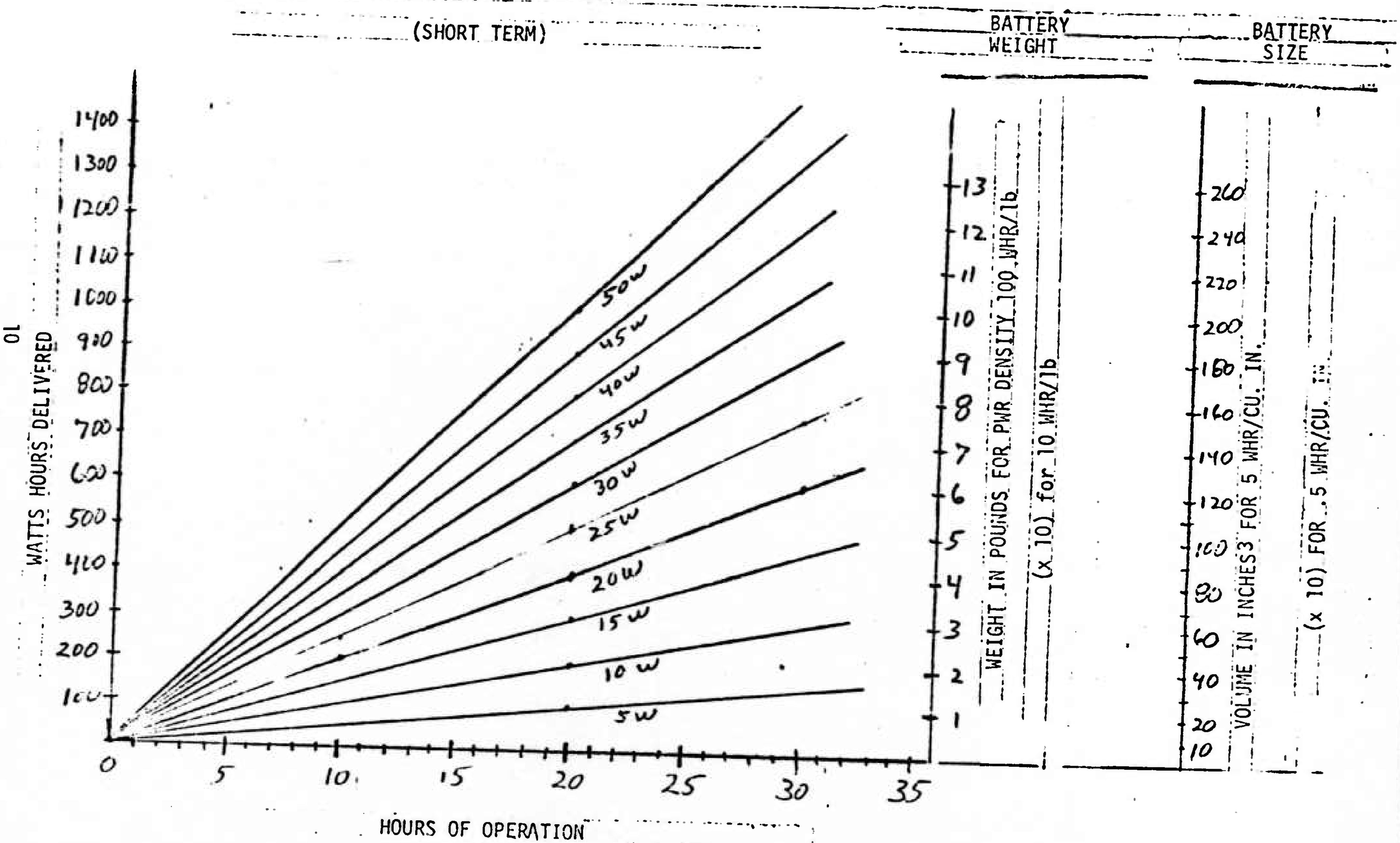
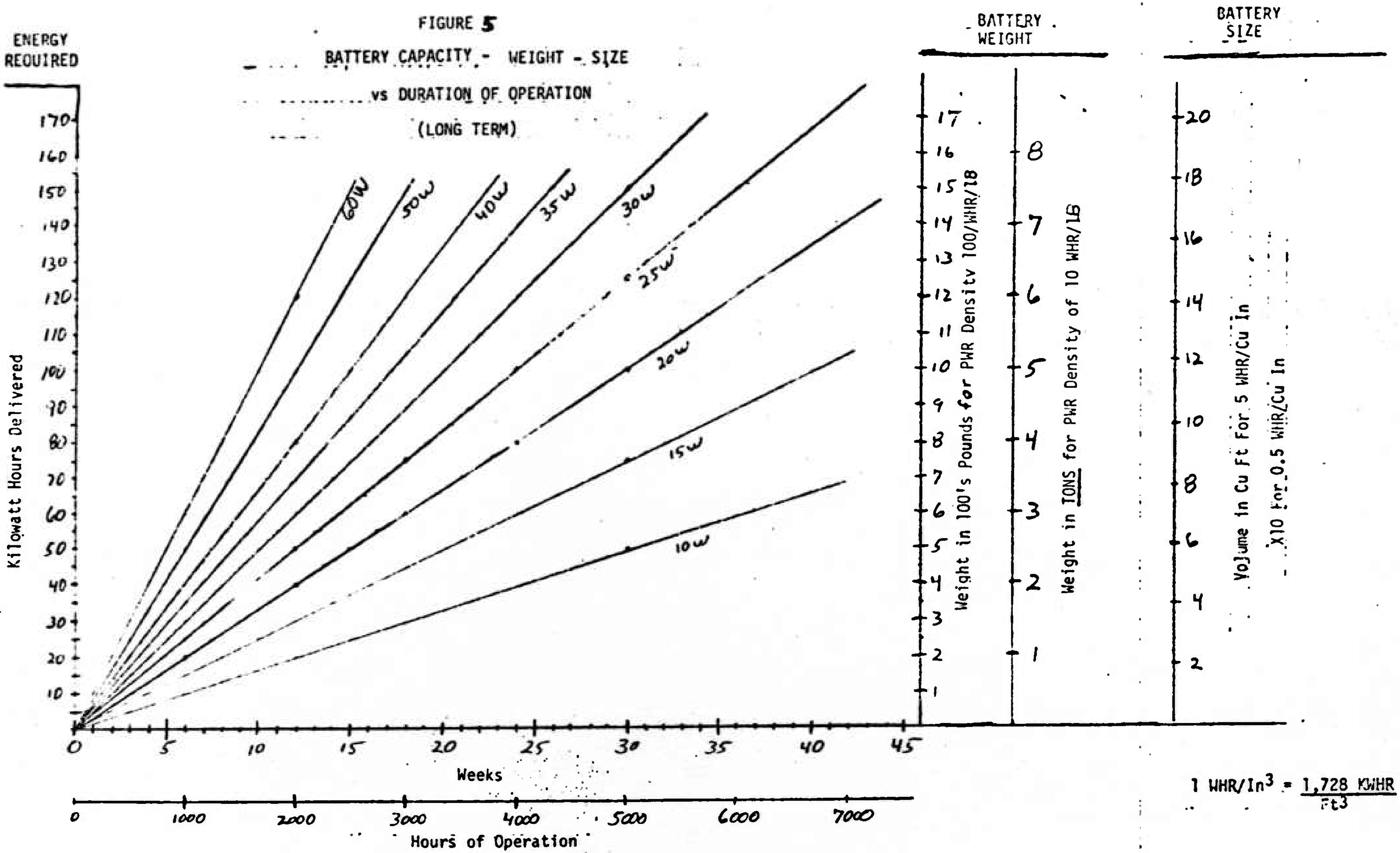


FIGURE 5

BATTERY CAPACITY - WEIGHT - SIZE

vs DURATION OF OPERATION

(LONG TERM)



$1\ WHR/In^3 = \frac{1{,}728\ KWHR}{Ft^3}$

Thermoelectric Generators

Thermoelectric generators are heat engines which convert heat energy to electrical energy. The electrical energy is obtained directly by causing a temperature difference across a thermopile consisting of pairs of positive and negative thermocouples. The thermocouple utilizes the Sebeck effect. The larger the difference between hot and cold junctions of the thermocouple, the larger the generated voltage. No moving parts are required to obtain power. This type of generator is commercially available from several companies. The heat source can be nuclear, gas (propane, butane, or natural gas) or liquid fuels. Because of the simplicity of the device, the life expectancy is long (greater than 10 years), maintenance is minimum, usually only being an annual cleaning of an air filter and gas jet, and reliability is thus high. The main considerations are:

Power requirement
Size and weight (generator and fuel)
Fuel consumption
Type of heat source
Environmental considerations
Failure modes

Several nuclear powered generators have been built for under-water, polar wastelands or spacecraft applications. Their reliability has been proven by systems operating in excess of five years. The cost and testing required by the Automic Energy Commission make this option only applicable for those situations where extremely long life and reliability are required.

Three gas powered thermoelectric manufacturers are compared on Table 2. Propane was selected as the fuel for all units because of its boiling point of -43°F. Butane, on the other hand, boils at +31°F, which is not low enough if the system is to be deployed in cold climates. All the generators were selected from manufacturers' brochures for 50 to 60 watts of power at 24VDC. The Teledyne unit has a flame-

TABLE 2: THERMOELECTRIC GENERATORS

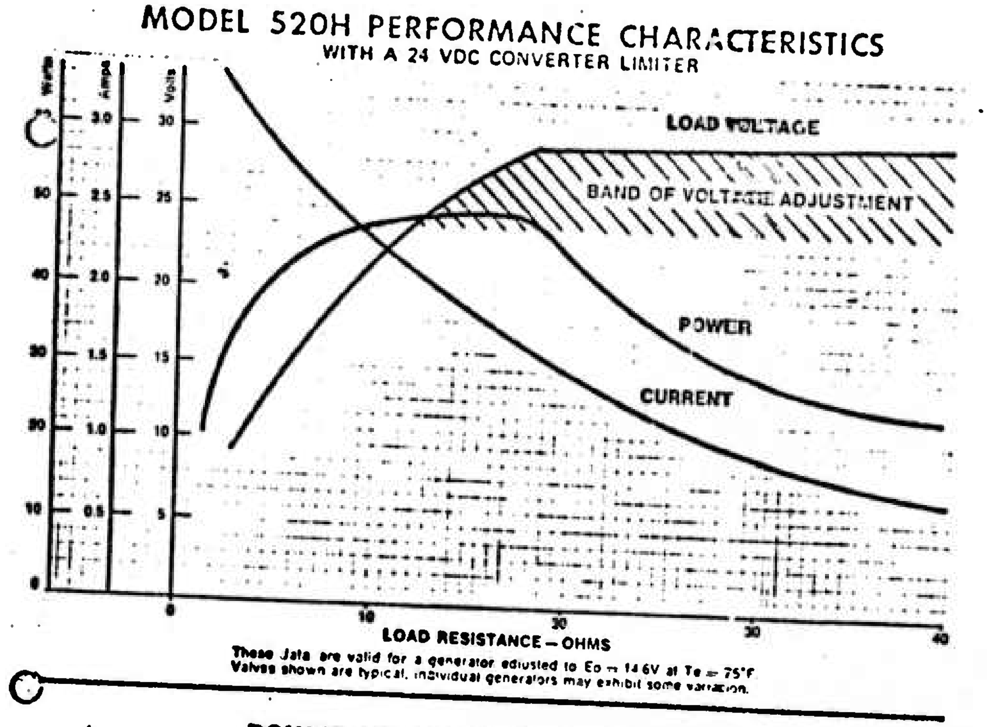
OUTPUT VOLTAGE +24VDC

MFGR	MODEL	COST ($)	OUTPUT POWER (WATTS)	PROPANE LBS/WK	UNIT WEIGHT (POUNDS)	DIMENSIONS (INCHES)	BLOWER
THERMO ENERGY CONVERTER CORP.	2A80L-24	2K	53	39.8	76	24 x 24 x 13 1/4	Yes
3M	520H-24	3K	50	45.4	90	37.5 x 20 x 17	No
TELEDYNE	2T6	-	60	67.2	130	30 x 25 x 17	No

less heating chamber. It thus avoids being extinguished by high winds as do the other two manufacturers that have a flame. A re-ignition capability is provided for the flame type units to overcome this failure mode. The flame is contained within several layers of housings so that rather high winds, in excess of 70 mph, are required to affect the chamber. The units require oxygen for burning and are affected by altitude. Operation from below sea level to 13,000 feet can be obtained with standard units. Figure 6 shows some of the performance parameters of the 3M Company device. As would be expected, the power output is a function of ambient temperature. The power and voltage characteristics as a function of load resistance show that the source impedance is relatively high, greater than 1 ohm. Efficient use of a thermoelectric generator requires that the load be matched to the device. For this reason, a hybrid system using batteries to absorb surges is normally used. The low impedance voltage regulation provided by a battery complements the thermoelectric generator.

The major factor not addressed is the storage space and weight required for fuel as a function of the time duration for unattended operation.

Figure 7 shows the weight of propane fuel as a function of manufacturer and duration of operation required. Six months of operation would require 1050 pounds of propane @ 40 lbs/wk or 1800 pounds @ 70 lbs/wk. This does not include a container for the propane. As a general rule, the tank weight is approximately 75% of the fuel capacity weight. Thus, to store 1000 lbs of propane would require a tank weighing 750 lbs. As a specific example, a commercially available tank weighing 300 lbs will hold 420 lbs (99 gallons) of propane. Its size is 29" in diameter and 56 1/4" high. Three to four tanks would be required for 6 months of operation.



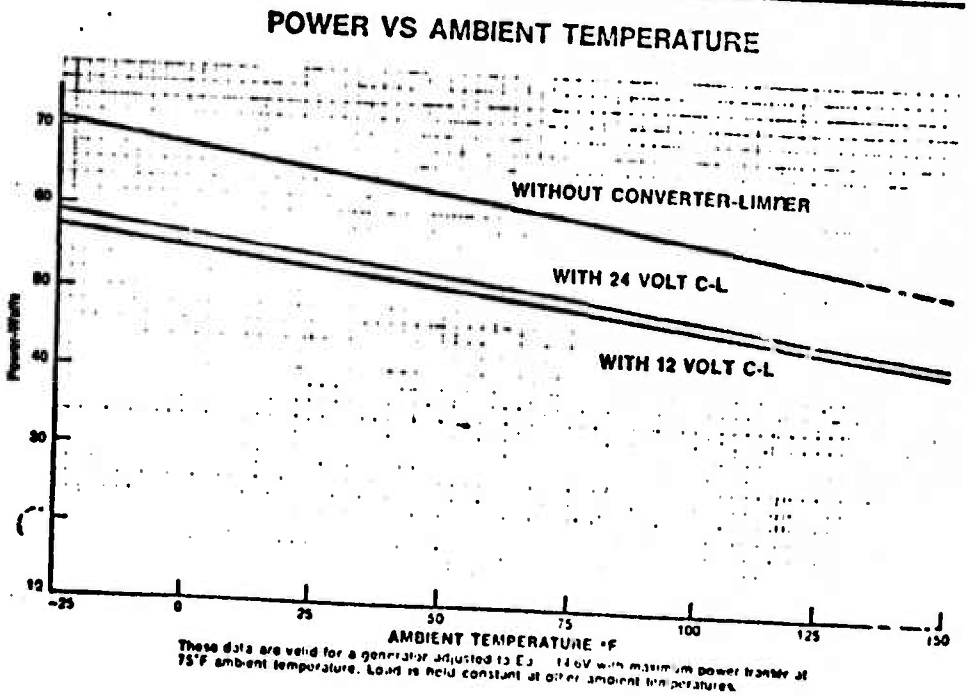


FIGURE 6: THERMOELECTRIC SOURCE CHARACTERISTICS

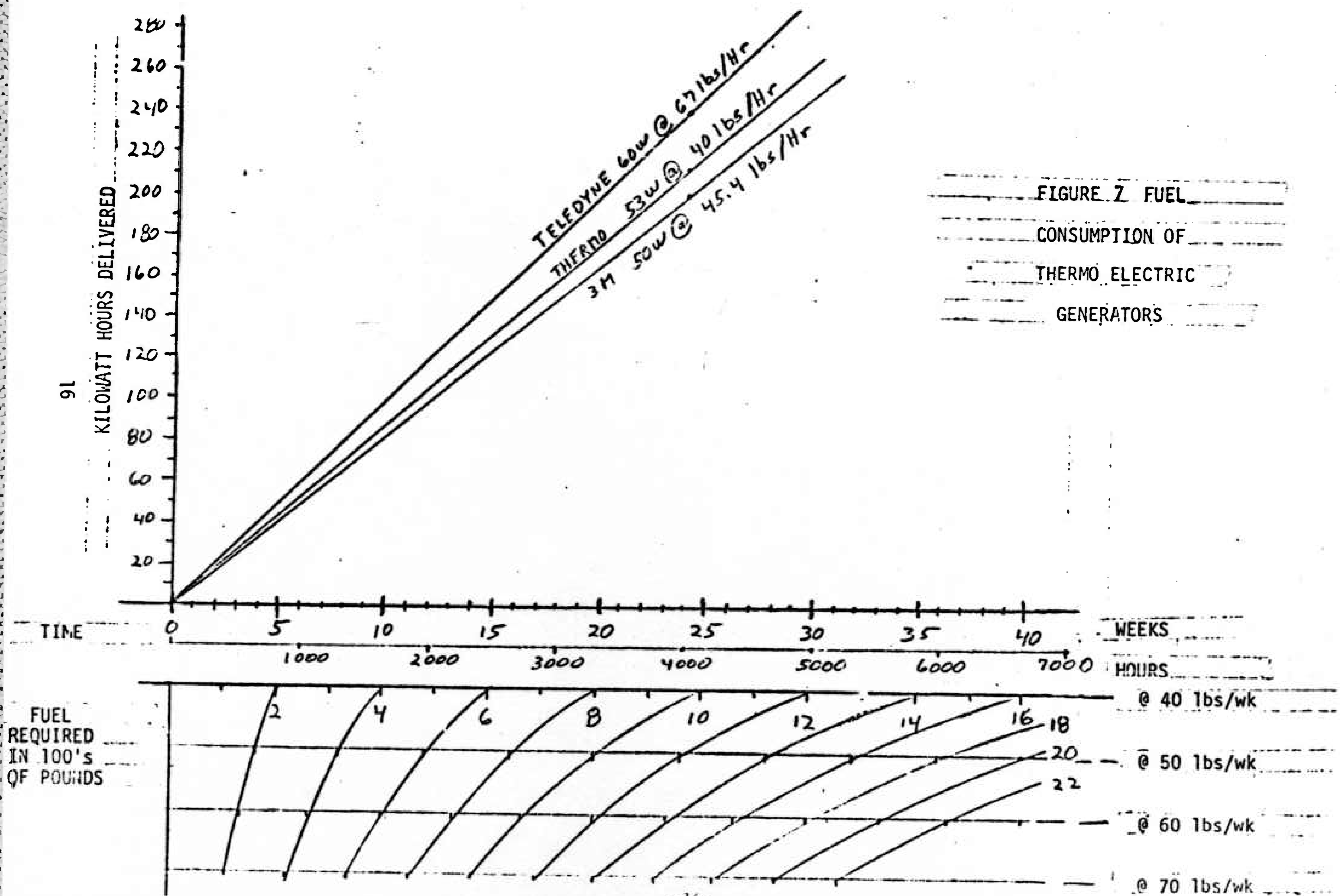


FIGURE 7 FUEL CONSUMPTION OF THERMO ELECTRIC GENERATORS

Recommended Approach

It appears obvious that a terminal or repeater cannot initially be constructed to be lightweight and compact if the power system is integrated into the unit. The equipment should be designed to operate from 24VDC which will allow the system to be powered in many ways. For general use, a conventional AC-DC power supply; for short term useage in an area where power is not available, a pair of 12VDC lead acid cells can be quickly and easily obtained. Where this is a continuing service with easy access but not a permanent installation, higher capacity Nickel Cadmium cells can be used and recharged. For permanent unattended installations, a nuclear or propane thermoelectric generator and nickel cadmium battery combination is recommended. (The nuclear system will be more mobile). A propane system must be considered a semi-fixed site and might be well to consider a shelter construction as a long term solution. For the terminal, an outboard battery pack of nickel cadmium or silver zinc is recommended. A zinc air system is an option where prolonged field use is required. A 24VDC, 25AH battery weighing 12 pounds can be deployed. When not in service, the plates can be removed, deactivating the cell. Recharging the system requires about 1 pound of zinc plates. Thus, carrying 15 pounds to the field would provide 100AH of service at 24VDC (2400 WHRS) or 2 days of continuous operation of a 50 watt device.

References

1. City Car With H_2-Air Fuel Cell/Lead Battery. by K. V. Kordesch, Consumer Products Division of Union Carbide Corporation; 1971 Intersociety Energy Conversion Engineering Conference Proceedings, page 103.

2. A Review of Cadmium-Mercuric Oxide Batteries for Underwater Applications by Roberg G. Barnhart and David P. Boden ESB Incorporated; 1971 Intersociety Energy Conversion Engineering Conference Proceedings, page 529.

3. Eveready Battery Applications Engineering Data Union Carbide Corporation 1971.

4. Eagle Pitcher Industries, Incorporated; Electronics Division Couples Dept.; Joplin, Missouri.

5. 3M Company; St. Paul, Minnesota.

6. Thermo-Energy Converters Corporation; Brooklyn, New York.

7. Teledyne Isotopes; Timonium, Maryland.

Appendix C.6

Criteria of Code Selection in MSK Sense

1.0 GENERAL

The MSK (Minimum Shift Keying) type of modulation utilizes a single physical channel consisting of a pair of equivalent quadrature channels. The practical implementation of MSK essentially involves transmitting two separate signals one chip (T_c) apart (assuming code spread) on the individual quadrature subchannel. Each signal is cosine weighted* over every two-chip interval on each subchannel. The resultant signal is phase continuous at each transition.

The MSK signal can either utilize a single code for each bit with alternating chips transmitted on the two orthogonal subchannels, or it can utilize two independent 1 bit length codes staggered by one chip interval on individual channels. In order that the matched filters (SAWD's) may be implemented for detection of signal corresponding to each subchannel, it is recommended that two codes of the same general class of family (e.g., maximal length, Golay's complementary, etc.) be used for the two subchannels. This ensures that the inherent autocorrelation properties of each code class will be undisturbed.

*Several different weighting functions can be used depending on the desired signal structure.

For example, a maximal length code results into -1 side lobes when sequentially repeated. This would not occur with one maximal length code with its alternate chips transmitted on two sub-channels and detected through independent matched filters on each subchannel. Similarly, in order to retain the property of Golay's complementary sequences, it is desirable that a complementary pair of codes be transmitted on the two subchannels, so that the sum of the autocorrelation side lobes is zero for all non-zero shifts.

Unless an additional constraint can be implemented upon the signal design, there is a cross-coupling between the two subchannels. One purpose of this paper is to identify mathematically the time and phase relationship between the various autocorrelating and cross-correlating terms in the MSK application, so they can be combined properly to indicate the output of the detector at each chip interval. This relationship is derived in Section 2.0.

Based on the time relationship between the autocorrelation and cross-correlation terms at any chip instant, the criteria for selecting the codes are outlined. The impact of dual data rate on the code selection is also considered.

2.0 A MATHEMATICAL MODEL OF MSK MODEM

Figure 1 depicts a model of MSK modem that has been used to establish time and phase relationship between the autocorrelating and cross-correlating terms. The effect of the cosine weighting function is disregarded in this model because the results of the model are to be used only to develop a time correlating technique on a chip-by-chip basis. However, since the auto-correlation or cross-correlation between individual subchannels

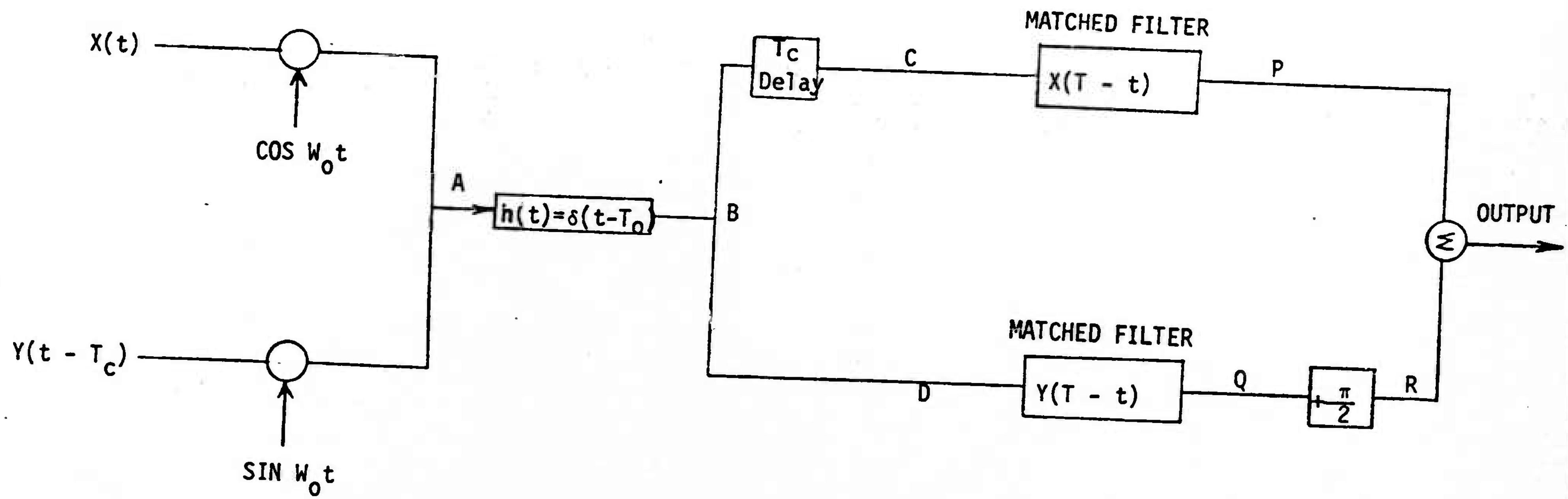


Figure 1

NOTE: Refer to Appendix 1 for Expressions at A, B, C, D, P, Q, R, etc.

(in digital sense) occurs over 2 chip intervals, the in-between chip correlation values can be obtained by accounting for the cosine weighting function; that is, a correlation factor can be determined for a 2-chip wide cosine weighting function for 1 chip shift, as shown in Figure 2. Thus, we are able to perform a digital chip-by-chip correlation using a digital computer program.

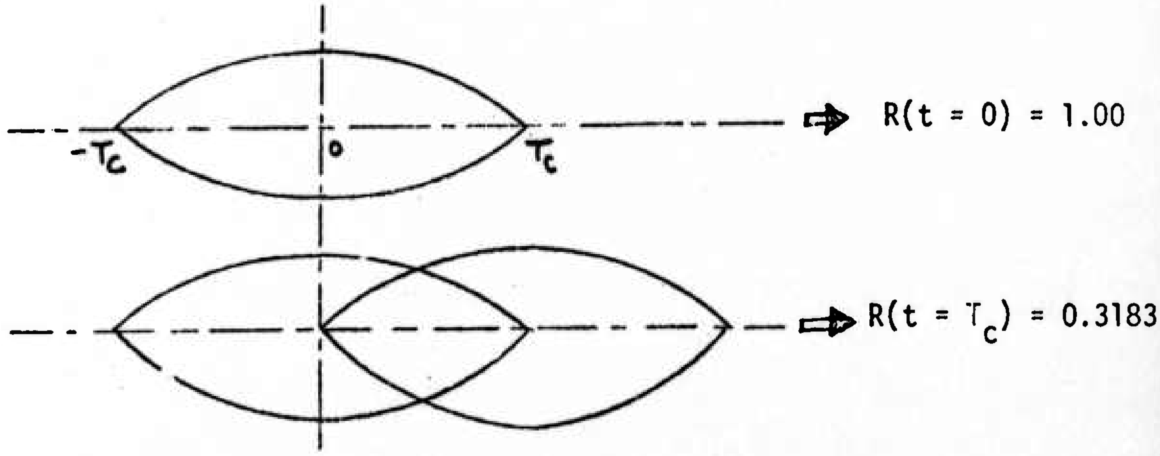


Figure 2

As obvious from Figure 1, once the input coded waveforms X(t), and Y(t - T_c) are sent through the orthogonal subchannels, and subjected to proper time delay, matched filtering, and phase translation, the resultant output appears in terms of four correlation components: R_{xx} and R_{yy} due to autocorrelations, and R_{xy} and R_{yx} due to cross-correlation. The equation relating these four components at the output as functions of time and phase is derived in Appendix 1, and given as follows:

$$\text{Output} = \left[R_{xx}(t - T_c) + R_{yy}(t - T_c)\right] \cos\alpha + \left[R_{yx}(t - 2T_c) - R_{xy}(t)\right] \sin\alpha$$

Where: R_{xx} = Autocorrelation of code X
R_{yy} = Autocorrelation of code Y
R_{yx} = Cross-correlation of Y with X
R_{xy} = Cross-correlation of X with Y

T_c = One chip interval
α = $W_o(t - T_o)$
W_o = $2\pi f_o$
T_o = Delay in the channel
f_o = Integer multiple of chip frequency

$$\left|\text{Output}\right| = \sqrt{\left[R_{xx}(t - T_c) + R_{yy}(t - T_c)\right]^2 + \left[R_{yx}(t - 2T_c) - R_{yy}(t)\right]^2}$$

2.1 Computer Program:

A computer program (see Appendix 2) has been written based on the above model. The program provides a tool to compare the side lobes due to various code pairs. An example that will explain the technique used in the program is given as follows:

Let X = 1 1 1 0
    Y = 1 1 0 1

Then, the correlation terms every two chips are:

R_{xx} =	-1	0	1	4	1	0	-1
R_{xy} =	-1	0	3	0	1	0	1
R_{yy} =	1	1	-1	4	-1	0	1
R_{yx} =	1	0	1	0	3	0	-1

Using the correlation factor of 0.3183 discussed in Section 2, the in-between chip correlation terms can be filled in to give the chip-by-chip correlation given as follows:

R_{xx} =	-.3	-1	-.3	0	.3	1	1.5	4	1.5	1	.3	0	-.3	-1
R_{xy} =	-.3	-1	-.3	0	.9	3	.9	0	.3	1	.3	0	.3	1
R_{yy} =	.3	1	.3	0	-.3	-1	.9	4	.9	-1	-.3	0	.3	1
R_{yx} =	.3	1	.3	0	.3	1	.3	0	.9	3	.9	0	-.3	-1

Now, according to the output equation of Section 2, R_{xx} and R_{yy} terms are slided to the right by 1 chip, and R_{yx} terms are slided by two chip, shown as follows:

R_{xx} =	X	-.3	-1	-.3	0	.3	1	1.5	4	1.5	1	.3	0	-.3	-1	-.3	0
R_{xy} =	-.3	-1	-.3	0	.9	3	.9	0	.3	1	.3	0	.3	1	.3	0	0
R_{yy} =	X	.3	1	.3	0	-.3	-1	.9	4	.9	-1	-.3	0	.3	1	.3	0
R_{yx} =	X	X	.3	1	.3	0	.3	1	.3	0	.9	3	.9	0	-.3	-1	-.3

Then,

$(R_{xx} + R_{yy})$ =	X	0	0	0	0	0	0	2.4	8	2.4	0	0	0	0	0	0	0
$(R_{yx} - R_{xy})$ =	.3	1	.6	1	-.6	-3	-.6	1	0	-1	.6	3	.6	-1	-.6	-1	-.3
Output =	.3	1	.6	1	-.6	-3	-.6	2.8	8	2.8	.6	3	.6	-1	-.6	-1	-.3

Notice that the autocorrelation in MSK is four chip wide.

Two more test examples are listed (as the program outputs) in Appendix 2. Based on this technique several code pairs consisting of maximal length class, or Golay's sequences will be studied to select the desirable code pairs. The criteria of their selection are outlined in the following section.

3.0 CRITERIA OF CODE SELECTION

Based on the result of Section 2.0, the following ground rules are applicable for selecting code pairs for M.S.K. applications:

A. Ideally, the absolute value of side lobes as a result of autocorrelation and cross-correlation of two codes should go to zero taking into account the proper time shifts described in the above section. Practically, however, they should be minimized.

B. For Golay's pairs, specifically, the only important criterion is that the difference of the cross-correlation terms staggered by 2 chips should result in minimum side lobes. This is because the side lobes due to autocorrelation are always zero.

C. With maximal length code pairs, the property that the autocorrelation side lobes are always -1 after the code is sequentially repeated is ruined due to the cross-correlation side lobe. However, this should not have a significant impact in the non-overlapping packet environment once we could attain synch to a starting pulse. Therefore, the main consideration with maximal length code pairs for non-overlapping packets is to minimize the side lobes during the time the first bit is propagated through the matched filters. During this interval, it is the overall impact of the output equation of Section 2.0 that has to be accounted for in minimizing the side lobes.

D. With maximal length code pairs in the overlapping packet environment (when multiple detectors are used) the side lobes have to be minimized not only during the first bit interval but during all bit intervals so that the sum of the correlations due to the interfering packets is minimum.

E. Since dual data rate is to be used in the ARPA system, it is important to take into account the cross-correlation between Hi-Lo rate codes. For example, if 128 chips/bit complementary sequence is used for the low rate, and 32 chips/bit complementary is used for the high rate, then it is important to determine in advance that the 32 chip SAWD will not result in high cross-correlations when 128 chip code is propagated through, and vice versa.

APPENDIX 1

Refer to Figure 1 in Section 2.0 of this paper.

$A = X(t)\cos W_0 t + Y(t - T_c) \sin W_0 t$

$B = X(t - T_0) \cos W_0(t - T_0) + Y(t - T_c - T_0) \sin W_0(t - T_0)$

$D = B$

$C = X(t - T_0 - T_c) \cos W_0(t - T_0 - T_c) + Y(t - 2T_c - T_0) \sin W_0(t - T_0 - T_c)$

Now let $W_0 = nW_c$ where W_c is the chip frequency.

$$= \frac{2\pi n}{T_c}$$

$$\therefore \quad \cos W_0(t - T_0 = T_c) = \cos \frac{2\pi n}{T_c} (t - T_0 - T_c)$$

$$= \left\{\cos \frac{2\pi n(t - T_0)}{T_c} - 2\pi n\right\}$$

$$= \left\{\cos 2\pi n - \frac{2\pi n(t - T_0)}{T_c}\right\}$$

$$= \cos\left\{\frac{2\pi n(t - T_0)}{T_c}\right\} \quad \text{for all integer values of } n$$

$$= \cos W_0(t - T_0)$$

$$= \cos\alpha$$

Similarly, $\sin W_0(t - T_0 - T_c) = \sin W_0(t - T_0)$ for all n.

$$= \sin\alpha$$

Hence, $C = X(t - T_o - T_c) \cos\alpha + Y(t - 2T_c - T_o) \sin\alpha$

$D = X(t - T_o) \cos\alpha + Y(t - T_c - T_o) \sin\alpha$

Then, through matched filters, we get

$$P = R_{xx}(t - T_o - T_c - T + t) \cos\alpha + R_{yx}(t - 2T_c - T_o - T + t) \sin\alpha$$

$$Q = R_{xy}(t - T_o - T + t)\cos\alpha + R_{yy}(t - T_c - T_o - T + t) \sin\alpha$$

$$R = R_{xy}(t - T_o - T + t) \cos(\alpha + \pi/2) + R_{yy}(t - T_c - T_o - T + t) \sin(\alpha + \pi/2)$$

$$= R_{yy}(t - T_c - T_o - T + t)\cos\alpha - R_{xy}(t - T_o - T + t) \sin\alpha$$

$$\begin{aligned} \text{Output} &= P + R \\ &= \left\{R_{xx}(t - T_o - T_c - T + t) + R_{yy}(t - T_c - T_o - T + t)\right\} \cos\alpha \\ &\quad + \left\{R_{yx}(t - 2T_c - T_o - T + t) - R_{xy}(t - T_o - T + t)\right\} \sin\alpha \\ &= \left\{R_{xx}(t - T_c) + R_{yy}(t - T_c)\right\} \cos\alpha + \left\{R_{yx}(t - 2T_c) - R_{xy}(t)\right\} \sin\alpha \end{aligned}$$

APPENDIX 2

This appendix lists a program that calls for a subroutine "COREL" listed in Packet Radio Note #73. Two examples are presented at the end. First example is for a 7-length maximal length code pair, and the second example is for an 8-length Golay's complementary pair.

```
      DIMENSION CX(500),CY(500),CXBAR( [illegible]0),CYBAR(500),XXX(1000),XXY(1000
     1),YYY(1000),YYX(1000),AAA(1000),  B(1000),BBB(1000),BBA(1000),
     2[illegible]FCRS(1000),CORACT(1000),ENV(10  )
      COMMON [illegible](200),CP(500),NR(400)
    5 READ(5,10,END=2000) N,(CX(I),I=1  )
      READ(5,10) N,(CY(I),I=1,N)
   10 FORMAT(I5/(16F5.0))
      WRITE(7,20)
   20 FORMAT(1H1,'   CODE X',/)
      WRITE(7,25) (CX(I),I=1,N)
   25 FORMAT(20F6.0)
      WRITE(7,30)
   30 FORMAT(1X//)
      WRITE(7,35)
   35 FORMAT(1X,'   CODE Y',/)
      WRITE(7,40) (CY(I),I=1,N)
```

```
      FORMAT(20F8.0)
      WRITE(7,30)
      DO 50 I=1,N
 THE TEST FOR EQUALITY BETWEEN NON-I  EGERS MAY NOT BE MEANINGFUL.
      IF(CX(I).EQ.0.) GO TO 45
      CXBAR(I)=0.
      GO TO 50
   45 CXBAR(I)=1.
   50 CONTINUE
      DO 60 I=1,N
 THE TEST FOR EQUALITY BETWEEN NON-I  EGERS MAY NOT BE MEANINGFUL.
      IF(CY(I).EQ.0.) GO TO 55
      CYBAR(I)=0.
      GO TO 60
   55 CYBAR(I)=1.
   60 CONTINUE
      NP=N
      DO 65 I=1,N
      CP(I)=CX(I)
      CP(I+N)=CX(I)
   65 C(I)=CX(I)
      WRITE(7,70)
   70 FORMAT(1X,'     XX/X CORRELATION',
      CALL CORL
      M=3*N-1
      DO 75 I=1,M
   75 XXX(I)=XX(I)
      DO 80 I=1,N
   80 C(I)=CY(I)
      WRITE(7,85)
   85 FORMAT(1X,'     XX/Y CORRELATION',
      CALL CORL
      DO 90 I=1,M
   90 XXY(I)=XX(I)
      DO 95 I=1,N
      CP(I)=CY(I)
   95 CP(I+N)=CY(I)
      WRITE(7,100)
  100 FORMAT(1X,'     YY/Y CORRELATION',
      CALL CORL
      DO 105 I=1,M
  105 YYY(I)=XX(I)
      DO 110 I=1,N
  110 C(I)=CX(I)
      WRITE(7,115)
  115 FORMAT(1X,'     YY/X CORRELATION',
      CALL CORL
      DO 120 I=1,M
  120 YYX(I)=XX(I)
      DO 130 I=1,M
      AAA(2*I)=XXX(I)
      ABB(2*I)=XXY(I)
      BBB(2*I)=YYY(I)
  130 BBA(2*I)=YYX(I)
      P=0.5
      AAA(1)=P*AAA(2)
      ABB(1)=P*ABB(2)
```

```
      BBB(1)=F*BBB(2)
      BBA(1)=F*BBA(2)
      MM=N-1
      DO 140 I=1,M
      J=2*I
      AAA(J+1)=F*(AAA(J)+AAA(J+2))
      AAB(J+1)=F*(AAB(J)+AAB(J+2))
      BBB(J+1)=F*(BBB(J)+BBB(J+2))
  140 BBA(J+1)=F*(BBA(J)+BBA(J+2))
      MMM=2*N
      DO 150 I=1,MMM
      XXX(I+1)=AAA(I)
      XXY(I)=AAB(I)
      YYY(I+1)=BBB(I)
  150 YYX(I+2)=BBA(I)
      XXX(1)=0.
      YYY(1)=0.
      YYX(1)=0.
      YYX(2)=0.
      XXY(MMM+1)=F*XXY(MMM)
      XXY(MMM+2)=0.
      XXX(MMM+2)=F*XXX(MMM+1)
      YYY(MMM+2)=F*YYY(MMM+1)
      YYX(MMM+3)=F*YYX(MMM+2)
      XXX(MMM+3)=0.
      YYY(MMM+3)=0.
      XXY(MMM+3)=0.
      K=MMM+3
      WRITE(7,200)(XXX(I),I=1,K)
      WRITE(7,210)
      WRITE(7,200)(XXY(I),I=1,K)
      WRITE(7,210)
      WRITE(7,200)(YYY(I),I=1,K)
      WRITE(7,210)
      WRITE(7,200)(YYX(I),I=1,K)
      WRITE(7,210)
  200 FORMAT(20F6.1)
  210 FORMAT(1X,/)
      DO 215 I=1,K
      DIFCRS(I)=YYX(I)-XXY(I)
  215 SUMAUT(I)=XXX(I)+YYY(I)
      WRITE(7,220)
  220 FORMAT(1X,'    DIF OF CROSSCORREL  ION TERMS',/)
      WRITE(7,200)(DIFCRS(I),I=1,K)
      WRITE(7,210)
      WRITE(7,230)
  230 FORMAT(1X,'    SUM OF AUTOCORRELA  ON TERMS',/)
      WRITE(7,200)(SUMAUT(I),I=1,K)
      WRITE(7,210)
      DO 300 I=1,K
  300 ENV(I)=SQRT(DIFCRS(I)*DIFCRS(I)+  MAUT(I)*SUMAUT(I))
      WRITE(7,250)
  250 FORMAT(1X,'    ENVELOP MAGNITUDE'  )
      WRITE(7,200)(ENV(I),I=1,K)
      GO TO 5
 [illegible]000 CONTINUE
      STOP
      END
```

EXAMPLE 1 (MAXIMAL-LENGTH)

CODE A

1. 1. 1. [illegible] 0. 1. 0.

CODE Y

1. 1. 1. [illegible] 1. [illegible] 0.

AA/A CORRELATION

[illegible]	[illegible]	[illegible]	[illegible]	-1	[illegible]	7	-1	-1	-1	-1	-1	-1	7	0	-1	0	-1	0	-1

AA/Y CORRELATION

-1	[illegible]	-1	[illegible]	-3	4	3	-1	3	-1	-1	-5	3	3	0	3	6	-1	-2	-1

Y/Y CORRELATION

-1	[illegible]	-1	[illegible]	1	[illegible]	7	-1	-1	-1	-1	-1	-1	7	0	1	0	-1	-2	-1

Y/A CORRELATION

-1	[illegible]	-1	[illegible]	3	[illegible]	3	3	[illegible]	-1	-1	3	[illegible]	3	4	-3	0	-1	0	-1

[illegible]	[illegible]	[illegible]	-.3	.0	-.3	-1.0	-.3	.0	.3	-1.0	-.3	.0	2.2	7.0	1.9	-1.0	[illegible]	-1.0	-.6
-1.0	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	-1.0	1.9	7.0	.2	.0	-.3	-1.0	-.3	.0	-.3	-1.0	-.3	.0	-.3
-1.0	[illegible]	[illegible]																	

[illegible]	[illegible]	[illegible]	[illegible]	-.3	[illegible]	[illegible]	.0	-1.0	.0	.3	4.0	2.2	[illegible]	.6	-1.0	.6	3.0	[illegible]	-1.0
[illegible]	[illegible]	-1.9	-3.0	-.0	3.0	1.9	3.0	1.0	.0	1.0	3.0	1.9	.0	-.3	-1.0	-1.0	-2.0	-1.9	-1.0
-.3	.0	.0																	

.3	[illegible]	[illegible]	-1.0	-2.0	-2.0	-1.0	-.3	.0	.3	1.0	.3	.0	2.2	7.0	1.9	-1.0	[illegible]	-1.0	-.6
[illegible]	[illegible]	[illegible]	[illegible]	-1.0	[illegible]	[illegible]	1.9	7.0	[illegible]	.0	.3	1.0	.3	.0	-.3	-1.0	-1.0	[illegible]	-1.0
-1.0	-.3	.0																	

.0	.0	-.3	-1.0	-1.0	-2.0	-1.0	-1.0	-.3	.0	1.0	3.0	1.0	.0	1.0	3.0	1.0	[illegible]	[illegible]	-1.0
[illegible]	[illegible]	-.3	-1.0	.0	3.0	.0	-1.0	.0	.0	2.2	4.0	.3	-3.0	-1.0	.0	-.3	-1.0	-.3	.0
[illegible]	[illegible]	[illegible]																	

.0	1.3	.0	-1.0	-.6	-1.0	-.0	-1.0	.0	.0	.0	-1.0	-1.3	-3.0	.3	4.0	1.3	.0	-1.3	-4.0
-1.3	.0	1.3	[illegible]	1.3	.0	-1.3	-4.0	-.3	.0	1.3	1.0	-.0	-3.0	-.6	1.0	.0	1.3	.6	1.0
.0	-1.0	-.3																	

SUM OF AUTOCORRELATION TERMS

.0	-.3	[illegible]	-1.3	-2.0	-1.3	-2.0	-.6	.0	.0	.0	.0	.0	4.5	14.0	3.8	-2.0	-1.3	-2.0	-1.3
-2.0	-1.3	-4.0	-1.3	-4.0	-1.3	-2.0	3.0	11.0	.5	.0	.0	.0	.0	.0	-.6	-2.0	-1.3	-2.0	-1.3
-2.0	-.0	.0																	

ENVELOPE MAGNITUDE

.0	1.2	2.0	1.0	2.1	1.0	2.1	1.2	.0	.0	.0	1.0	1.3	5.4	14.0	5.5	2.4	1.3	2.4	4.2
2.4	1.3	2.4	4.2	2.4	1.3	2.4	5.5	14.0	.4	1.3	1.0	.6	2.0	.0	1.2	2.1	1.6	2.1	1.6
[illegible]	[illegible]	[illegible]																	

EXAMPLE 2 (GOLAY'S COMPLEMENTARY)

```
CODE X
 1.   1.   1.   0.   0.   0.   1.   0.

CODE Y
 1.   1.   1.   0.   1.   1.   0.   1.

X/X CORRELATION
 -1   0  -1   0  -3   0   1   8   1   0  -4   0  -4   0   0   8   1   0  -3   0
 -1   0  -1

X/Y CORRELATION
 -1   0  -1   0  -3   0   3   0   0   0   0   0  -4   0   4   0   1   0   1   0
  1   0   1

Y/Y CORRELATION
  1   0   1   0   3   0  -1   8   0   0   4   0   4   0   0   8  -1   0   3   0
  1   0   1

Y/X CORRELATION
  1   0   1   0   1   0   1   0   4   0  -4   0   0   0   0   0   3   0  -5   0
 -1   0  -1

 .0  [illegible] [illegible] [illegible]  .0  [illegible] -1.0  -.3  .0  .0 -3.0 -1.0  .0  .3  1.0  2.9  0.0  2.5  .0  .0
 [illegible] [illegible] [illegible] -1.3  .0  [illegible] -4.3 -1.3  .0  .0  .0  2.3  3.0  2.9  1.0  .3  .0 -1.0 -3.0 -1.0
 .0  [illegible] [illegible] [illegible]  .0  -.3 -1.0  -.3  .0

 [illegible] [illegible] [illegible]  .0  -.3 -1.0  [illegible]  .0 -1.0  .0 -1.0  .0  1.0  3.0  1.0  .0  .0  .0  .0  .0
 .0  .0  .0  .0 -2.3 -4.3 -1.3  .0  1.3  .0  1.3  .0  .3  1.0  .3  .0  .3  1.0  .3  .0
 [illegible]

 [illegible] [illegible] [illegible] [illegible]  .0  [illegible]  1.0  [illegible]  .0  .0  3.0  1.0  .0  -.3 -1.0  2.2  0.0  2.5  .0  .0
 .0  [illegible] [illegible] 1.3  .0  [illegible]  4.0  1.3  .0  .0  .0  2.3  [illegible]  2.2 -1.0  -.3  .0  1.3  3.0  1.0
 .0  [illegible] [illegible]  .3  .0  [illegible]  1.0  .3  .0
```

```
-1.0    .0   -.3  -1.0   -.3    .0   -.3  -1.0   -.3

 SUM OF CROSSCORRELATION TERMS

  .0   1.0    .0   1.0    .0   1.0    .0   1.0   1.9    .0   1.9   1.0   -.0  -3.0   -.0   1.0    .3    .0   1.3   4.0
 1.3    .0  -1.0  -4.0    .0   4.0   1.3    .0  -1.3    .0  -1.3    .0   -.3  -1.0    .0   3.0    .0  -1.0  -1.9  -5.0
-1.9  -1.0   -.0  -1.0   -.0  -1.0   -.0  -1.0   -.3

 SUM OF AUTOCORRELATION TERMS

  .0    .0    .0    .0    .0    .0    .0    .0    .0    .0    .0    .0    .0    .0    .0   5.1  16.0   5.1    .0    .0
  .0    .0    .0    .0    .0    .0    .0    .0    .0    .0    .0   5.1  16.0   5.1    .0    .0    .0    .0    .0    .0
  .0    .0    .0    .0    .0    .0    .0    .0    .0

  .0   1.0    .0   1.0    .0   1.0    .0   1.0   1.9    .0   1.9   1.0    .0   3.0    .0   5.2  16.0   5.1   1.3   4.0
 1.0    .0    .0   5.0    .0   4.0   1.0    .0   1.3    .0   1.3   5.1  16.0   5.2    .0   3.0    .0   1.0   1.9   5.0
 1.9   1.0    .0   1.0    .0   1.0    .0   1.0    .3
```

Appendix C.7

Microprocessor Components

1.0 INTRODUCTION

The purpose of this investigation was to characterize the technology of microprocessor components, memories and structures. The study is heavily weighted in favor of components that are available now. The many designs that are being planned or that are on-going at the various semiconductor houses are not considered in this evaluation. It is felt that there is not enough assurance that these devices will be available for equipment delivery in 1974.

2.0

Refer to Table 1 for a list of the microprocessor sets available. Also included are lists of random access memories and programmable read only memories.

A listing of the several factors considered in evaluating the various component sets is included in the following tables:

Table 2 - MOS Random Access Memories
Table 3 - MOS PROMS
Table 4 - CPU Parameters Used in Evaluation

Table 5 - RAM Parameters Used in Evaluation

Table 6 - Comparison of Microprocessor Sets

3.0 CONCLUSIONS

Based in the investigation, the following selections have been made:

A) The National General Purpose Controller/Processor appears most promising for the Packet Radio repeater requirements.

B) Complementary MOS (C-MOS) for Random Access Memories.

C) Programmable (electrically erasable) Read Only Memories will be used.

TABLE 1

<u>COMPANIES WITH MICROPROCESSOR SETS</u>

1. <u>INTEL</u> - 4 BIT - 4004, 8 BIT - 8008, BOTH ARE SINGLE CHIP CPU'S.
2. <u>NATIONAL SEMICONDUCTOR</u> - MULTICHIP CPU, CAN CHOOSE WORD WIDTH.
3. <u>ROCKWELL-MICROELECTRONICS</u> - 4 BIT SINGLE CHIP CPU.
4. <u>FAIRCHILD</u> - PPS 25 - DECIMAL ORIENTED SET.

<u>COMPANIES WITH PLANS FOR MICROPROCESSOR SETS</u>

1. <u>INTEL</u> - 8 BIT N-CHANNEL CPU - 8080 - DEC. 73.
2. <u>MOTOROLA</u> - 8 BIT N-CHANNEL CPU - FIRST QUARTER 74.
3. <u>ROCKWELL-MICROELECTRONICS</u> - 8 BIT CPU.
4. <u>WESTERN DIGITAL CORP</u>. - 8 BIT N-CHANNEL CPU - BEFORE 74.
5. <u>SIGNETICS</u> - NO INFORMATION.
6. <u>HUGHES AIRCRAFT CO</u>. - NO INFORMATION.
7. <u>AMI</u> - 16 BIT GENERAL PURPOSE SET.

TABLE 2

MOS RAM'S

	INTEL				MOSTEK	NATIONAL		EA	AMS	RCA (256 BIT)
	1103-1	2102	2102-1	2105-1	4006-6	4260	5260	1502	6002	CD4061
...m Power Dissipation (MW)	475	300	300	500	575	575	575	165	275	80
...l Power Dissipation (MW)	375	150	150	325	400	400	400	100	180	5
...y. Power (MW)	75	150	150	65	50 50	100	100	30	2	25μw
...s Time (N. Sec)	150	700	350	80	400	450	375	150	150	280
... Times: Read (N. Sec)	300	1000	500	180	400	600	500	290	250	400
Write (N. Sec)	340	1000	500	180	650	750	625	450	250	500
...of Input	Mos[1]	TTL[2]	TTL[2]	Mos-TTL[3]	TTL[2]	Mos-TTL[3]	Mos-TTL[3]	Mos-TTL[3]	Mos[1]	TTL[2]
...of Output	Current Sense	TTL	TTL	Current Sense	Current Sense	TTL	TTL	TTL	Current Sense	TTL
...h Interval/No. Of Refresh Cycles	2M Sec 32 Cycles	Static	Static	10μ Sec 1 Cycle	2m Sec 32 Cycles	1m Sec 32 Cycles	2m Sec 32 Cycles	2m Sec 1 Cycle	2m Sec 32 Cycles	Static
...ing Temperature	0°C TO 70°C	0°C TO 70° C	0°C TO 70°C	0°C TO 70°C	0°C TO 70°C	-55°C TO 125°C	-25°C TO 70°C	0°C TO 70°C	0°C TO 70°C	-55°C TO 125°C
...Supplies (Volts)	+19, +22	+5	+5	+12, -5	+5, -12	+5, -12	+5, -12	± 12	+7, +20 +22.5	Variable +3-+15
... Pins	18	18	18	18	16	16	16	18	22	16
..., 1-24 Quantities	$13.50	$24.00	$44.50	$60.00	$16.75	$47.80	$18.00	$41.20	$25.00	$12.00
... Level Shifters And Amps-1024x8 Bit	29	0	0	10	8	2	2	2	29	0
...of Mos	P	N	N	N	P	P	P	N	P	Both

1. Mos - 15-20V Voltages Swings Are Needed.
2. TTL - 5V Logic Levels
3. Mos-TTL-Part Of The Inputs Require Mos Voltages And Part Can Be

TABLE 3

MOS ELECTRICALLY PROGRAMMABLE AND ERASABLE ROM'S

		Intel 1702A	National MM4203
1.	Power Dissipation	700 mw	700 mw
2.	Cycle Time	700 n sec.	700 nsec.
3.	Chip Enable Logic	Single Line	3 Bit Control
4.	TTL Compatible	Yes	Yes
5.	Power Supplies	+5, -9	+5, -12
6.	No. of Pins	24	24
7.	Operating Temp.	o°C to 70°C	-55°C to 85°C
8.	Required Clocks	None	None

TABLE 4

CPU PARAMETERS USED IN EVALUATION: REFER TO TABLE 6

1. Power Requirements - Includes CPU, supporting circuitry, and 1 I/O port. Does not include any memory.

2. Program Execution Efficiency - Predicted operating time for a particular benchmark program.

3. Supporting Circuitry Required - The number of IC's required to make the CPU a workable system. Does not include memory or TTY interface.

4. Program Storage Efficiency - The number of memory locations necessary to store a benchmark program.

5. Power Supplies
   a. Number required
   b. Voltages required are the voltages of common usage?

6. I/O Interfacing
   a. Ease of design - what's available to aid hardware design?
   b. Number of input and output ports.
   c. Software - what's available under software control?

7. P.C.B. partitioning ease - breaking up components on 2 or 3 boards, how many pin-outs are necessary?

8. Clock requirements
   a. Frequency - is it easy to derive?
   b. Number of phases
   c. TTL or MOS level

9. Allowable voltage fluctuations or ripple

TABLE 4 (Cont'd)

10. Cost

    a. CPU chip(s)

    b. Supporting circuitry

    c. P.C. boards

    d. Prototype kit

11. Memory addressing

    a. Direct

    b. Indirect

    c. Indexed

    d. Literal

12. Instruction set utility

    a. Number of instructions

    b. Number of different instruction types

    c. Unusual instructions

13. Number of registers

    a. Accumulator

    b. Index

    c. Scratch pad

14. Stack capabilities

    a. Size of stack

    b. Is it accessible through software?

15. Largest memory capacity addressable

16. Capability for customer microprogramming

<u>TABLE 5</u>

<u>RANDOM ACCESS MEMORY PARAMETERS USED IN EVALUATION</u>: REFER TO TABLE 6

1. Cost per bit - interface and control circuitry included. (Refresh circuitry included if necessary).
   a. 1024 word x 8 bit system
   b. 1024 word x 16 bit system
2. Power dissipation per bit - interface and control circuitry. (Refresh circuitry included if necessary).
3. Cycle Times
   a. Access - time from chip select until data available.
   b. Read - access time plus deselect time.
   c. Write/refresh
4. Refresh requirements
   a. Number of addr[illegible] changes required for complete memory refresh.
   b. Minimum time between refreshed.
5. TTL compatibility
6. Operating temperature - if not specified over the military range, will the memory function at high temperatures.
7. Power supplies
   a. Number required
   b. Voltages required - are the voltages of common usage?
8. Availability
   a. Can the part be had in large quantities?
   b. Second sourced
   c. State of the art - is it the current design?
9. Packaging Efficiency - Using 16,192 bits as a standard, how many IC's are necessary to produce a memory system with interface, control, and refresh circuitry included.

		8008-1
1.	Power Requirements (Watts)	With TTL - 7W With CMOS - 1W
2.	Program Execution Efficiency	
3.	Supporting Circuitry Required	37 IC's
4.	Program Storage Efficiency	124 Words x 8 Bits
5.	Power Supplies	+5V, -9V
6.	I/O Interfacing	8 possible I/O Ports No Input Lines for communicating with CPU except Control Panel Interr
7.	P.C.B. Partitioning	An 8 Bit Bidirectional bus is used for Data I/O and Addressing. This allows for fewer Pin Outs from board.
8.	Clock Requirements	TTL Level Clocks 2 Phase
9.	Allowable Voltage Fluctuations	+5%
10.	Cost	CPU Chip - $180 SIM 8-01 - $900 INTELLEC - 8 $3650
11.	Memory Addressing	Indexed only
12.	Instruction Set Utility	48 Instructions
13.	Number of Registers	1 8-Bit Accumulator 6 General Registers 2 of the 6 can be used for Indexing
14.	Stack Capabilities	7 14-Bit Registers for Program Counter Storage only
15.	Largest Memory Addressable	16K word x 8 Bit
16.	Capability for Microprogramming	None

TABLE 6

COMPARISON OF MICROPROCESSOR SETS

IMP-16	IMP-8
With TTL - 14W With CMOS - 3W	With TTL - 8W With CMOS - 2W
40 IC's	34 IC's
49 Words x 16-Bits	
+5V, -12V	+5V, -12V
6 General Purpose Flags 4 General Purpose Jump condition inputs 1 General Interrupt Input	Same as IMP-16
16-Bit Bidirectional bus is used for I/O and Addressing. The wider word and greater number od CPU lines makes for more Pin Outs	8-Bit Bidirectional bus is used for I/O and Addressing
-12V to +5V Swing 4 Phase	-12V to +5V 4 Phase
±5%	±5%
CPU Chips - $430 IMP - 16C - $950 IMP - 16P - $3850	CPU Chips - $270 IMP - 8C - $800 IMP - 8P - $3750
Direct, Indirect, Indexed, Literal Relative Type	Direct, Indirect, Indexed, Literal Paged Type
43 Instructions Software Control of Stack and External Flags. External Jump Condition Inputs	38 Instructions Software Control of Stack And External Flags. External Jump Condition Inputs
4 16-Bit Accumulators	4 16-Bit Accumulators
16 16-Bit Registers Accessable through Software	16 8-Bit Registers Accessable through Software
65K word x 16-Bit	65K word x 8-Bit
Instruction Sets can be developed	Same as IMP-16