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A DATA LINK DESIGN WITH A VARIABLE DATA RATE AND VARIABLE JAM RESISTANT CAPABILITY

The MITRE Corporation

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This report describes work performed on Project 7130 during FY79. The objective of this phase was to address the technical issues involved in variable data rate/variable jam resistant data link technology. The design approach was based on cost-performance trade-offs for the signal processing elements of a two-terminal, data link system with up to 1 GHz spread spectrum bandwidth operating with data at rates of from 100bps to 1Mbps. The design and design process for a laboratory model are

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presented. Cost estimates for large production quantities of the data link are also presented.

ABSTRACT

The FY79 work program on Project 7130 has included analyses of jam-resistant technology for variable-data rate links and detailed design for a laboratory brassboard. The selection of a design approach has been based on cost-versus-performance trade-offs for the signal processing elements of a two-terminal, data link system possessing up to 1000 MHz of spread spectrum bandwidth with data rates ranging from 100 bps to 1 Mbps. This design approach has been applied to a laboratory model, which will be fully tested and evaluated in an FY80 follow-on effort. Cost estimates for large production quantities of the data link have also been projected.

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INTRODUCTION

1.1 Scope

In FY79, Project 7130 completed analyses for a variable-data-rate, variable-jam-resistant data link design. The analyses focused on technology which can be applied particularly to a number of unmanned tactical missions such as standoff attack/weapon delivery and reconnaissance. This report documents the results of these analyses and the rationale which subsequently was used in the design of a laboratory data link model. An attempt is made here to:

1. give perspective to some specific technical issues in the design of variable data rate links,
2. provide insight into the available jam-resistant data link technology, and
3. document the basic laboratory design.

Integration, testing, and evaluation of the laboratory system comprise a follow-on FY80 effort.

This year's work focused on many of the performance-versus-cost aspects of a variable data rate capability for data links with expendable vehicle platforms on one end. The project emphasized the design and implementation of a two-terminal laboratory brassboard link. In the course of the design for the laboratory link compromises have been made which make our implementation somewhat different from that recommended by our analyses. Factors such as component availability and our laboratory-oriented versus a production-type technology influenced many of these decisions. Although these limitations do not constitute serious constraints to our experimental program, the

reader will note, at times, differences between our recommendations and our implementation. We have attempted to maintain clarity in distinguishing these two aspects of the data link, and we hope that any dichotomy in our presentation is not interpreted as contradiction.

Often, we identified several alternative approaches to the design. These alternatives were analyzed from a cost-versus-performance standpoint, but in some cases we had to rely on subjective reasoning to select an approach. Through our experimental program of detailed design, implementation, and evaluation, we expect to gain insight and experience in further quantifying the merits of at least the selected approach.

1.2 Background

Jam resistance is an important concern for most tactical communication systems. Maintaining reliable communication in the presence of powerful jamming is a difficult task for any system, but becomes particularly stressing when the system must be low cost.

Spread spectrum techniques have become a fundamental element in improving jam resistance. Spread spectrum in the form of binary phase and linear FM modulation has been used for years in radars to improve combined range and Doppler performance, but only recently has the state of technology reached a point where designers can consider similar techniques for virtually any tactical communication problem. Although we can now appropriately consider the application of these jam resistance techniques to expendable equipment, relevant cost-versus-performance factors must first be closely examined. Spread spectrum should not be indiscriminantly incorporated in any system expected to operate in a jamming environment.

MITRE initiated an IR&D program in FY77 to focus on jam-resistant techniques for a dedicated weapon control data link. By carefully tailoring that data link design to meet only specific mission needs and by the judicious combination of several anti-jam (AJ) techniques, a data link with low-cost potential and substantial jam resistance (potentially several orders of magnitude improvement over the best existing command and control links) was demonstrated. That work also demonstrated how a cost-effective trade could be gained by decreasing the spread spectrum bandwidth by one order of magnitude.

That IR&D program had initially focused on obtaining low data rates (50 - 200 bps) with up to 63 dB of spread spectrum processing gain against jamming. A follow-on program was then conducted in FY78 to examine the potential for applying the same techniques to a variable-data-rate link. Approaches were developed and experimentally investigated to incorporate into the existing data link three distinct data rates spanning several orders of magnitude, while providing a direct trade of jam resistance for data rate. Through that experience it became apparent that, while the data rate/jam resistance trade was conceptually feasible and also attractive from a system use point of view, it presented some interesting cost/technology options.

That background served as a base for the FY79 project, which was part of the MITRE Mission Oriented Investigation and Experimentation (MOIE) Program. The emphasis for the project was placed on a laboratory-oriented effort that would examine the technology/cost factors involved in obtaining a substantial jam-resistant data link capability with an effective tradeoff between data rate and jam resistance. A data rate range from tens of bits per second (bps) to one or two megabits per second (Mbps) was considered as representative of potential needs for many tactical weapon delivery and RPV-type reconnaissance applications. A spread spectrum bandwidth of 1000 MHz was initially

viewed as a realistic technical challenge for a data link which possibly included expendable terminals.

The work program concentrated on the signal and data processing technologies involved in obtaining a substantial jam-resistant capability. Those technologies represent an important piece of the jam resistance problem. To be most effective, the results of this work must be integrated with other elements such as antenna processing techniques, geometry considerations, operational strategies, and realistic jammer threats, to obtain a substantial low-cost jam resistance capability for these missions.

1.3 Summary

The work during the FY79 program included analyses of jam resistant technology for variable-data-rate links, detailed design for a laboratory brassboard, and subsequent implementation of that design.

The perceived applications for this work include tactical weapon delivery and reconnaissance. The data rate needs for a single link span a wide range from less than 100 bps to over 1 Mbps. The low end of the range is needed for transmission of control and/or status data between the master (control) terminal and a remote, airborne (slave) terminal. The high end of the data rate range is required for transmission of compressed video data, typically from the slave to the control terminal. Since these tactical missions potentially encounter a high jamming environment, a substantial amount of jam resistance is a critical need for mission success.

This work limited detailed study to the considerations involved in communication between two terminals. One terminal (control) is assumed to be a high-valued airborne or ground station; the other terminal (slave) is on an airborne, potentially expendable vehicle.

Where possible, the complexity (and resulting expense) of the data link can be placed in the control terminal. Extension of the techniques described here to the problems of a single control terminal and multiple vehicle terminals is possible but significant cost/performance compromises are indicated.

We examined signaling techniques and RF power trades involved in obtaining substantial jam-resistance using approximately 1000 MHz of spread spectrum bandwidth. A design incorporating 800 MHz of bandwidth has resulted. Through the analyses and design, we developed insight into the problems of obtaining low-cost implementations with a direct trade in jam resistance over a wide span of data rates. Complex signal generation and reception procedures are necessary and expensive transmitter technology is involved.

The technology requirements are eased somewhat by segmenting the data rate needs into bands and treating each band individually. By using this approach, we have divided the problem into three ranges covering these resulting data rates:

- Low range (control data) 61 - 981 bps
- Medium range (voice-rate data) 6 - 96 kbps
- High range (video data) 36 - 575 kbps

Two-way communication between the command and the slave terminal is normally desired. In addition, the ability for the command terminal to use the data link for round-trip ranging is often necessary. A ranging capability was included in the control data signaling requirements. Communication from the control terminal to the slave terminal (the forward link) generally does not require video data. Back link (slave to command) communication could involve all three types of data. Data rates are independently selected for each (one-way) path. The following two-way communication system is shown in Table 1-1.

TABLE 1-1: TWO-WAY DATA COMMUNICATION MATRIX

	FORWARD LINK	BACK LINK
CONTROL DATA	X	X
VOICE-RATE DATA	X	X
VIDEO DATA		X

Analyses of signaling strategies show that different techniques for each range should be employed to obtain a direct trade of signal processing gain for data rate over each range. Use of the full 800 MHz bandwidth for each range of data rates realizes the overall processing gain versus data rate trade. For our parameters, this resulting trade is shown in Figure 1-1. From a signal processing viewpoint, we have used the spread spectrum bandwidth effectively.

The jam resistance of the design is dependent also on the amount of transmitted power used for each data rate. We have used a figure of merit for the relative jam resistance of any given data rate. This factor is the energy transmitted per information bit times the total spread spectrum bandwidth used for that data rate. Use of this factor as a true indicator of jam resistance assumes (1) appropriate stratigies are employed to force a jammer to continuously jam over the entire spread bandwidth, (2) the jammer has a fixed average power constraint, and (3) cost to the jammer of obtaining a given power level for the various bandwidths of interest to us is not a first order effect.

Using this figure of merit, jam resistance is a direct function of processing gain and transmitter power characteristics. To determine the most cost-effective way to achieve a given amount of relative jam resistance, RF power generation technology was examined in conjunction with the signal processing analyses.

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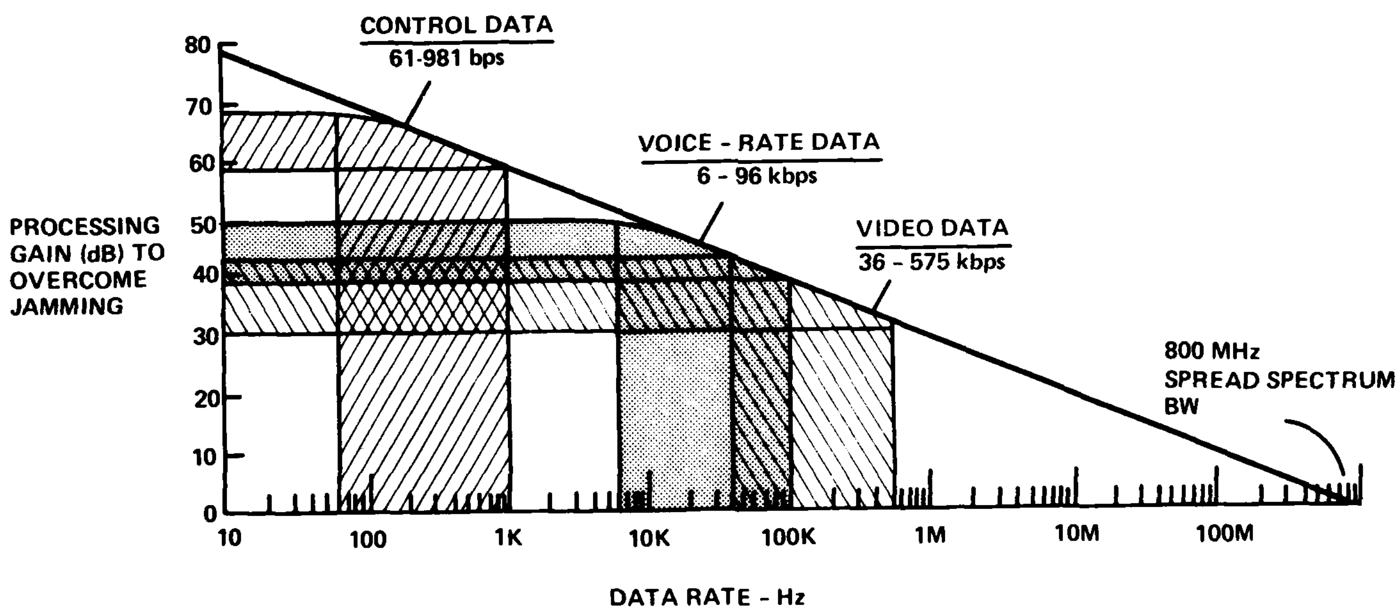

FIGURE 1-1: DATA RATE/ PROCESSING GAIN TRADE

The most attractive transmitter options include travelling-wave tube (TWT) and voltage-tuned magnetron (VTM) technology. To achieve the range of data rates while maintaining the desired jam resistance, a multi-mode (pulsed and CW) transmitter is necessary. An approximately linear trade in jam resistance figure of merit and data rate over the full (segmented) range can then be obtained with the combined contributions of spread spectrum signal processing and RF power generation technology.

The capability available with this design is substantial when compared with existing data links. A cost estimate for the slave terminal was done to gain some insight into the feasibility of the design and to identify the resulting technology cost drivers. Parts cost for this terminal, assuming a production quantity buy for 10,000 terminals, approaches $10,000. The transmitter cost is the most significant item, followed by the frequency synthesizer and then the detection, timing and control logic. The integrated effect of including high RF power, large spread spectrum bandwidth and both multiple and high data rates has a significant impact on the resulting cost.

A laboratory brassboard model of the two-terminal data link design is being implemented. Only the low-power parts of the RF section are included in the brassboard. Frequency hopping is implemented first in a 16-channel (200 MHz) system which will be extended in FY80 to the full 64-channel (800 MHz) system. Full integration, test and evaluation of both the 200 MHz and 800 MHz systems are planned for the follow-on FY80 program.

The remainder of this report provides the results of the analyses and design approaches for the variable-data-rate/variable-jam-resistant data link. Section II gives some background and perspective on the basic elements of jam resistance and relates jam resistance to a variable-data-rate capability. In Section III we discuss the design

issues involved in achieving the desired performance. Technology trade-offs and the rationale for our design decisions are presented. The results of the data link design are given in Section IV. A block diagram description of the laboratory brassboard is presented along with a projection of high-volume production costs for an expendable terminal.

SECTION II

JAM RESISTANCE AND VARIABLE DATA RATE

In this section we will describe some basic elements of jam resistance and their implications on the variable-data-rate tradeoffs.

2.1 The Problem

The basic problem is illustrated in Figure 2-1. We desire reliable data communication over a channel in which the interference or noise level is dominated by the jammer. The signal-to-noise ratio at the input to the receiver $\left(\frac{S}{N}\right)_i$ is set by both our signal power and the jammer power. The signal-to-noise ratio at the data decision circuitry is $\left(\frac{S}{N}\right)_o$. Through various processing procedures in the receiver, we hope to realize a substantial increase in $\left(\frac{S}{N}\right)_o$ compared to $\left(\frac{S}{N}\right)_i$. In the design of appropriate signaling techniques for the data link and subsequent processing procedures, emphasis is placed on two areas: (1) techniques that provide reliable data recovery at low $\left(\frac{S}{N}\right)_o$, and (2) realistic approaches to achieve the highest value of $\left(\frac{S}{N}\right)_o$.

The input signal-to-noise ratio is affected by many factors, including:

- signal and jammer transmitter characteristics
- antenna factors
- geometry (e.g., distance) factors
- propagation effects.

The output signal-to-noise ratio is affected by the $\left(\frac{S}{N}\right)_i$ plus the processing in the receiver. We are interested here in examining only the technologies appropriate for the signaling factors in obtaining jam resistance.

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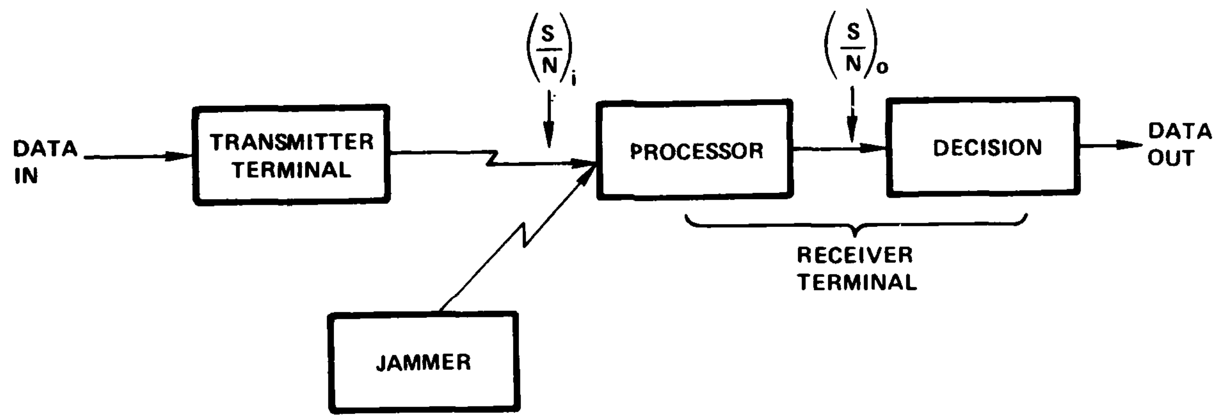


FIGURE 2-1: DATA TRANSMISSION IN THE PRESENCE OF JAMMER

First we will assume that the jammer provides white Gaussian noise. The optimum linear processor will produce an output signal-to-noise ratio at our data decision time

$$\left(\frac{S}{N}\right)_o = \frac{E_b}{N_o}$$

where E_b = energy/information bit and

N_o = jammer noise power/Hz

The bit energy is

$$E_b = \overline{P}_s \times T_{bit}$$

where $\overline{P}_s$ = average signal power during a bit time

T_{bit} = time duration of a data bit

$= \dfrac{1}{\text{Data Rate}}$

The jammer noise density is

$$N_o = \frac{\overline{P}_j}{W_{ss}}$$

where $\overline{P}_j$ = average jammer power

W_{ss} = bandwidth over which the jammer uniformly spreads power

We can now see that

$$\left(\frac{S}{N}\right)_o = \frac{E_b}{N_o} = \frac{\overline{P}}{\overline{P}_j} \times T_{bit} \times W_{ss} \qquad (1)$$

Since the ratio of average signal-to-average jammer power is the average input signal-to-noise ratio, we have

$$\left(\frac{S}{N}\right)_o = \left(\frac{S}{N}\right)_i \times T_{bit} \times W_{ss} \qquad (2)$$

The factor $T_{bit} \times W_{ss}$ will be called the processing gain available to provide jam resistance. No specific restrictions have been placed on the use of this available time-bandwidth space by the data link signaling; the use is generally restricted only in that the jammer must be forced to a white noise jammer, spread evenly over time and bandwidth.

If we assume that the amount of average jamming power $\overline{P}_j$, is fixed, we can control the remaining parameters and adjust them independently. Several observations can immediately be made.

- Spreading our signal over a wider band gives a linear improvement in performance (this shows a fundamental virtue of spread spectrum in a jamming environment. Similar improvement in the standard communication problem is not normally available since N_o, not $N_o W_{ss}$, is fixed).
- Increasing the bit time T_{bit} (decreasing the data rate) gives a direct, linear improvement. This fact provides the motivation for source coding, such as data compression and demonstrates the direct tradeoff available between data rate and jam resistance.
- Maximizing the performance for any fixed data rate involves maximizing $\overline{P}_s \times W_{ss}$.

We want to effectively exploit the factors under our control to maximize $\left(\frac{S}{N}\right)_o$ and minimize cost. Throughout the course of our work, we have used these factors to define a figure of merit (FOM) for the relative amount of jam resistance provided.

$$\text{FOM} = 10 \log (\overline{P}_s \times T_{bit} \times W_{ss}) \text{ dB} \tag{3}$$

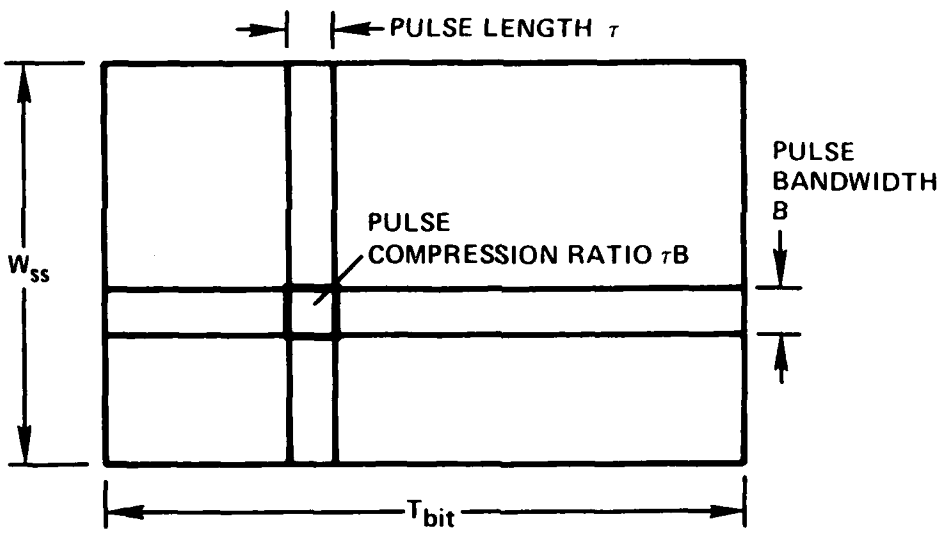
$$= 10 \log (E_b \times W_{ss}) \text{ dB} \tag{4}$$

The data link design reduces to finding the combination of signal and data processing factors with the proper transmitter technology to achieve the maximum figure of merit at minimum cost. When expressed in the form of Equation (3), the available jam resistance is a function of the average power and total spread spectrum bandwidth used for a given value of T_{bit}. Many practical and technology-related considerations influence the selection of values for these factors. Although we apparently have considerable flexibility in optimizing the parameters, the solution for one value of T_{bit} may not be the same as that for another. Thus, a single optimum data link design that encompasses a wide range of data rates may be considerably complicated by this additional dimension.

2.2 Considerations for Parameter Selection

In developing practical signaling techniques which use the available time-bandwidth signal space ($T_{bit} \times W_{ss}$), we consider approaches which partition the space into smaller segments as shown in Figure 2-2. We will consider the use of an instantaneous bandwidth B and a pulse length τ for the basic signal transmission. As long as the use of B and τ force the jammer to continuously fill the entire $T_{bit} \times W_{ss}$ space, we have considerable freedom in the selection of τ and B.* The time-bandwidth product τB of the basic signal should be limited to a size which allows matched filtering in the receiver with available technology. The instantaneous bandwidth B must be large enough to provide acceptable fine timing and ranging but small enough to reduce complications in synchronization and signal tracking and lessen speed/power requirements on the associated circuitry. The pulse length τ

* Note that signaling need not demand coherent processing over the entire $T_{bit} \times W_{ss}$ space, and the spread spectrum bandwidth W_{ss} does not even have to be contiguous.



W_{ss} = TOTAL SPREAD SPECTRUM BANDWIDTH

T_{bit} = TIME DURATION OF A DATA BIT $\left(= \frac{1}{\text{DATA RATE}} \right)$

τB = TIME-BANDWIDTH PRODUCT OF EACH TRANSMISSION

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FIGURE 2-2: SIGNAL SPACE FOR SPREAD SPECTRUM SIGNALING

should be short enough to protect against repeat jamming strategies, but long enough to provide enough energy for each transmission (i.e., short, low duty factor transmissions require a high peak power transmitter to obtain the necessary energy E_b).

As a result of partitioning the signal space, we can utilize frequency and time positions in a random-hopping pattern and have the following parameters:

- number of frequency-hop positions $\frac{W_{ss}}{B}$
- number of time-hop positions $\frac{T_{bit}}{\tau}$
- pulse compression on each transmission τB

Implementation of frequency hopping requires a frequency synthesizer in the transmitter and receiver. Each synthesizer must be capable of generating $\frac{W_{ss}}{B}$ frequencies. The time hopping requires a synchronized timing window for detection in the two terminals, and, for low duty factors, a high peak power is required. The pulse compression requires some form of intrapulse modulation, such as phase coding, and a matched filter of complexity τB in the receiver. If reasonable choices are made for each parameter, the overall FOM will be maximized and the resulting cost will be minimized.

The signaling technique must also allow data to be transmitted. For this we want signal modulation and coding schemes that minimize the required $\left(\frac{S}{N}\right)_o$. Coherent M-ary signaling and robust error correction coding are well known techniques to accomplish this.

Error coding will also be necessary to control the jamming strategy and drive the jammer toward white noise jamming. Even relatively simple error coding can neutralize various jammer strategies. Figure 2-3 illustrates typical results of communicating with and

without error coding in the presence of a jammer. Performance without error coding against the white noise jammer is shown in curve (a). By choosing a different jamming strategy, such as a combination of partial-time and partial-band jamming, the jammer (under a constrained average power) can optimize jamming effectiveness and limit data link performance to curve (b). A substantial advantage (tens of dB) to the jammer generally results here. With the addition of error coding and a re-optimized jammer, the performance moves to curve (c). The difference between curves (a) and (c) can be negligible when even modestly powerful codes are employed. Finally, if the jammer returns to a white noise condition when error coding is employed, curve (d) results; in comparison to the white noise jamming without coding, the error coding typically provides a performance improvement.

For best results the *choice of error* coding should be compatible with the selection of signaling and should favor M-ary codes. Additional improvement against optimum jammer strategies can also be gained by using codes capable of handling erasures.*

2.3 Variable Data Rate

In Section 2.1 we saw that a direct tradeoff between data rate and jam resistance is available. Our data link design strives to obtain maximum jam resistance over a wide range of data rates.

Variable-data-rate capability p vides various options for transmission of fixed or multiple-source data. If the trade were linear and continuous over a wide range of data rates and we had truly maximized the figure of merit for the cost, the data link design would

* An erasure is defined as an error whose symbol location is known. Declaration of an erased received symbol can be made when jamming is detected or when the receiver provides low confidence in the detected symbol.

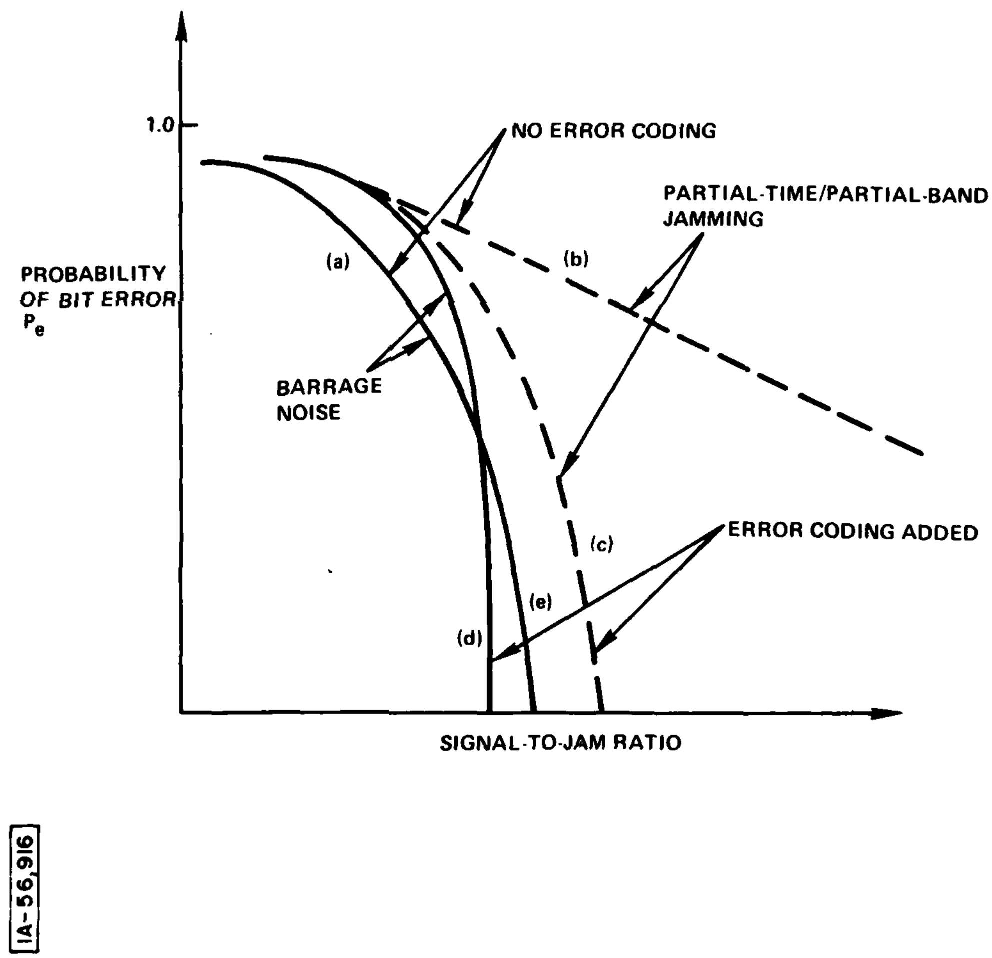

FIGURE 2-3: PERFORMANCE AGAINST JAMMING

provide the solution to many problems. A common design and set of equipment could be available for any data rate requirement. While obviously desirable, such a solution does not usually result. Tailoring to individual needs will almost always provide a more cost-effective answer for a specific requirement. A design in which this trade is available over a more limited range of data rates may, however, be cost effective and have multiple system applications. The desire to achieve multiple system applications is a commonly accepted motivation for a variable-data-rate capability.

Other system use opportunities arise when this capability exists. If we have a direct trade of data rate for jam resistance, the data rate can be changed as the jamming changes. A fixed level of performance can thereby be achieved. The quality of data transmission can remain high as the jamming level increases while the quantity (data rate) decreases. In the case of extreme jamming, at least a small amount of information (e.g., a limited but important set of messages) can be communicated reliably. Another possibility is to source code the data such that the most important data is encoded for the lowest data ra . One implementation of this approach might be to highly compr s the data source for transmission at the lowest (and most reliable) data rate, some of the redundancy might be sent at higher rates and at the highest rate the uncompressed data could be sent. The increasing levels of data compression then have increasing amounts of jam resistance.

In summary, a variable-data-rate capability can provide various system level alternatives. It may provide a cost-effective approach to needs of multiple systems, and it may open up new approaches to data transfer and graceful degradation of system performance in a jamming environment.

SECTION III

THE DATA LINK DESIGN: APPROACHES AND CONSIDERATIONS

This section will describe the significant approaches considered for the variable-data-rate, variable-AJ design as well as the resulting design. A laboratory model of a two-terminal system is based on these results.

3.1 Signaling Schemes for Variable-Data-Rate/Variable-AJ System

The design of a jam-resistant data link which operates at one data rate is relatively simple. However, the requirements for a system with variable-data-rate capabilities are much more difficult to satisfy. In this section, we will discuss the factors which affect the selection of a signaling scheme for a variable-data-rate link.

3.1.1 Transmitter Considerations

One of the major factors in determining a signaling scheme is choosing from several types of available transmitters. Maximum performance at a minimum cost is a prime consideration in this choice.

Transmitters can operate in a CW mode or a pulsed mode. Transmitters which are designed for pulsed operation have high peak power output specifications and typically a maximum duty factor specification of ≈1%. In order to maximize the average signal power ($\overline{P}_S$ in equation 3) and thus maximize the jam resistance, these devices should be operated at their peak output power and maximum duty factor. If a pulsed transmitter is used, the signaling scheme should allow the duty cycle to remain fixed at its maximum value regardless of the data rate. Transmitters which are designed for CW operation can also be used in a pulsed mode at no increase in peak output power as the

duty factor is lowered. Thus, the average signal power in a CW transmitter used for pulsed operation is reduced by the duty factor and the available energy per transmitted bit is reduced. Hence, a transmitter of this type should be operated in the CW mode to be used most efficiently. In summary, the signaling scheme should be designed for a pulsed transmitter operating at $\approx$1% duty cycle and/or a CW transmitter operating continuously. A more detailed discussion of transmitter options is given in Section 3.4.1.

3.1.2 Basic Parameter Selection

To obtain high processing gain (high AJ), the signal energy must be spread over a bandwidth which is large compared to the basic data bandwidth. A goal in this project was a spread spectrum bandwidth W_{ss} of approximately 1000 MHz to accommodate signaling with data transmission rates up to 1 MHz. The amount of spread spectrum bandwidth that can be realistically obtained is limited by carrier frequency, transmitter and antenna considerations. For data link communication over distances in the order of 100 km, the maximum frequency should be limited to less than approximately 12 GHz to minimize attenuation due to rain.[1] The design of desirably-shaped antenna patterns usually restricts the bandwidth to less than 10% of the carrier frequency. The choice of carrier frequency and bandwidth affects transmitter cost, as discussed in Section 3.4.1. These factors and others related to the signaling structure, as discussed in later sections of this report, have led to a nominal choice of 800 MHz for the total spread spectrum bandwidth and 9 GHz for the carrier frequency.

The manner of obtaining a spread spectrum bandwidth of 800 MHz must be considered. Two choices exist: (1) use an instantaneous bandwidth of 800 MHz, or (2) use a smaller instantaneous bandwidth plus frequency hopping. By using an instantaneous bandwidth of 800 MHz,

we require basic decision timing of 1.25 ns. Encoding of the data could be performed by transmitting a pseudo-noise (PN) sequence with a 1.25 ns chip period. Detection and appropriate processing of chips which are this narrow is not practical in a low-cost system at the present state of the art. Therefore, some amount of frequency hopping is required. An instantaneous bandwidth of 10 MHz (chip width = 100 ns) allows the hardware design to be simple and inexpensive and still allows the system to meet timing/ranging requirements in the order of 10 ns/10 feet.* For simplicity in the hardware, the number of frequency channels should be a power of two. These constraints can be accommodated with an instantaneous bandwidth of 12.5 MHz (chip width = 80 ns) and frequency hopping over 64 channels to produce a spread spectrum bandwidth of 800 MHz.

For low data rates, time hopping can be used to provide some of the processing gain. From both a technology and a geometry point of view, pulses which are 10 μs wide are short enough to prevent a responsive jammer from locking onto the signal and repeating it within the time duration of the pulse. This pulse width is also consistent with typical low duty factor transmitter specifications. The selected pulse consists of 128 80 ns chips for a total pulse width of 10.24 μs. The intrapulse PN coding on each 10.24 μs pulse increases the processing gain by a factor of 128:1, which is also a reasonable time-bandwidth product for matched filter processing. In order to match the duty factor of ≈1% in a high peak power transmitter, a pulse will be transmitted in one 10.24 μs time slot every 128 time slots. A frame is defined as 128 time slots (1.31 ms). The combination of frequency hopping, time hopping, and PN coding provides an

* A timing/ranging need of 10 ns/10 feet may not be required in all applications, but an instantaneous bandwidth of 10 MHz or more can provide the accuracy in all cases that need these functions.

available processing gain as listed below:

Frequency hopping:	64:1 = 18 dB
Timing hopping:	128:1 = 21 dB
PN coding:	128:1 = 21 dB

3.1.3 Continuously Variable Versus Stepped Variable Data Rates

The optimum system which accommodates a direct trade of data rate for jam resistance would allow the data rate (and processing gain) to vary continuously and responsively with the jamming level. In this way, a maximum amount of information transfer would occur. Various alternatives were considered for obtaining this data rate/AJ trade over a wide range of data rates.

One method of automatically varying the data rate over a continuous range with the jamming level is to have the transmitting terminal repeat messages which are received incorrectly. Thus, as the jamming level increases, more and more messages are repeated thereby lowering the data rate. Note that this system requires reliable transmission on the back link (slave to command) so that the transmitter will know when data has been received incorrectly. The characteristics of this repeat request scheme are summarized in Figure 3-1. Because of the problems cited in Figure 3-1, this system was rejected.

As discussed in Section II, a combination of PN coding, frequency hopping and time hopping forms an attractive way to provide the required processing gain. We desired to cover a range of data rates from approximately 100 bps to 1 Mbps corresponding to a variation in theoretical processing gain from 69 dB to 29 dB when 800 MHz of spread spectrum bandwidth is used. At low data rates, all three methods of obtaining processing gain can be used. However, at high

REPEAT REQUEST SCHEME	
Advantages	**Disadvantages**
1. The data rate is continuously variable.	1. An analysis has shown that this technique does not provide the desired linear trade as data rate is lowered.
2. The data rate is adaptive to the jamming level.	2. Processing gain does not increase as the data rate is lowered. This is contrary to one of the fundamental objectives of the program.
3. The adaption is automatic.	3. This type of system can not be used for high data rates because the propagation delays between the two terminals do not allow corrections to be made fast enough.
4. If a perfect back link is assumed, there are never any final message errors.	4. The error rate on the back link must be low. If this method is used for two-way data transmission, only the path with the good back link will provide acceptable performance.
	5. If two-way data transmission is required, the transmission rate must be increased so as to also provide feedback from the receiver to the transmitter.
	6. The system is too complex to be practical for M-ary signaling.

FIGURE 3-1: CHARACTERISTICS OF REPEAT REQUEST SCHEME

data rates at least one of the contributors to processing gain must be eliminated. We found no practical method for varying the processing gain and data rate over this large a range. A more realistic approach consisted of varying the data rate over separate, narrower bands for various types of data such as video, control data, etc.

The data rate needed to control an RPV is in the range of 50 - 1000 bps.[2] Similar data rates are expected for the weapon delivery problem. Compressed video data from many unmanned tactical platforms requires a data rate of 50 kbps to 1 Mbps. Between these two ranges of data rates lies a third potential range encompassing voice and other sensor link applications. The system design was thus reduced to one which could transmit data in each of these three data rate bands. For each band, processing gain would be obtained by some combination of PN coding, frequency hopping, and time hopping. Within each band, there should be a linear trade between data rate and processing gain.

The practicality of varying the data rate continuously rather than in steps within each band was then considered. It was found that a continuously variable data rate causes an enormous number of problems in implementation. Data rate changes quantized in 2:1 (3 dB) steps are much easier to accomplish and provide a reasonably fine adjustment. The data link system design covers a 16:1 range of data rates within each of the three data rate bands.

3.1.4 Methods of Varying Data Rates

3.1.4.1 Control Data

Since the control data rate is low (less than 100 bps), PN coding, frequency hopping and time hopping can all be used to provide processing gain. The parameters described in Section

3.1.2 have been selected to be optimum and cannot be varied without degrading performance or increasing costs. If one pulse is transmitted every frame and if each pulse represents one bit, then the data rate is 763 bps. This is near the upper limit of the control data range. The problem now is to devise a method of lowering the data rate in steps of 2:1 with a corresponding increase in processing gain without changing any of the basic system parameters and without any indication of data rate or changes in data rate from the signal externals. Although there are some losses, post-detection integration of pulses is a simple method of lowering the control data rate. The processing gain increases almost linearly for a small number of pulses. For integration of 16 pulses (frames of data), the increase in processing gain is 10 dB rather than the ideal value of 12 dB.[3] Therefore, the decision was made to use post-detection integration to vary the data rate. Note that an external observer is unable to determine what the actual control data rate is because the transmission appears to be the same at all data rates and there is no indication of the number of pulses being integrated.

3.1.4.2 Video Data

Because of the high data rates (up to 1 Mbps) required for transmission of compressed video, continuous or nearly continuous transmission is required. Therefore, CW transmission should be used to maximize the average power output, and no processing gain can be obtained by time hopping. Processing gain is still available from frequency hopping and PN coding. Since transmission is continuous and we want to avoid follower jamming, the frequency must be changed faster than a jammer can respond or about every 10 μs. A fast-switching synthesizer is thus required; this can significantly impact the system cost. Some additional processing gain can be obtained by letting each bit be represented by an M-chip binary PN

sequence. A further advantage of this method is that the value of M can be varied to change the data rate. This scheme is desirable because the processing gain varies linearly with the value of M. A 16:1 range of data rates can be obtained for values of M from 8 to 128. The resulting data rates lie within the desired range. Note that binary PN modulation produces a signal which is essentially the same as the control data signal so that the transmitter design is simplified.

3.1.4.3 Voice or Medium Rate Data

For this intermediate range of data rates, frequency hopping will again be used in this case to provide processing gain. A pulsed transmitter can be used for data at these rates so that time hopping provides some processing gain. The best method of varying the data rate over the desired range is to let each of the 128 PN coded chips in a pulse represent a different bit and then to add, in a post-detection integration fashion, the outputs of corresponding chips in m successive pulse transmissions. As was the case with the control data for values of m from 1 to 16, the processing gain increases almost linearly as the data rate is lowered. Binary PN modulation is convenient for the 128-chip pulse and is consistent with transmission at the other data rates.

3.1.5 Timing

Several of the tactical applications envisioned require multiple-data-rate transmission. Reconnaissance missions and RPV applications typically need low-data-rate transmissions from the command terminal to one or more vehicles. Control/status data and frequently video data must be transmitted from the vehicle(s) to the command terminal. Applications may require voice-rate transmission between terminals in addition to control and/or video data.

Full duplex transmission of data at different rates is needed.

In our analyses, several possibilities were considered. For instance, two terminals may transmit and receive only control data. Since the data rate is low and each transmitter is on for only $\approx 1\%$ of the time, transmitted pulses seldom overlap. Error coding can effectively reduce the probability of error when infrequent pulse overlaps do occur. At such a low duty factor, a small number of vehicle terminals can communicate with one command terminal with only a minor degradation in performance. Voice-rate data is also transmitted by low-duty-cycle pulses so that both control and voice-rate data can be transmitted between the command terminal and a few vehicles with minimum self interference.

On the other hand, video data requires continuous transmission. Hence, a problem of self-jamming occurs when video data is being received from more than one source. This problem could be solved by using different frequency bands for each source and multiple receivers at the command terminal. This increases the complexity in the equipment of the command terminal and also requires a more complex frequency synthesizer, one of the most expensive units in the system, at both ends of the link. Therefore, the data link system design described here will consist of only two terminals with control and voice data being transmitted on the forward link (command terminal to slave) and control, voice and video data being transmitted on the back link (slave to command terminal). Figure 3-2 shows this concept.

There are also certain timing problems associated with multiple data transmission between two terminals. Consider first the basic timing in the two terminals. In order for full processing gain to be realized from time hopping, the receiving terminal must know when to expect the signal. Hence, the two terminals must be synchronized in time. This is accomplished by establishing the command

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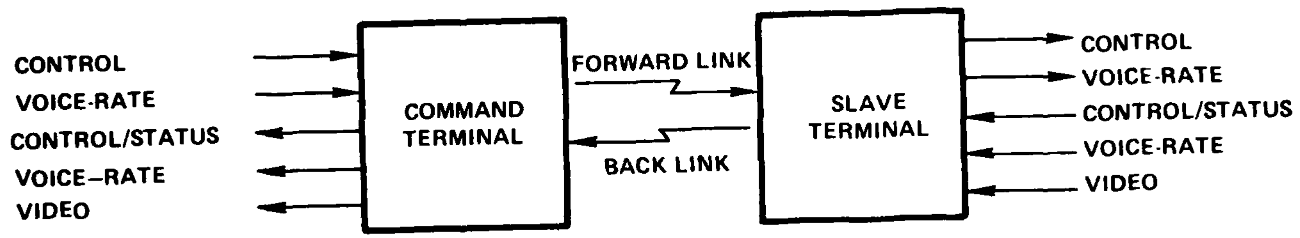
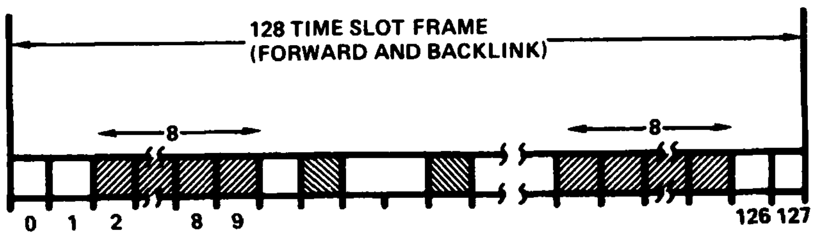


FIGURE 3-2: TWO-TERMINAL DATA TRANSMISSION

terminal transmitter as the basic timing reference for the system. The slave terminal timing for both its transmitter and receiver is locked to the timing of the signals which the slave receives from the command terminal. The command terminal receiver is then locked to the signals it receives from the slave terminal. Thus, the timing in the slave terminal will be delayed from the timing in the command transmitter by the propagation delay between the two terminals. Likewise, the timing of the command receiver is delayed from the slave terminal timing by the same propagation delay.

Consider next what can be done to minimize the probability of data overlap. At the slave terminal the problem is simplified because the slaved timing concept provides common forward and backlink timing. The time slots during each frame allocated for control and voice-rate data are known, and these transmission slots can be declared illegal *a priori* for video transmissions. The resulting effect is to decrease the allowable video data rate slightly. In each frame (128 time slots), one time slot is allocated for voice-rate transmission. Control data is transmitted using pulse-position modulation for M-ary signaling of digital data ($M=2^n$, n = number of bits per symbol transmitted). For 3-bit symbols, (n = 3, M = 8), eight different time slots are set aside for the eight possible symbols to be transmitted each frame. To protect against sophisticated jammers, each of the eight slots should be randomly sprinkled within the frame. For full duplex transmission of 8-ary control data plus voice-rate data, the slave terminal must reserve 18 time slots in each frame for non-video transmission as shown in Figure 3-3. In this way, there will be no conflict at the slave terminal between transmission and reception of the different types of data.

Unfortunately, there is no way of preventing simultaneous transmission and reception in the command terminal. Because of the



= TIME SLOTS ALLOCATED FOR 8-ARY CONTROL TRANSMISSION

= TIME SLOTS ALLOCATED FOR VOICE-RATE TRANSMISSION

= TIME SLOTS (TOTAL = 110 = 128 - 18) REMAINING FOR VIDEO DATA TRANSMISSION FROM SLAVE TO COMMAND

NOTE: THE TIME SLOTS FOR 8-ARY TRANSMISSION SHOWN ABOVE ARE GROUPED IN TWO CONTIGUOUS BLOCKS FOR SIMPLICITY IN THE PRESENTATION ONLY. ALL 18 SHADED TIME SLOTS ARE RANDOMLY DISPERSED THROUGHOUT THE FRAME.

IA-56,911

FIGURE 3-3: EXAMPLE OF TIME SLOT ALLOCATION AT SLAVE TERMINAL

propagation delays, the command terminal will probably be receiving video data when it is time to transmit control or voice-rate data. Refer to the timing diagram in Figure 3-4. The received video data will not be valid during the time when the transmitter is on. In order to minimize errors in the video data during these times of self-interference, error coding can be used on the video data. Since the time of potential errors is known, erasures can be declared. This leads to a selection of error-correction codes with good erasure capabilities to be used with the video data. Whenever the command terminal transmitter is turned on, erasures are declared for the received video data. There are still several problems with this scheme. The erasures on the video occur in bursts which are one pulse width (time slot) in duration or up to 16 successive video data bits. Since error coding will not correct long bursts of errors or erasures, interleaving of the video data must be included after encoding to sprinkle a burst of errors over several received words. If the interleaving pattern is fixed, a pulsed jammer could selectively jam the video data at the most opportune times. Therefore, the interleaving must appear to be random. Because of the added redundancy inherent in error coding, video data rates will be reduced to values somewhat lower than those desired.

3.1.6 Error Coding

The purpose of the variable-data-rate features in the link is to maintain the highest data rate which still meets the desired error performance. Acceptable performance for the low-data-rate mode is application dependent but would probably require error rates of less than 10^{-5} (possibly 10^{-4}).

In the presence of only broadband Gaussion noise, the data link operation is not significantly influenced by the use of error coding unless very powerful codes (with the associated expense) are

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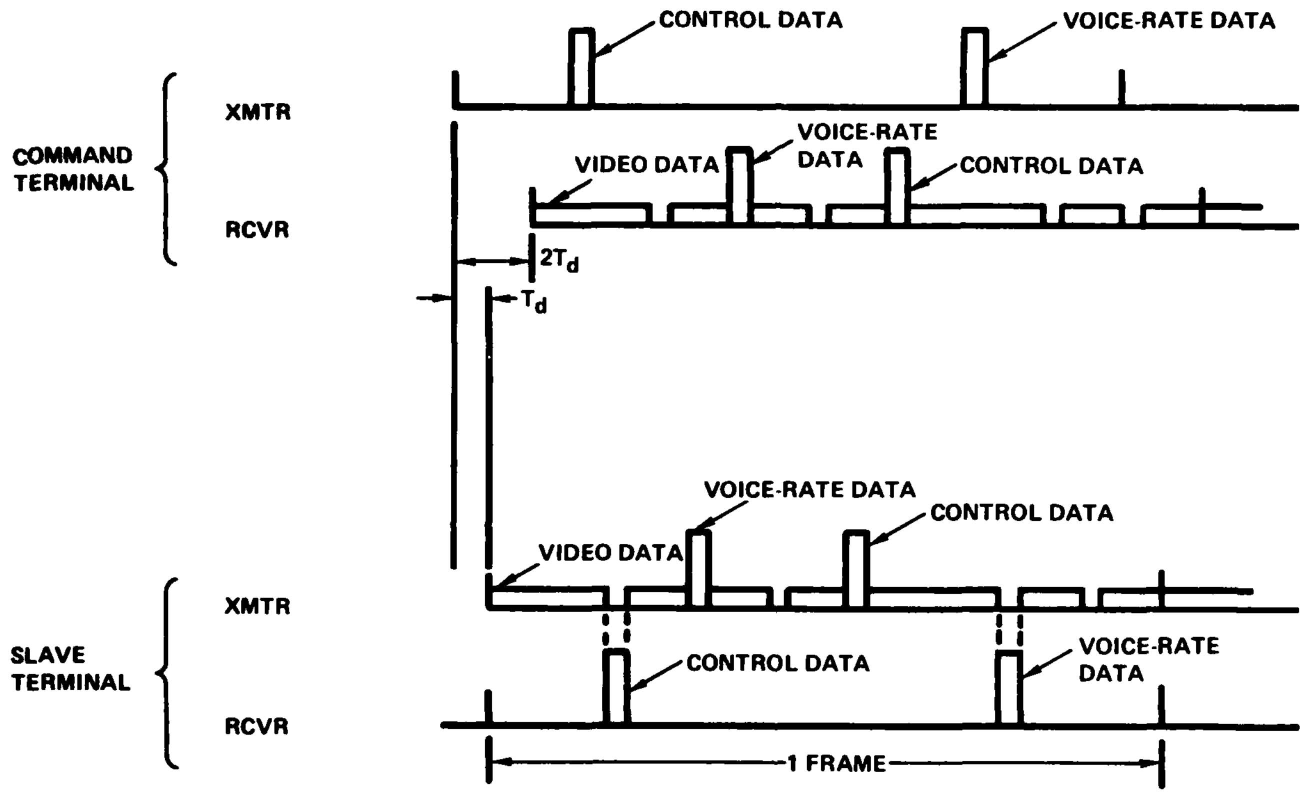


T_d = PROPAGATION DELAY TIME BETWEEN TERMINALS.

FIGURE 3-4: SYSTEM TIMING

incorporated. With the possibility of more adaptive jamming, (such as partial-time or partial-band jamming), the use of even relatively simple error codes can provide substantial benefit. Short block codes selected from the Bose-Chaudhuri-Hocquenghem (BCH) and Reed-Solomon codes were good candidates for our design.

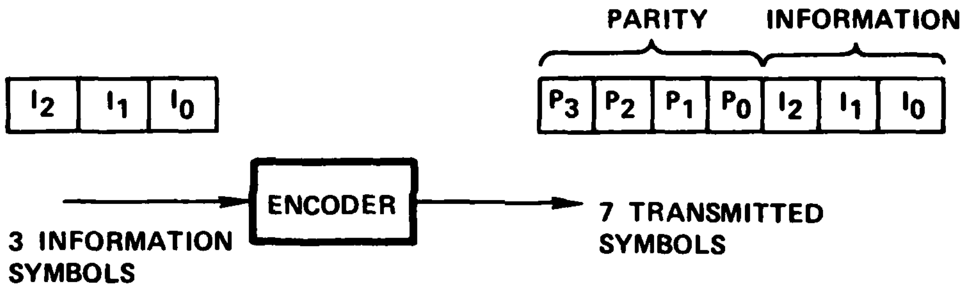
The Reed-Solomon (7,3) code was selected for use in the low-data-rate mode. Its significant parameters are summarized in Figure 3-5. Among the variety of reasons influencing its selection were:

- The 8-ary symbol is compatible with the basic signaling scheme
- Simple and potentially low-cost decoding techniques are available [4]
- An analysis of its performance indicated impressive capabilities in context of partial-time and/or partial-band noise

The probability of error requirements for compressed video data are not as stringent as those for control data, especially for only moderate amounts of compression. The residual redundancy in tactical reconnaissance pictures generally allows for acceptable picture reconstruction with error rates up to 10^{-3}. Nevertheless, the problems of simultaneous transmission and reception in the command terminal necessitate the use of a code which can handle erasures as discussed in 3.1.5. Because the video data rate is high, decoding must be handled in hardware rather than in software. The complexity of implementing the decoding algorithm for the Reed-Solomon (7,3) code in hardware prohibits its use. On the other hand, the BCH (7,3) code provides erasure handling capability and can be decoded with a reasonable amount of hardware. Since encoding is also extremely easy to implement, the BCH code is used for video data.

Voice-rate data may or may not require low error rates. Voice transmission itself is not degraded appreciably unless the error

NUMBER OF TRANSMITTED SYMBOLS IN A BLOCK N = 7
NUMBER OF INFORMATION SYMBOLS IN A BLOCK k = 3
NUMBER OF BITS/SYMBOL = 3 (=k)
NUMBER OF LEGITIMATE CODEWORDS = 512 = $(2^k)^k$
MINIMUM DISTANCE BETWEEN CODEWARDS = N - k + 1 = 5



ERROR/ERASURE CAPABILITY OF 7-SYMBOL WORD

NUMBER OF ERASURES	NUMBER OF ERRORS
0	≤ 2
1	≤ 1
2	≤ 1
3	0
4	0

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FIGURE 3-5: REED-SOLOMON (7, 3) CODE

rate is very high ($P_e \approx 10^{-1}$), so that the need for error coding is not only questionable but may be self-defeating.* However, if data other than voice data, such as sensor data, is to be transmitted at this medium data rate, error coding may be required. In addition to errors caused by a jammer, erasures will occasionally be introduced on the backlink when the command terminal transmits at the same time that it is trying to receive this medium rate data. The microprocessor-based Reed-Solomon decoder implementation, which is being used for control data, is too slow for use here. Several options are available, however. A simpler code such as the BCH (7,3) code could be used, or additional equipment complexity could be incorporated to perform the decoding of the more powerful codes at a higher speed. For the laboratory demonstration system, the added complexity of the error coder/decoder will not be incurred for the medium rate data.

3.2 Basic Design

Each terminal in the system consists of the basic units shown in Figure 3.6. The RF receiver removes the frequency hopping, amplifies and filters the signal, and outputs an IF signal to the receiver logic which demodulates and detects the data. The transmitter logic formats the data to be transmitted. This data is then modulated onto a carrier which is hopped in frequency. A random number generator is incorporated to select values for the following AJ parameters.

- Carrier frequencies which are generated in the frequency synthesizer
- Time slots for the control and voice-rate data signals
- PN codes for intrapulse coding

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* There is always some channel error rate above which error coding increases the probability of decoded errors

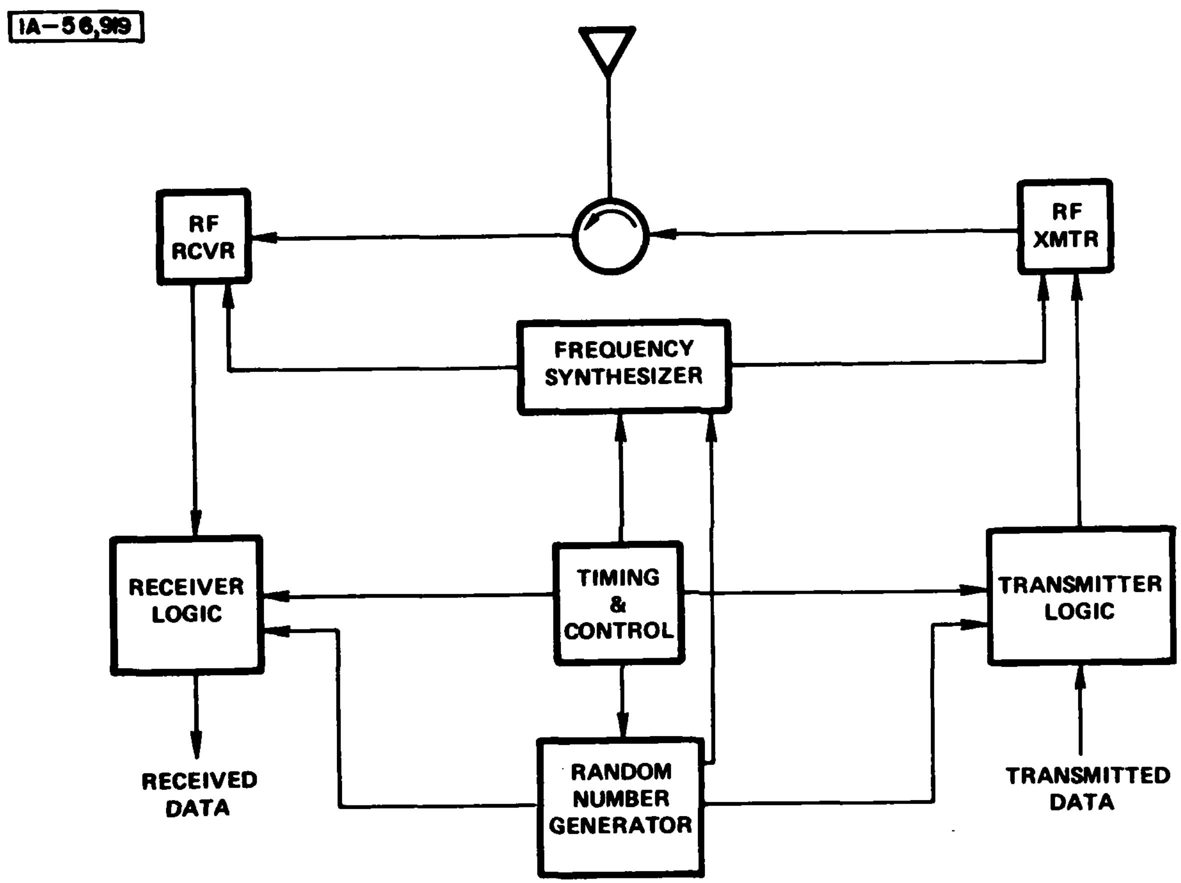

FIGURE 3-6: BASIC TERMINAL BLOCK DIAGRAM

The timing and control unit contains one or two oscillators plus the logic necessary to provide the required timing signals. The frequency synthesizer generates the RF/IF frequencies which are needed in the RF transmitter and receiver. The details of these system blocks are described in the following sections.

3.3 Signal and Data Processing Parts of the Data Link

3.3.1 Demodulator and Detector

3.3.1.1 Control Data

One symbol of control data is transmitted during one time slot every frame. Each transmission consists of 128 binary PN-coded chips. The time slot in which the pulse occurs determines which one of eight symbols (a symbol carries 3 bits of data) is being transmitted. The random number generator provides one particular 128-chip sequence each frame. The receiving terminal, which knows the selected sequence, contains a matched filter consisting of a surface acoustic wave (SAW) convolver.* The biphase modulated IF input signal is one input to the convolver. The other input is a biphase modulated IF reference signal with the expected PN modulation in reverse order. These two signals are injected into opposite ends of the convolver delay line and produce a correlation peak when the two signals are exactly coincident in the interaction region. The SAW convolver was chosen for this application because of its ability to perform the complicated signal processing function of matched filtering in a small package and also because the convolver has the potential for becoming a low-cost component.[5]

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* For a tutorial reference in the convolver, see reference[6]

After matched filtering, a decision must be made to determine which symbol was received. The convolver output is detected and then sampled by a 3-bit A/D converter. The sampling rate is two samples per chip. Corresponding samples for a block of m frames (m = 1, 2, 4, 8, 16) are added as necessary depending on the data rate selected over a 16:1 range. Then the amplitudes of the samples from all eight time slots are compared and a decision is made as to the most likely received 3-bit symbol. The relative chip amplitudes also provide an indication of the degree of confidence in the symbol decision. The Reed-Solomon decoding algorithm makes use of this additional information to perform simplified soft-decision decoding. This detection and decision process is shown in Figure 3.7. The time of arrival of the control data is received and used to synchronize the two terminals and to calculate the distance between terminals.

3.3.1.2 Video Data

Transmission of video data is in 10.24 μs bursts but is almost continuous, with gaps being left only to prevent overlaps with the control and voice-rate data at the slave terminal. Each data bit is represented by 8 to 128 binary PN-coded, phase-modulated chips. Thus, each 10.24 μs, 128-chip transmission contains 1 to 16 data bits.

Several signaling schemes were investigated and the cost/complexity impact of signal generation in the slave terminal and subsequent processing in the command terminal were considered. Some of the schemes considered most seriously were:

- Coherent matched filters for PN sequences
- Orthogonal signaling with matched filtering and non-coherent detection
- Chip-by-chip DPSK demodulation followed by digital PN sequence correlators.

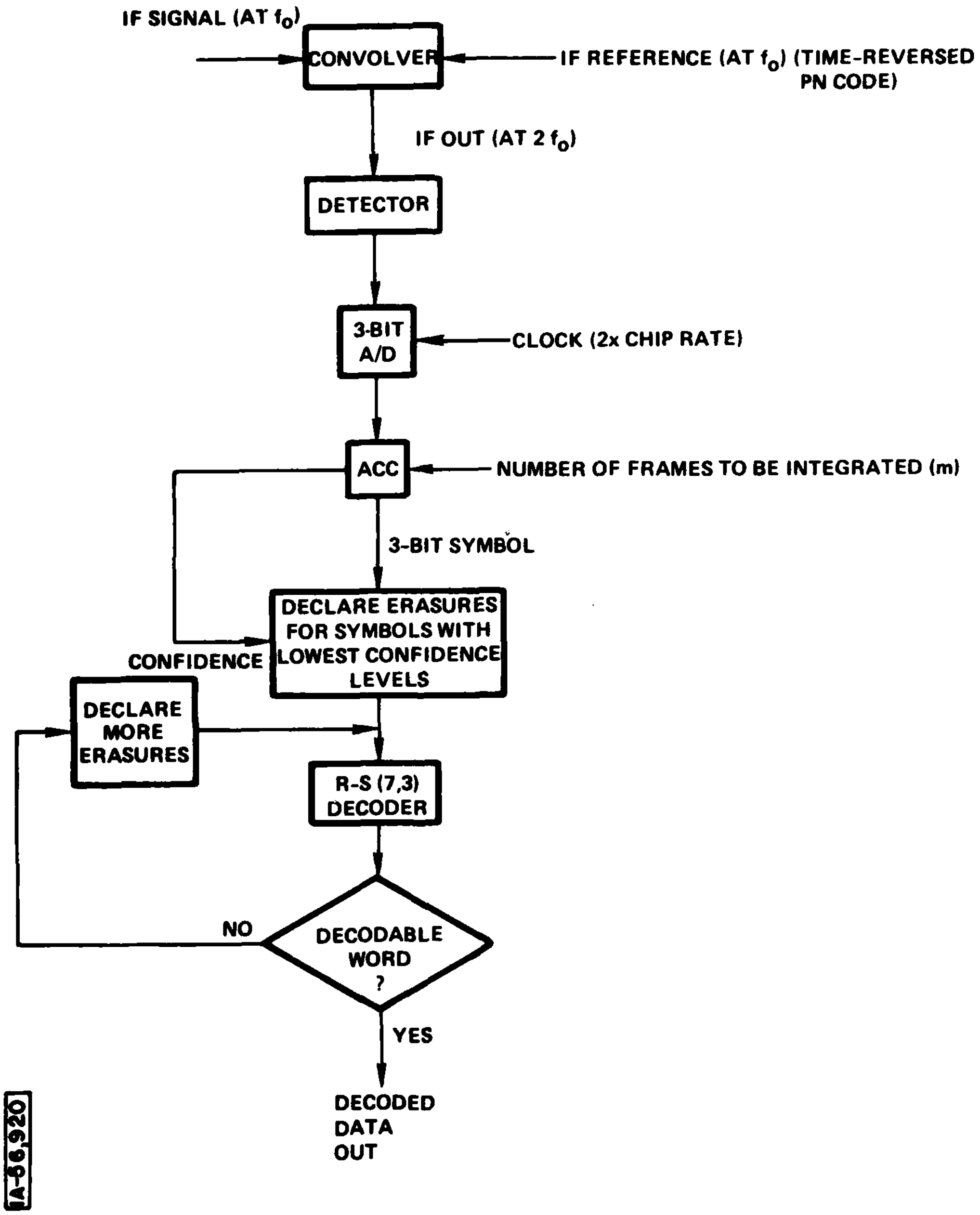

FIGURE 3-7: SYMBOL DETECTION AND DECISION FOR LOW-DATA-RATE MODE

Because of the frequency hopping, which is not coherent from one time slot to the next, techniques requiring coherence from time slot to time slot were ruled out. The second approach was ruled out because there is a loss in performance with non-coherent detection and because the matched filtering is expensive. The third scheme does not require coherence between 10.24 μs bursts but still provides good performance. This type of demodulator allows the use of complementary (antipodal) rather than orthogonal codes and reduces the number of required correlators to one. Differential encoding of the chips of the PN sequences is required but is simple to implement. Since phase changes occur every time slot, there is no phase reference for the first chip in each time slot. Hence, there will be a slight degradation in the first bit of each time slot for the higher video data rates. Because video data is transmitted only on the back link, only the command terminal contains a video receiver. Since low cost is not as important for the command terminal, a 3-bit A/D converter, rather than a simpler hard limiter, can be used to digitize the output of the DPSK demodulator and gain a slight performance improvement. Good timing for sampling (to within a fraction of a chip) is assumed to be available from the synchronization and timing/ranging features of the link. Correlation of the demodulator output with the expected PN sequence is handled digitally in a simple accumulator. The correlator output is compared with a threshold level, and a decision is made as to whether the data is a zero, a one or an erasure. The randomized interleaving is removed from the data, and the video data is decoded in hardware. This process is depicted in Figure 3-8.

3.3.1.3 Voice-Rate Data

Voice-rate data is transmitted by a 128-chip pulse once every frame. Binary DPSK transmission is again used with each chip representing a bit. In the command terminal receiver, the

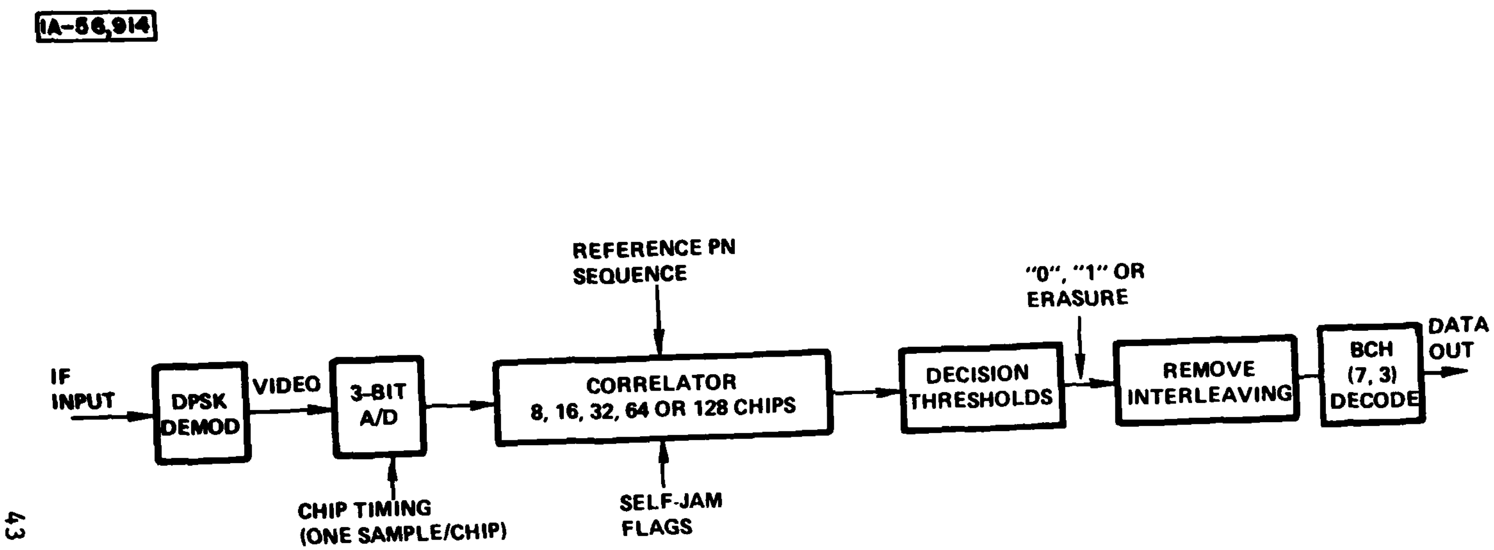

FIGURE 3-8: DEMODULATION AND DETECTION OF VIDEO DATA (COMMAND TERMINAL ONLY)

same A/D converter and DPSK demodulator are used for this function and for the video rate processing just described. In the slave terminal, where cost is important, a DPSK demodulator and a hard limiter are used. Depending upon the data rate selected (over a 16:1 range), the output of the demodulator may be used directly or summed for several frames (1, 2, 4, 8 or 16) before making a final decision on each data bit. Because of the lack of a phase reference for the first chip in each pulse, that chip is discarded. Other design problems require that the second chip also be discarded. Thus, 126 chips remain to be used as data bits in each pulse. This processing technique is shown in Figure 3-9.

3.3.2 Data Rate Control

A desired feature of this data link design is the ability to automatically adjust the transmission data rates as the jamming level changes. To do this, the receiver in each terminal must monitor parameters which indicate the probability of error performance. For the control data, the data rate and the desired probability of error are so low that it would take too long to make a decision using as a criterion the number of times the Reed-Solomon decoder was unable to decode a word. Instead of using indicators in the decoder, the confidence levels of the symbol decisions prior to decoding can be used to monitor control data performance. For the video data, the numbers of erasures and errors in the received 7-bit word are added for several seconds, and a decision is then made. For voice-rate data, the number of low amplitude bits at the demodulator output is an indication of the signal-to-noise ratio and, therefore, of theprobability of error. These performance indicators for control voice-rate, and video data are evaluated once every 4 seconds, and a decision is made to raise, lower or maintain each data rate. These decisions, made at the receivers in each terminal, must be communicated to the transmitter in the other terminal. This is accomplished

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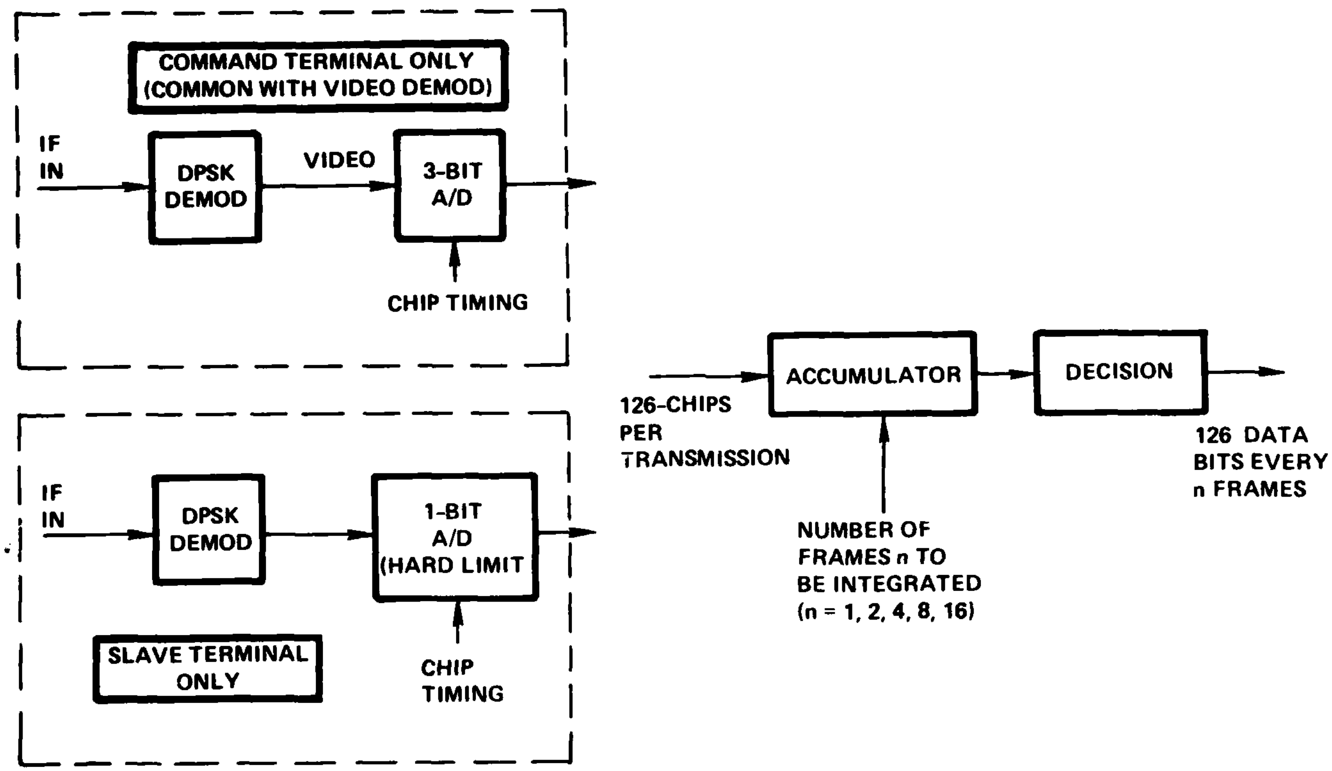

FIGURE 3-9: DEMODULATION AND DETECTION FOR VOICE-RATE DATA

by inserting the data rate information in the control data stream between the two terminals. The system also provides for manual selection of data rates.

3.3.3 Random Number Generator

The random number generator selects frequency-hop positions, time slots for the transmission of control and voice-rate data, and PN codes for control and video data. All of these parameters must appear to be randomly selected so that a jammer can not predict the next transmission parameters. The random number generators must also be synchronized so that they produce the same numbers in both terminals. In the data link work done under the IR&D program, the outputs of several very long maximal-length-feedback shift registers were combined in a nonlinear way to provide random numbers.[7] This sequential-logic method proved unreliable because a single error, which might be caused by noise, would be fed back in the random number generator, causing all subsequent numbers to be wrong. In the current system, we decided to use the new data encryption standard (DES) chip set which uses an algorithm approved by NBS. The implementation uses combinatorial logic avoiding the potential runaway problems inherent in the previous design. A key and a plaintext input produce a ciphertext output. Although the security of the algorithm has generated some controversy from a cryptographic viewpoint, the randomness of the output is more than sufficient for selecting the AJ parameters for the tactical missions envisioned here. For the laboratory model of the data link, the key will be hard wired in some random pattern, a simple binary counter will provide the plaintext input, and the ciphertext output will be the random numbers. Since there is no feedback of one ciphertext output to the next, the reliability should be greatly improved. Figure 3-10 shows this concept.

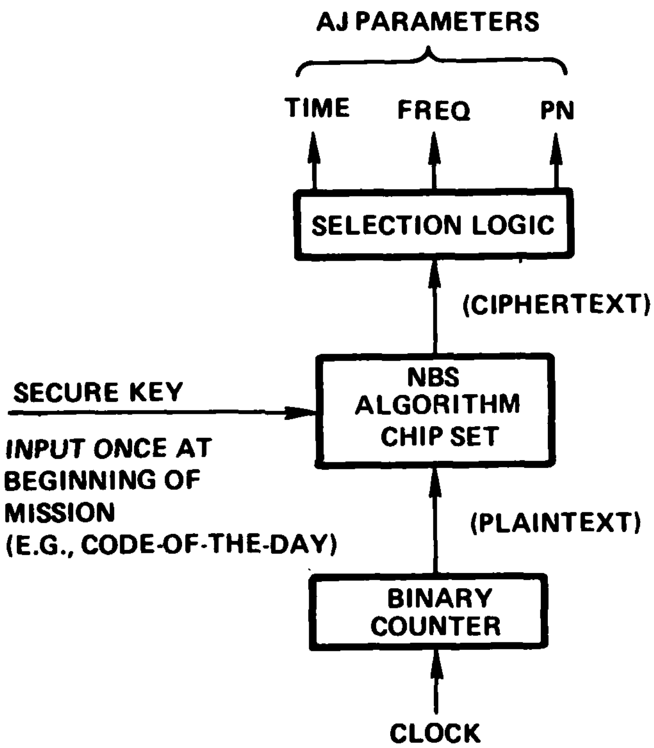

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FIGURE 3-10: RANDOM SELECTION OF AJ PARAMETERS BASED ON NBS ALGORITHM CHIP SET

3.3.4 Synchronization of Both Terminals

For this system to work, the two terminals must always agree on carrier frequencies, transmission time slots, and PN codes. The timing in the two terminals must be synchronized and the random number generators must be started together. The terminals must acquire synchronization when power is turned on or at the beginning of the mission, and if, for any reason (e.g., jamming) during the mission, synchronization is lost. The synchronization procedures for these two cases are different.

For initial synchronization, the random number generators are stopped, and fixed values are used for the random parameters. Only control data is transmitted during this phase. The slave terminal searches in time for the control data signal from the command terminal. Then the command terminal receiver searches in time for the control data signal from the slave terminal. After both terminals have been time synchronized, their transmission times are changed in a fixed pattern providing handshaking and timing countdown prior to starting the random number generators. The voice-rate and video transmitters are also turned on when the random number generator starts. This synchronization procedure is simple and also easy to implement, but it has no inherent jamming protection because there is no processing gain due to frequency hopping, time hopping or PN coding. It is assumed that this initial synchronization will occur in a relatively benign environment where the jamming level is low or non-existent. Under such conditions, this initial synchronization procedure is adequate. No practical synchronization method, which allows random hopping and can be easily implemented, appears to be available.

If synchronization is lost during the mission, the jamming level may be so high that repeating the initial synchronization

procedure will fail. However, because of the stability of the oscillators, the timing in the two terminals should not have drifted far apart as long as the distance between terminals has not changed significantly. The timing error will probably be less than a time slot.

The random frequency numbers, which change once a time slot, and the random numbers for control time slots and PN codes, which change once a frame, are still valid. Hence, there is no need to stop the random number generator, and there is no loss of processing gain. The resynchronization procedure adjusts the timing to look for the control data signal in adjacent windows within the same time slot. In most cases, this method will work. Only if it fails is the initial synchronization procedure attempted.

3.4 Analog Portions of the Link

The analog section of each terminal generates the transmitted RF signals and receives and processes the RF signals to the point where digital data is recovered. For our purposes, the analog section as shown in Figure 3-11 can be viewed as consisting of: the transmitter which contains a driver and power amplifier(s); the transmitter/receiver (T/R) section, which contains the duplexer, the receiver section and the transmitter modulators; and the frequency synthesizer, which generates the frequency-hopped signal carriers required for transmission and reception. The detectors for the received data signals, shown in the figure, have been briefly discussed earlier. The duplexer channels the transmitted and received signals to and from the antenna. The receiver processes the received signals prior to their detection. The transmitter modulators impress the PN phase coding on the transmitted carrier and pulse modulate the transmission.

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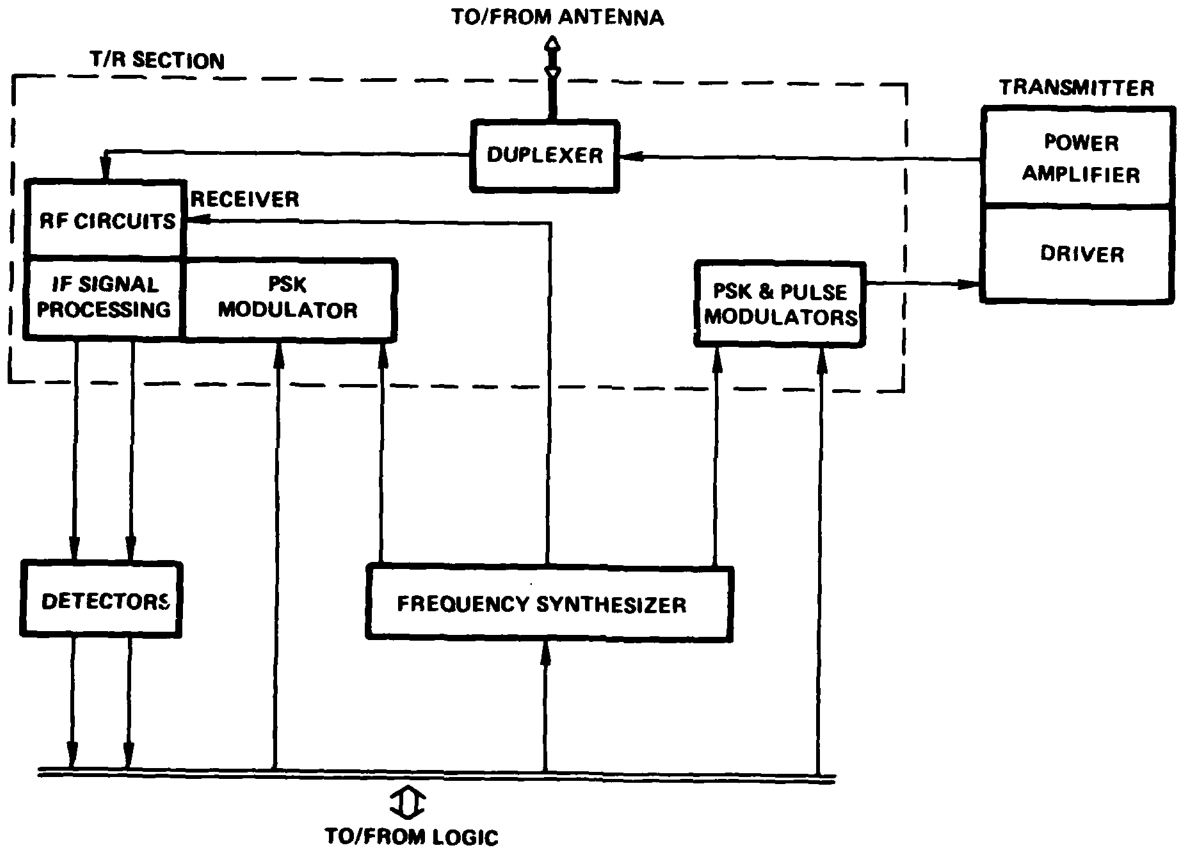

FIGURE 3-11: ANALOG PART OF DATA LINK TERMINAL

In this section we will examine individually the transmitter, the T/R section, and the frequency synthesizer of our proposed data link system. Based on our analyses and evaluation of technology, we will suggest some near-term and long-term possibilities in light of cost projections and performance capabilities for the three parts of the analog section. In Section IV we will present details of a laboratory demonstration design which yields simplified block diagrams of a 16-channel system, a 64-channel system, and a system-oriented explanation of the diagrams.

3.4.1 The Transmitter

In our analyses of transmitter effectiveness, we have used a figure of merit (FOM) to represent the relative jam resistance in dB.

$$\text{FOM} = 10 \log (E \times W_{ss}) \text{ dB}$$

where E is the energy per bit transmitted and W_{ss} is the maximum useable spread spectrum bandwidth. Maximizing this figure of merit while minimizing cost will provide the most cost-effective jam resistance against an average power-constrained noise jammer.

The ideal strategy for a single transmitter with a constrained average power would be to provide an FOM, or jam resistance, that is inversely proportional to the data rate. Ideally, we desire one low-cost transmitter that is capable of providing a variable peak power output with constant mean power, over a range of duty factor from unity (at the high data rates) to below 1% (at the low data rates). Practical considerations, however, limit peak power control to about a 6 to 10 dB range using tube pulse amplifiers. Solid state pulse amplifiers do not allow any significant range of power control. Thus,

a piecewise approximation to the ideal linear jam resistance versus data-rate curve has to be implemented at a minimum cost, and two or more transmitters will probably be required to obtain an acceptable AJ/data rate tradeoff over a wide range of data rates.

Since low data rate transmissions will normally consist of short (typically 10 μs) and infrequent bursts of energy, a transmitter with high peak power and low duty factor is required for this data. For higher (video) data rates, more continuous transmissions are required, and a CW transmitter of appropriate power is required for this data. The transmitted power level may also be constrained by prime power limitations and transmitter efficiency as well as by cost. For voice-rate data, optimum signaling might place intermediate requirements on the transmitter and a third transmitter might be used if the gain in AJ margin per dollar proved to be a good tradeoff.

Each transmitter must also be capable of handling an input signal, frequency-hopped at a rate high enough to prevent repeater-jammer interference and to allow efficient system operation at desired data rates. As noted earlier, this may require hopping in frequency every 10 μs, and, for the high data rate, CW signals may require settling times on the order of a chip time (80 ns).

While the results of the transmitter investigation have broad utility, the focus was on obtaining cost-effective approaches to the transmitter section of a slave terminal. Factors affecting the selection of candidate approaches were size, weight, reliability, and prime power requirements. For example, transmitters requiring more than a few hundred watts of prime power would clearly be impractical in most unmanned tactical vehicles and can be eliminated from consideration immediately, even though the technology may be well in hand.

3.4.1.1 Summary of Transmitter Results

Because the role of the RF transmitter is so important in the data link design and will impact so significantly on the resulting cost of an expendable terminal, a considerable amount of effort was devoted to examining transmitter options. These options along with estimated cost and FOM are discussed in Appendix A. The most promising approaches are summarized here.

The transmitters for the slave terminal were specifically considered in the analysis. The results obtained are also applicable to the command terminal which is not so constrained by cost considerations.

Low duty factor control data transmission requires a pulsed transmitter. The only practical options providing an adequate FOM are the traveling wave tube (TWT) and voltage tuned magnetron (VTM). The VTM is a relatively recent development. The tube itself has a large cost advantage over the TWT because of its simple construction, but the VTM transmitter is complicated by the need for a high-voltage tuning control not necessary with the broadband TWT. The design of a VTM transmitter with fast switching time (below 100 ns) does not appear to exist at present and will require considerable development effort. In addition, the attainable FOM of the VTM transmitter is about 9 dB inferior to that of the TWT. The VTM can be used in conjunction with its driver stage in a pulsed/CW mode transmitter. This advantage will be described later in this section.

For the CW transmitter required for the transmission of video data, the only practical near-term option is the TWT. This is a well-tried, low-risk solution for video data (and control or voice-rate data also, but providing a relatively small AJ protection), but it is expensive.

A promising transmitter approach using two transmitter tubes can provide a piecewise approximation to the ideal curve, $E \times W_{ss} \propto (\text{Data Rate})^{-1}$, at a lower cost than the sum of the two (independent) transmitters described above. The proposed near-term scheme uses a pulsed VTM driven by a CW/pulsed TWT. It requires the development of a high-voltage frequency-control circuit for the VTM. This approach provides both the high-peak-power pulses for low-data-rate transmissions and continuous operation (CW) for the higher data rates.

A long-term option for the two-transmitter approaches is to use the specially-developed pulsed/CW VTM, described previously for the low-power transmitter. This represents the lowest cost at a slight degradation in performance over the near-term option. We expect that this approach will give faster settling times for frequency hopping of the continuous, high chip-rate signals.

3.4.2 The T/R Section

The T/R Section must provide duplexing of the transmitted and received signals and protect the receiver mixer from transmitter power leakage. RF filtering of the input signal must be provided to protect the receiver from being jammed by out-of-band signals (e.g., at IF). The receiver mixer must have a single-sideband response (image-rejection circuit) to prevent the reception of a broadband noise jammer at two frequencies, which could degrade the jam resistance by 3 dB.

An image rejection of 10 dB is sufficient to reduce the jammer's advantage to 0.4 dB without requiring an expensive mixer. A low-noise receiver is not required since we are combatting high power jamming which will be well above normal receiver noise. Low-cost provisions should be made, however, to reduce the receiver noise level to a reasonable level (say 13 dB).

Since AGC cannot be used reliably in a jamming environment, a soft limiter is proposed to compress the signal dynamic range (about 40 to 50 dB) to a range that low-cost signal processors and detectors can handle (about 10 dB). A good limiter design will also protect against the negative effects (and potential advantages to a jammer) of impulsive interference (e.g., a partial-time partial-band jammer with a low filling factor of our TW space).

As described in Section 3.3.1.1, a SAW convolver is proposed to provide matched filter processing gain for the low data rates where a modest amount of PN coding will be used. While providing matched filtering for signals with a TW-product of 128 ($T = 10.24$ μs, $B = 12.5$ MHz), the convolver must accommodate some time-of-arrival uncertainties. Thus, its interaction region must be longer than th pulse length by perhaps a few microseconds.

Two binary-phase modulators are used for generation of PN-coded signals. One provides transmitter modulation; the other provides a reference waveform for the convolver.

A block diagram of the analog section showing more detail of the T/R section is shown in Figure 3-12. The conversion to a second IF is only necessary if the convolver frequency and the first IF are incompatible. If a single high IF were used at (say) 150 MHz*, the corresponding offset between the forward and backlink frequency bands would unnecessarily increase the bandwidth required of the antennas, or, conversely, the antenna bandwidth limitation would force reduction of the system bandwidth.** A low first IF (typically 50 MHz) is therefore proposed. The convolver reference waveform (PN modulated)

* Compatible with available convolver designs.

** As noted in Section 3.4.3, the forward and backlink bands are common except for a differential offset equal to the first IF.

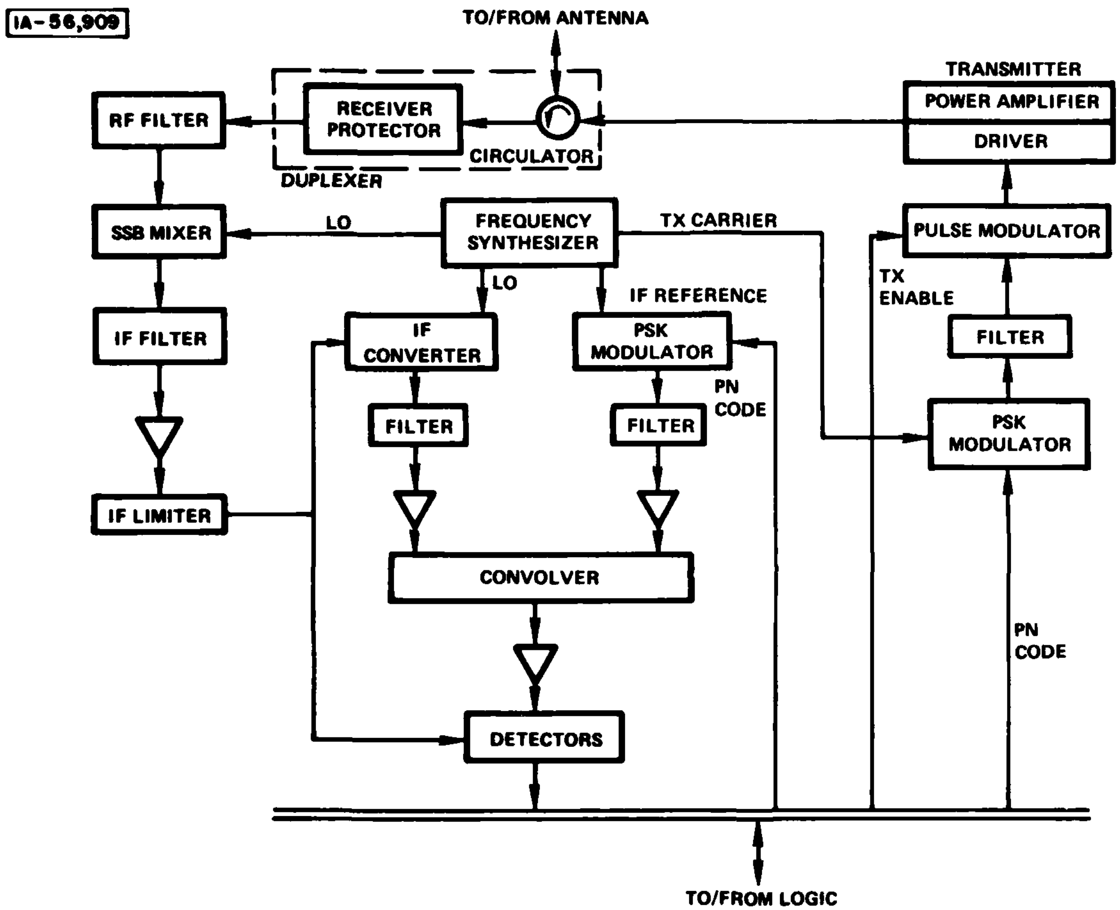

FIGURE 3-12: RF/IF ANALOG SECTION

is a time-reversed replica of the expected received signal (PN modulated at the second IF). The transmitter PSK modulator provides PN modulation of the transmitter carrier by codes provided by the random number generator. Both PSK modulators should have sufficiently low distortion (asymmetry and rise time), so that the convolver output at either the slave or the command receiver is not significantly distorted. The pulse modulator should provide sufficient carrier suppression between pulses to prevent a compromise in jam resistance or mission effectiveness due to transmitter carrier leakage. About 40 dB suppression of the carrier seems adequate. The convolver reference and signal preamplifiers and the post-amplifier must be designed so as to preserve adequate signal-to-noise ratio and to minimize distortion. Distortion is caused by input second harmonies (convolver output frequency) and by mismatch reflections at the input and output. The post-amplifier noise level, also, must be kept reasonably low.

3.4.3 The Frequency Synthesizer

The frequency synthesizer provides frequency hopping with transition times set by the maximum chip rate. Assuming that one chip can be lost during a hop, the frequency must settle to a new value in less than 80 ns at a 12.5 MHz chip rate.

To reduce the cost, the proposed system would use the same synthesizer at the slave terminal for transmission and for the receiver local oscillator function. The command-terminal synthesizer would be offset from the slave frequencies by the IF to allow two-way transmission.

In view of the wide bandwidth desired for maximum jam resistance and the consequent need for operation in I band, we took a fresh look at the possible design approaches. As explained in Section 3.1.2, a spread spectrum bandwidth of 800 MHz, comprising 64 channels of 12.5 MHz signal bandwidth, is a good tradeoff between transmitter and logic complexity and the degree of AJ protection possible. Figure 3-13 shows the tentative frequency spectrum and the expansion of a 16-channel design to the 64 channels. We plan to generate the 16 channels with 200 MHz bandwidth in the 600 to 800 MHz band. The output of the 16-channel portion of the synthesizer will be used as input to the 64-channel synthesizer extension to generate the full bandwidth. In addition to the spread spectrum bandwidth requirements on the synthesizer, we decided to see whether a reasonable specification on spurious output frequencies could be met without serious cost increases. A 20 dB figure can be justified based purely on efficiency of transmission and a 60 dB figure would clearly present severe technical restraints and certainly run up the costs significantly. A 40 dB specification was adopted as a cost/performance compromise.

Several basically different approaches to the synthesizer were considered. From the minimum cost standpoint, two favorable candidates emerged. Neither of these, however, is sufficiently developed to be considered for the near-term implementation.

The first low-cost technique and, at best, a long-range consideration, achives direct synthesis by using filter-bank selection of two (or more) frequencies. These frequencies are added (and possibly frequency shifted or multiplied) to provide the desired output frequencies. Figure 3-14 shows a 64-channel synthesizer using this technique. Where possible, SAW filter banks are used, since these

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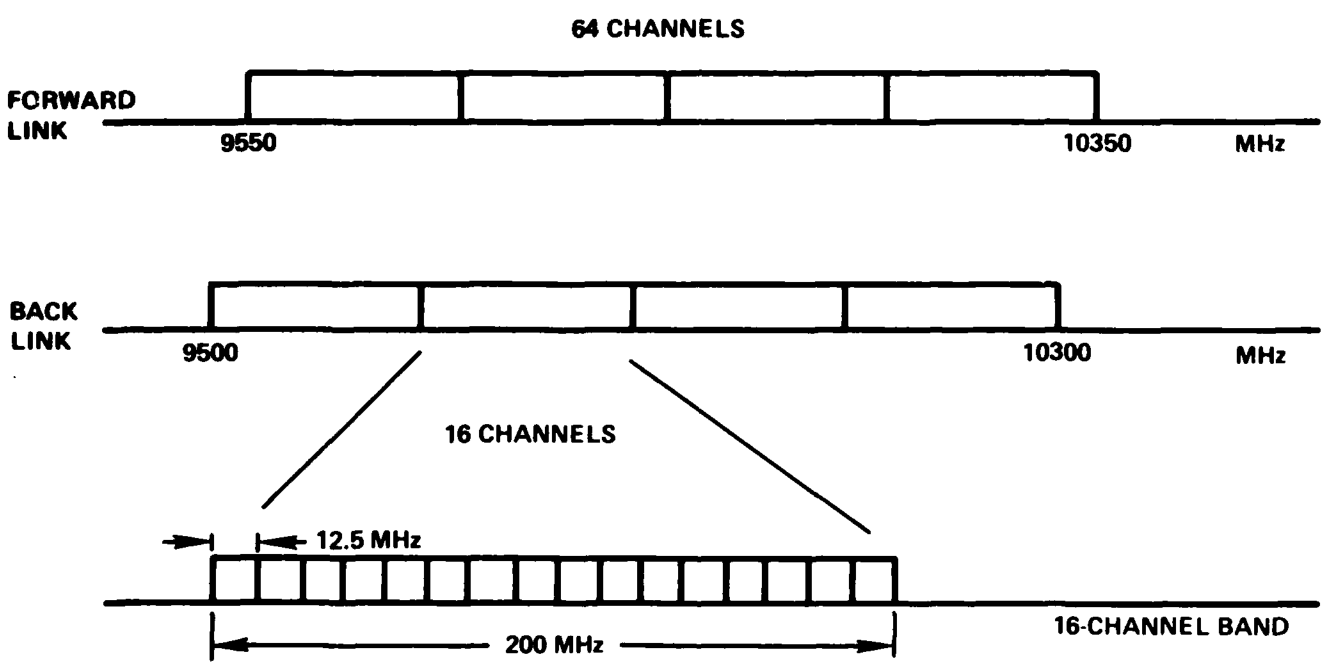

Figure 3-13. DATA LINK SPECTRUM

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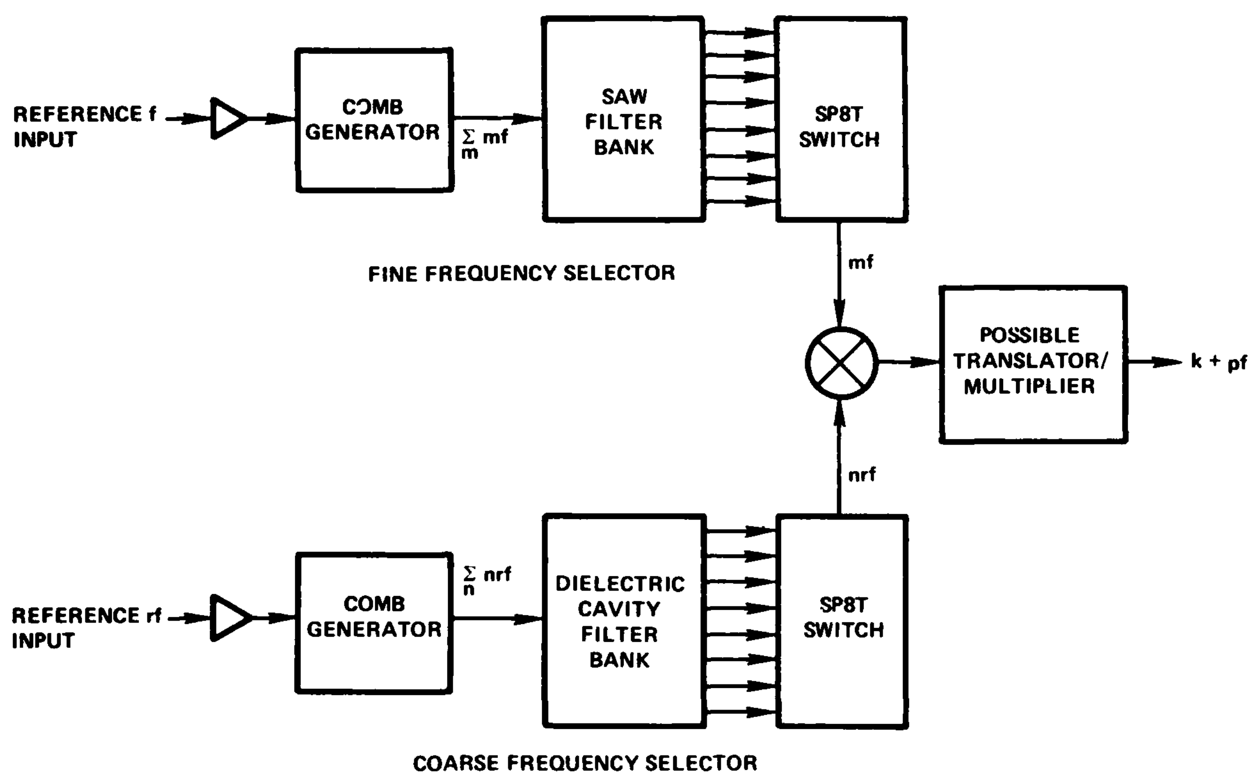

FIGURE 3-14: FILTER BANK SYNTHESIZER

have considerable potential for cost reduction. Work in this area is being conducted by RADC, partly in connection with JTIDS.[8] The fine frequency increments selected by a SAW filter bank can be accomplished at about 500 MHz, where good performance is readily obtained. A second SAW filter bank might be possible at around 1.3 GHz, where research efforts are currently being made. The scheme shown, however, uses a set of dielectric cavities for a second filter bank. These resonators would be made from temperature-compensated, high-K dielectric material that has recently been developed.[9] The use of such filters allows a higher frequency, say 4-5 GHz, to be used. Operation at about 9-10 GHz may also be feasible, and would eliminate the need for a frequency shift of the fine/coarse frequency sum. The result is cost efficient. Each filter bank is driven by a high energy comb harmonic generator, essentially a pulse train at the rate of the reference frequency.

The second low-cost technique uses microwave digitally programmable, phase-locked-loops (PPLOs) in the 1-2 GHz range. Microwave logic components are rapidly becoming available and developmental PPLOs have been demonstrated. In order to obtain the required rapid frequency switching (every 10 μs) with a short frequency-settling time (< 100 ns) two PPLOs are required, as shown in Figure 3-15. The frequency of the unused PPLO is changed in advance of the transition time. The minimum settling time for a PPLO is of the order of 1 μs for a reference frequency (channel spacing) of 12.5 MHz.

A third technique, which can be considered "brute force", calls for the combination of a set of basic frequencies from a clock reference to provide the required output channels. The combination is implemented by mixing and filtering, and many circuit arrangements are possible. For the near-term system, this approach appears to be the lowest-cost implementation.

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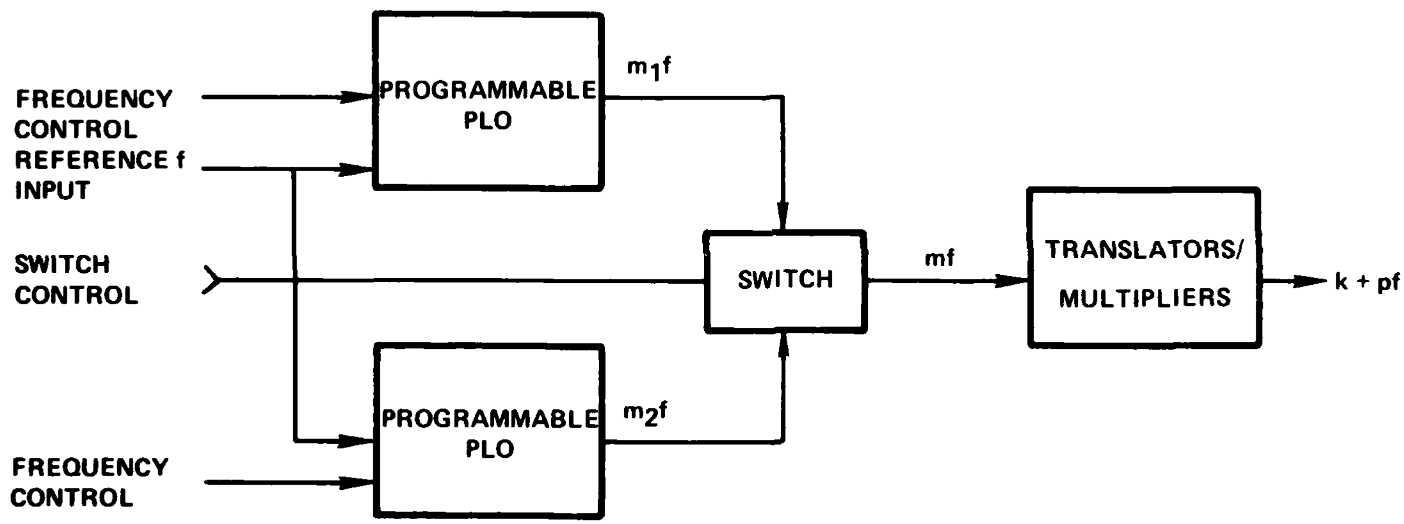

FIGURE 3-15: PROGRAMMABLE PLO SYNTHESIZER

The proposed arrangement uses three quaternary frequency selectors to provide 64 channels, as shown in Figure 3-16. This represents a relatively low-cost approach, although it may not be an optimum arrangement. Each selector uses three reference frequencies to generate four equally-spaced frequencies. The frequency spacings for the three selectors are one-, four-, and sixteen-channel spacings. Appropriate frequency offsets are introduced at convenient points in the circuit to facilitate filtering and to produce an I-band output compatible with the wideband transmitters to be used.

In a modification of the latter technique, the Fine Step Selector of Figure 3-16 can be replaced by a switched pair of PPLOs (as described earlier). Costs can be kept down by synthesizing at VHF, where very low-cost components are currently available. The relatively low frequency restricts the maximum number of frequencies from the PPLO to about ten, so that further selectors are required for, say, 64 channels. A number of circuit arrangements are possible, however, in a combined PPLO/direct synthesizer approach.

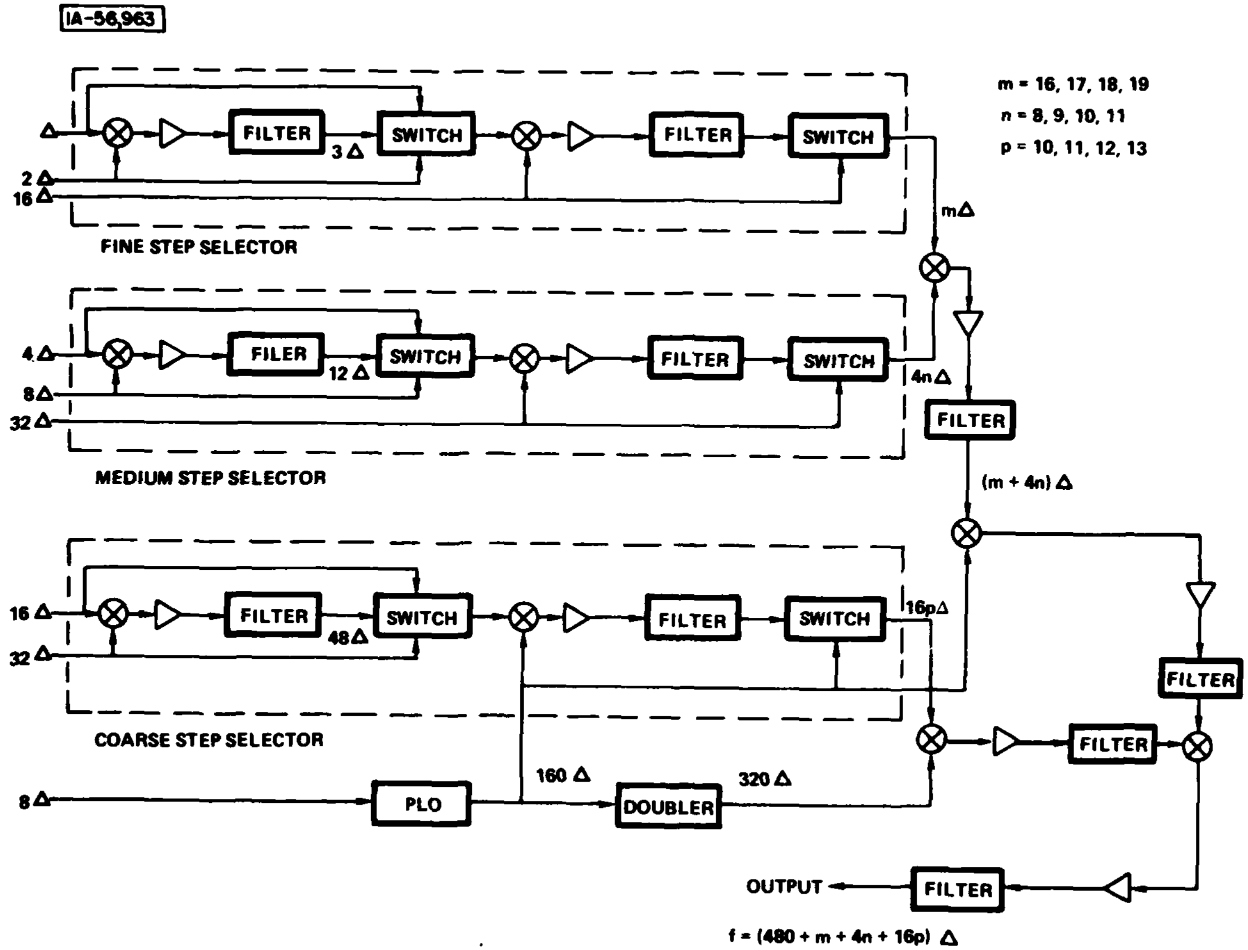

FIGURE 3-16: DIRECT MIX AND FILTER FREQUENCY SYNTHESIZER

SECTION IV

THE DATA LINK DESIGN: RESULTS

The requirement for a variable-data-rate/variable-AJ system has resulted in a design which is considerably more complex than a single data rate design would have been. Although some hardware is common for all three data rates, there are portions of the system which are used specifically for one band of data rates.

In this section, we present the two-terminal data link design that has resulted from the considerations given in the previous section. The details given here are basically for the laboratory brassboard model which will be completed and evaluated as part of the *FY80 program.*

Data from this design has been used in Section 4.2 to extrapolate a parts cost of the slave terminal assuming a production buy of 10,000 units. While careful use of these figures can give an indication of the expected procurement costs to the military, we see also that by examination of the cost distribution within the terminal we can learn much about both the cost of obtaining jam resistance and the current role of technology.

4.1 The Two-Terminal Design

The block diagrams of the two terminals in Figures 4-1 and 4-2 incorporate the design details discussed in the previous sections. Table IV-1 gives the data link parameters. From the figures, several differences between the terminals can be easily noted. One of the differences is the necessity for two clock generators in the command terminal but only one in the slave. The method of synchronizing the two terminals requires that there be a fixed frequency oscillator in

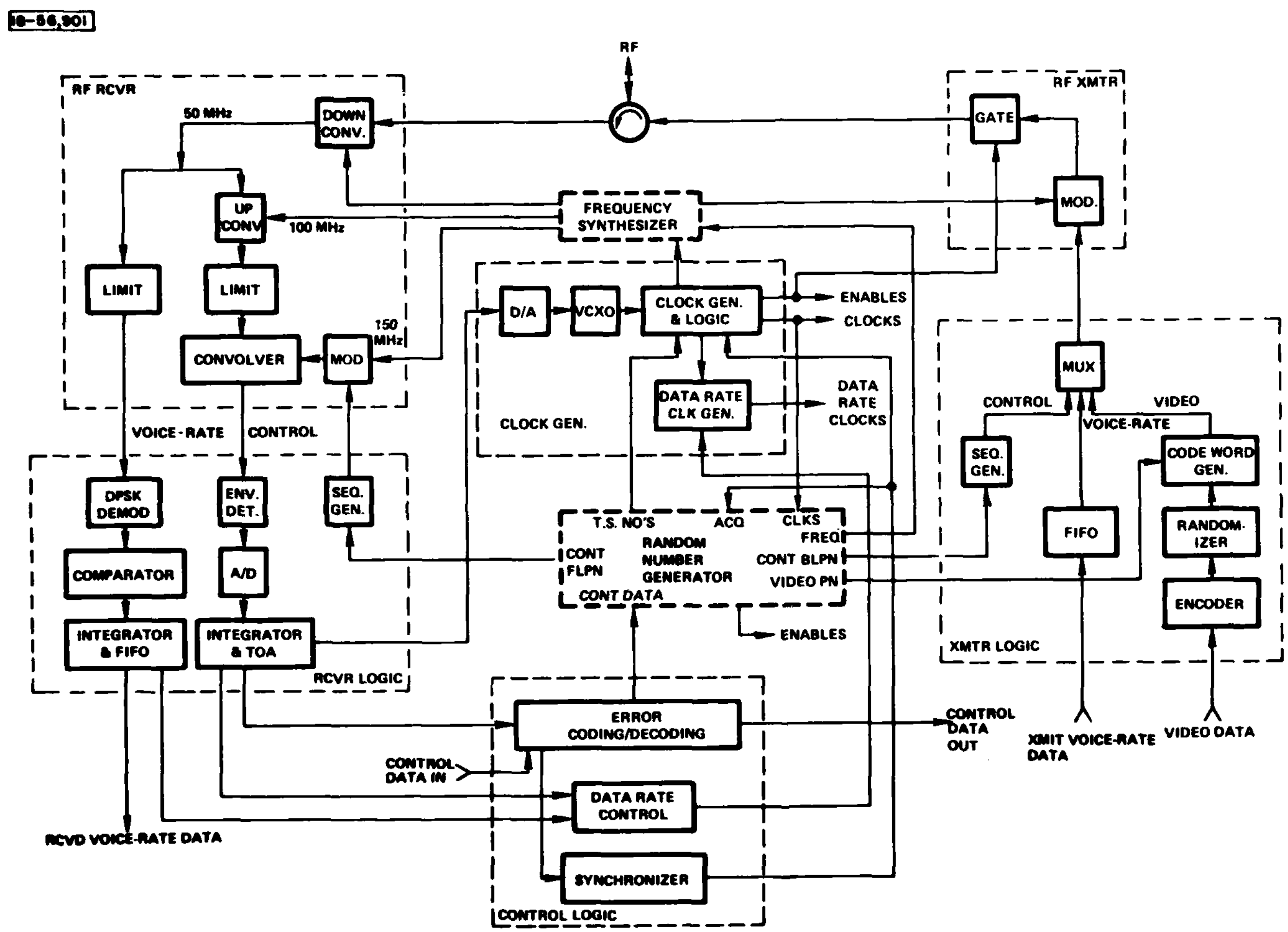

FIGURE 4-1 SLAVE TERMINAL BLOCK DIAGRAM

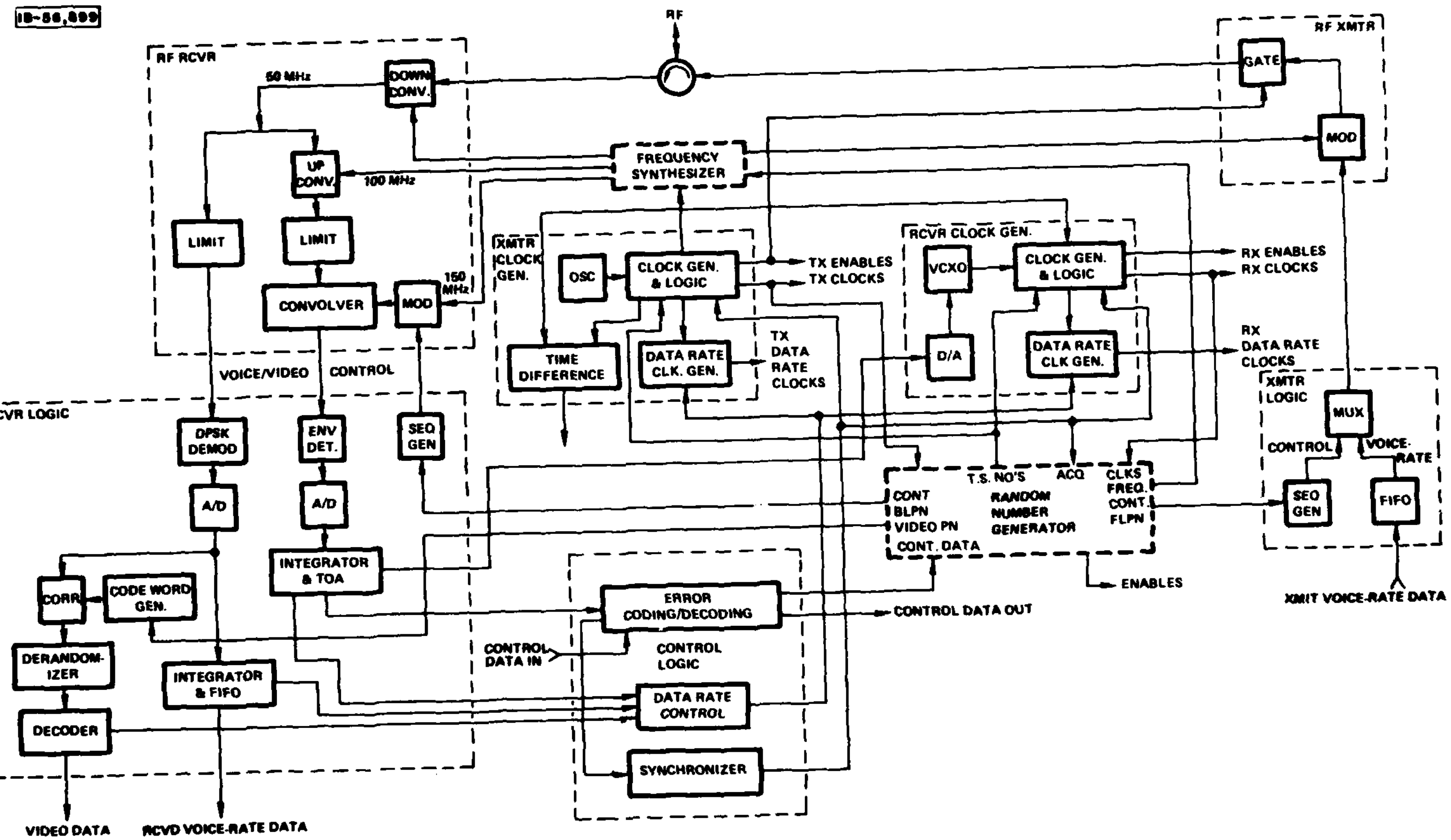

FIGURE 4-2 COMMAND TERMINAL BLOCK DIAGRAM

TABLE IV-1

DATA LINK PARAMETERS

CONTROL DATA	61.3,122.6,245.2,490.5, or 980.9 bps
VOICE-RATE DATA	6.0,12.0,24.0,48.1, or 96.1 kbps
VIDEO DATA	36.0,71.9,143.9,287.7 or 575.5 kbps
SPREAD SPECTRUM BANDWIDTH (64 channel system)	800 MHz
FORWARD LINK FREQUENCY BAND	9550 - 10350 MHz
BACK LINK FREQUENCY BAND	9500 - 10300 MHz
DURATION OF EACH FREQUENCY HOP	10.24 μs
CONTROL AND VOICE DATA PULSE WIDTH	10.24 μs
TIME SLOT DURATION	10.24 μs
PN CHIP RATE	12.5 MHz
NUMBER OF PN CHIPS PER TRANSMITTED PULSE	128
NUMBER OF TIME SLOTS PER FRAME	128

the command transmitter to act as the master timing for the system. The slave terminal contains a VCXO which locks the slave transmitter and receiver to the timing received from the command terminal. The command terminal receiver also contains a VCXO to lock its receiver timing to the slave transmitter. Another difference in the two terminals is the addition of video transmitter logic in the slave terminal and video receiver logic in the command terminal. Also, the command terminal contains the hardware required to measure round-trip propagation delay (distance).

4.1.1 Signal and Data Processing

The incoming RF signal is dehopped in the receiver. For control data signals, the 50 MHz IF signal is upconverted to 150 MHz and before being matched filtered in the convolver. The convolver is followed by an envelope detector and a 3-bit A/D converter. A 2903 bit-slice microprocessor (μP) performs post-detection integration (accumulation over several frames for variable-rate control data transmissions), makes symbol decisions and determines the time of arrival. Decoding of the R-S (7,3) code is handled by a 6800 MOS μP in the control logic. For voice-rate and video data, the 50 MHz IF signal is limited, demodulated and digitized. Depending upon the data rate, voice-rate data may be summed over several frames before bit decisions are made. In the command terminal, the video data path shares the same DPSK demodulator and digitizer. Then, each 8 to 128 chip sequence is correlated with the expected code word sequence, and a "1", "0" or erasure bit decision is made. The randomized interleaving is removed, and the video data is decoded.

At the slave terminal, video data is encoded, randomized, and multipliexed with encoded control and voice-rate data. Encoding in the command terminal is similar but there is no video data. This encoded PN sequence is used to phase modulate a frequency-hopped carrier.

A gate permits the transmitter to be turned off.

The clock generators provide the timing signals required by each terminal. Some of these clocks have a fixed frequency and others have frequencies which vary with the data rates. The frequency synthesizer generates the hopped carrier frequencies required by the transmitter and receiver.

The random number generator selects frequencies, time slot numbers and PN codes. The control logic includes the synchronizer and data rate controller as well as the encoder and decoder for control data.

4.1.2 RF/IF Analog Processing

The analog portion of the laboratory design does not contain the high power transmitter components. The transmitter chain is complete through the low-level RF drive point (output level ≈ 0 dBm). In addition, frequency hopping is being included in two steps. First, a 200 MHz bandwidth system at UHF will be completed and evaluated, then as part of the FY80 program, an extension to the full 800 MHz implementation in I band will be completed and evaluated.

Figure 4-3 shows a simplified block diagram of the RF/IF analog section. The laboratory design implements closely the conceptual design as described in Section 3.4.2. Separate limiters are used for the control data and the video data (at 1^{st} IF) to provide flexibility of design. The SAW convolver was provided with resistive padding to cut down reflections from mis-matches in impedance. Adequate amplification and filtering were provided to minimize distortion and to preserve an adequate signal-to-noise ratio. The ratio of signal-to-spurious and signal-to-harmonic levels should exceed 27 dB and the signal-to-noise ratio should exceed 23 dB (excluding the effect of post-detection filtering).

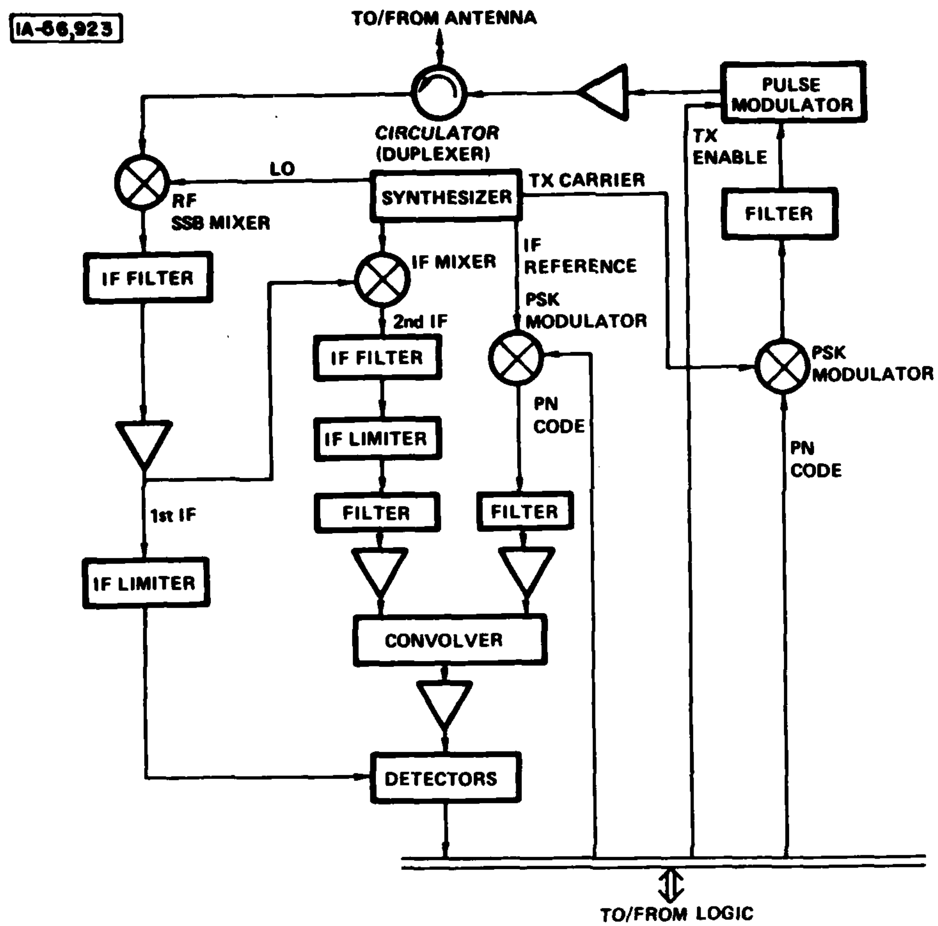

FIGURE 4-3: RF/IF SECTION (LABORATORY MODEL)

The planned extension of the system from 16 to 64 channels (200 to 800 MHz) in FY80 will be carried out by substituting the UHF RF components by I-band components, as shown in Figure 4-4. A quaternary frequency selector with appropriate frequency offsets will be added to the 16-channel synthesizer to convert it to 64 channels, along the general lines described previously and shown in Figure 3-16.

Figure 4-5 shows the basic configuration of the slave synthesizer, which follows the general principles previously described and shown in Figure 3-16, except that the coarse (quaternary) step selector and final frequency-converter stages were omitted. The command synthesizer is very similar, but contains circuits to provide the 50 MHz offset frequency required.

4.2 Costing of the Slave Terminal

In order to gain some insight into the potential cost impact to the military of procuring a data link of our design, we have estimated parts cost of the slave terminal assuming approximately a 10,000 unit production buy. The assumption has been made that the slave terminal would be needed in quantity for the expendable end of the link (if such an end exists). The command terminal would probably be associated with a high-value airborne platform or ground station and thus would be procured in much more limited quantities. Concern for the slave terminal cost being an overall system cost driver warrants the focus on the unit cost (and cost distribution) of that terminal.

With non-recurring costs, (such as development charges) excluded, a rough indication of the procurement cost of an electronics system like the slave terminal can be obtained by multiplying total parts cost by a factor of two to two and one half. From the preliminary parts cost from Table IV-2, we see that the resulting parts cost estimate for the slave

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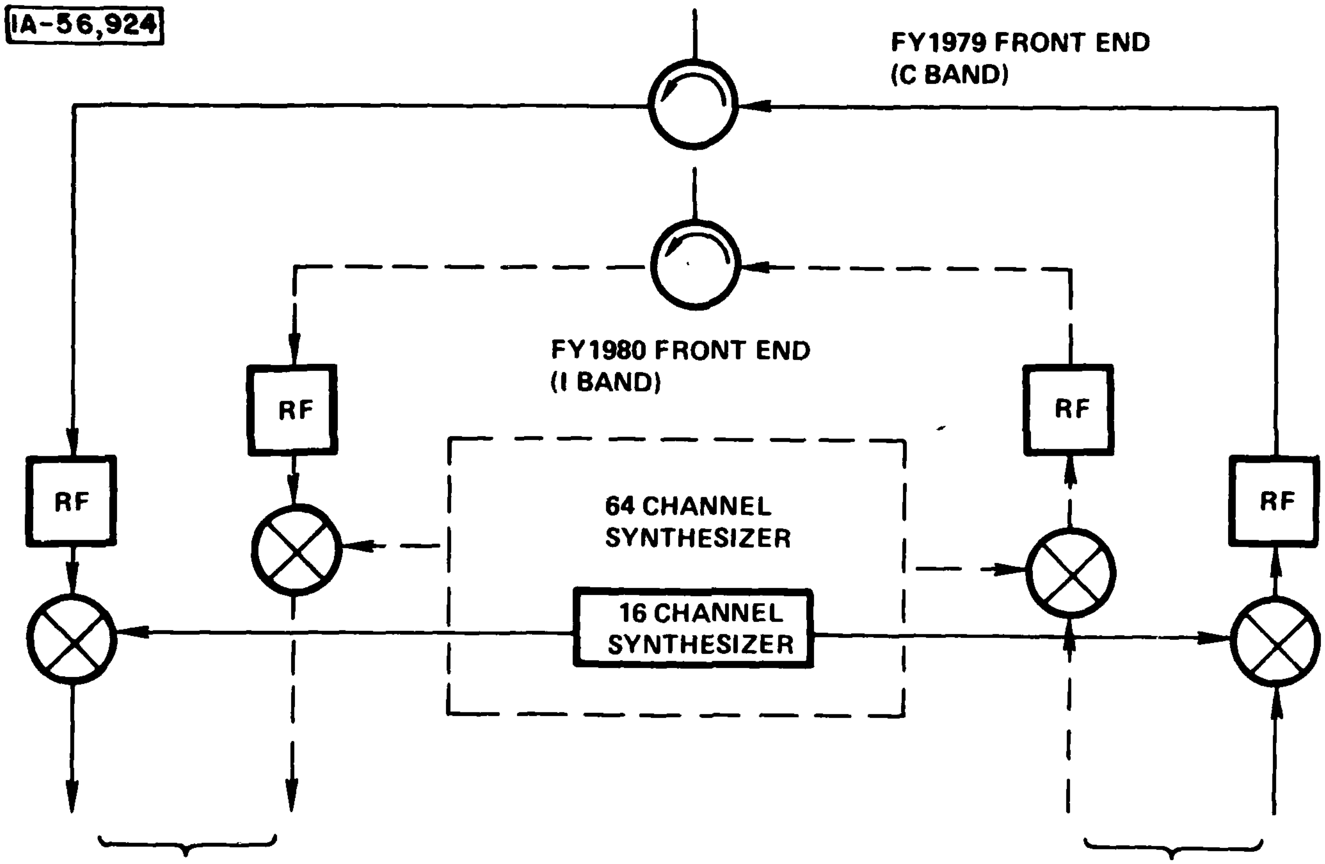

FIGURE 4-4: EXPANDABLE FRONT END

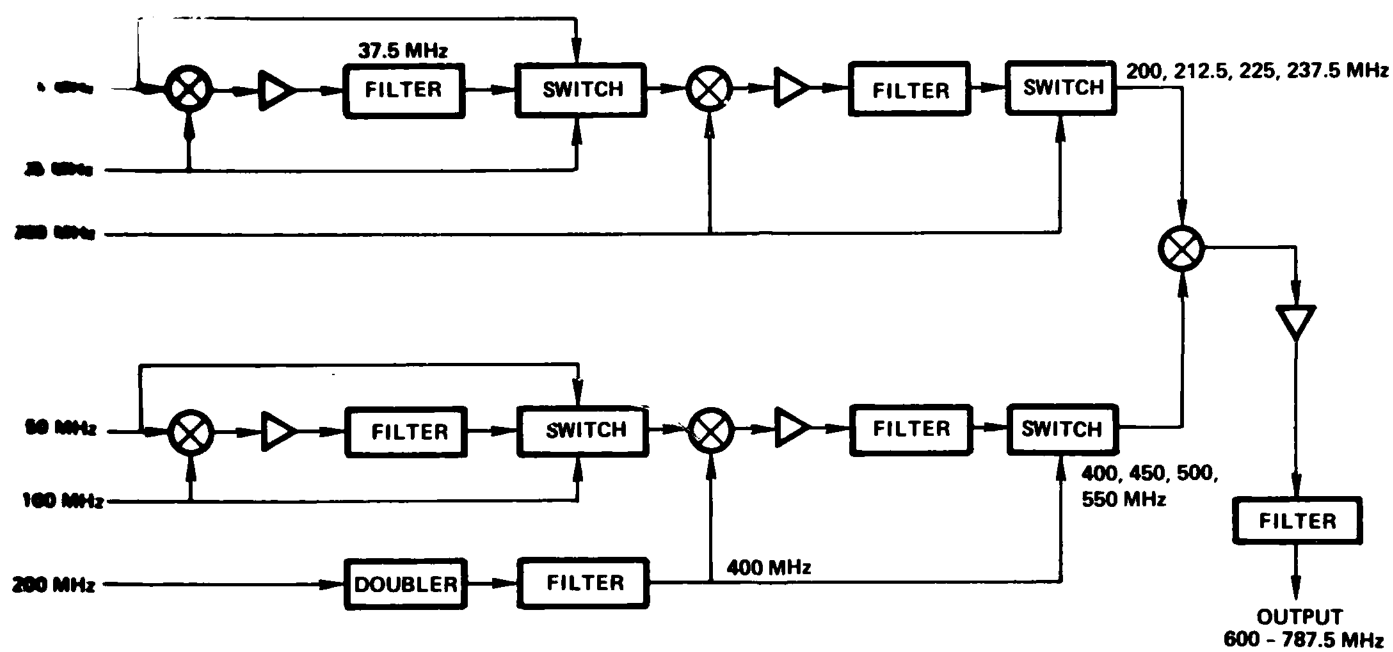

FIGURE 4-5: 16-CHANNEL SLAVE SYNTHESIZER (LABORATORY MODEL)

TABLE IV-2

SLAVE TERMINAL PARTS COSTS
(HIGH VOLUME)

BACK LINK TRANSMITTER	3200
RF/IF	860
FREQUENCY SYNTHESIZER	1940
MATCHED FILTER*	330
DETECTION, TIMING AND CONTROL	2030
CLOCKS	580
POWER SUPPLIES	290
MISCELLANEOUS HARDWARE	250
TOTAL	$9480

* Includes the convolver and its support circuitry needed for matched filtering in control data path only.

terminal is $9480. The full cost of such a production buy makes one skeptical on the reality of a committment to a 10,000 quantity procurement. Costs for a smaller quantity were not made, but unit parts costs would probably increase by ten to twenty percent for each factor of ten decrease in overall quantity. More important, however, non-recurring costs may begin to dominate the total cost and for small to moderate (less than 1000) quantities dedicated production processes and testing are not cost effective.

Interesting observations can also be made from the cost distribution of the terminal as shown in Figure 4-6. A large share of the cost is in the RF power generation area, reflecting the critical importance of this technology area to the overall jam resistance problem. The two other cost drivers are the frequency synthesizer and the logic area. We see that each of these significant cost items is driven by the combined need for substantial jam resistance (in terms of power-bandwidth product) and the full duplex operation of multiple data rates.

The RF power generation section will be a cost-critical area in any data link design, but several factors compound the cost issue here.

a. The need for wide bandwidth forces the center frequency to a region of high-cost components. Mass production techniques are not a factor yet in transmitter technology at these frequencies.

b. The desire for variable duty factor designs to provide maximum available jam resistance over the range of data rates stresses the technology.

c. The combination of high power and wideband frequency hopping with fast settling times needed for high data rate transmission limits the options.

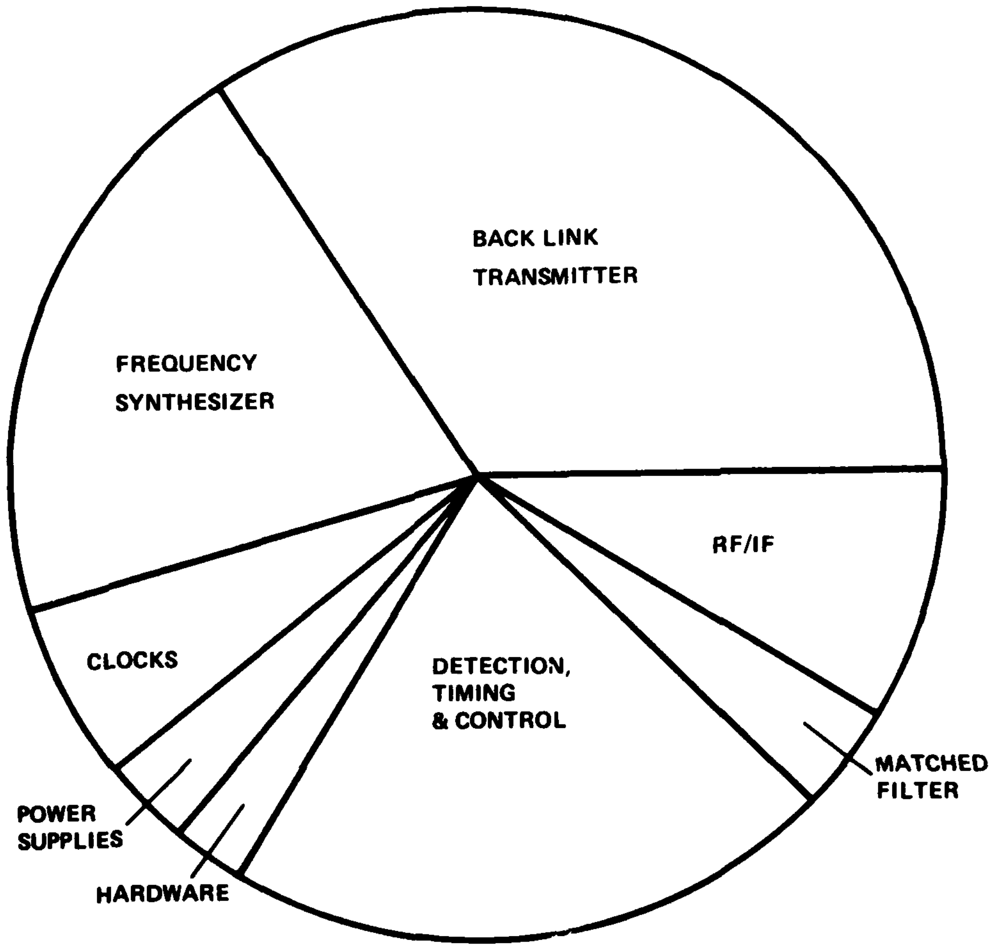

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FIGURE 4-6: RELATIVE COST OF SLAVE TERMINAL

The frequency synthesizer costs are affected by some of the same factors, too.

a. The wide bandwidth and high frequency band of operation increases both complexity and cost of the implementation.

b. The need for wider bandwidth and for fast settling times at the higher data rates impacts by eliminating some of the more attractive phase-locked oscillator approaches.

c. Spectral purity needs beyond that necessary for efficiency of transmission restrict component selection and require an increase in the number of components and their costs.

The cost of the detection, timing and control logic has increased significantly from our previous experiences. Again, a number of factors drive this cost.

a. Off-loading the processing functions for the multiple data rates from traditionally more expensive analog circuitry has been cost-effective in providing the variable-data-rate capability, but the implementation is still expensive. For example, the microprocessor and memory IC's required for post-detection integration of control and voice-rate data are significant contributors to the overall cost.

b. The complexity of handling multiple data rates in a duplex mode forces use of additional circuitry.

c. Even at the modest instantaneous bandwidths used, some of the logic circuitry needed is in the more expensive power consuming ECL line.

Technology is moving rapidly in a direction to cause more impact on the design and cost than has been accounted for here. For example, some of the logic circuitry used is relatively new and current learning

curve projections of quantity pricing may be far too pessimistic. Similar possibilities lie in the frequency synthesizer area where much work in high-speed analog and digital technology may provide a high payoff. At present, we do not believe that such dramatic cost reduction (or performance improvement) benefits exist in the RF power generation area.

APPENDIX A

Transmitter Options for the Slave Terminal

A number of transmitter options were studied in detail for low duty factor operation (control data) and for CW operation (video data). The options of greatest practical value at this time are listed in Figures A-1 and A-2. Other classes of transmitters were considered but are not included, either because they do not meet our needs or because their state of development is not adequate for a near-term solution. These other transmitters, which include negative resistance diode amplifiers and oscillators and electron beam amplifiers, may prove to be of value in the future, however.

The cost estimates shown in Figures A-1 and A-2 include all associated circuits and power supplies. Total production quantities of ten thousand units and production batches of thousands are assumed in these estimates. In most cases, costs are based on verbal estimates or extrapolations from data for similar components. Some of the cost estimates based on multiple sources varied considerably. In these cases, the lowest numbers were used. In some cases, a feel for the types of production processes (manual versus automated) plus design knowledge led to a calculated guess.

Figure A-1 shows the main candidates considered for the pulsed transmitter which can be used for the control data transmissions with approximately a 1% duty factor. We will consider the merits and weaknesses of each design.

a. The pulsed triode can be used as an amplifier or as a phase-locked oscillator. The latter, shown in Figure A-1(a), represents an efficient and very low-cost implementation.

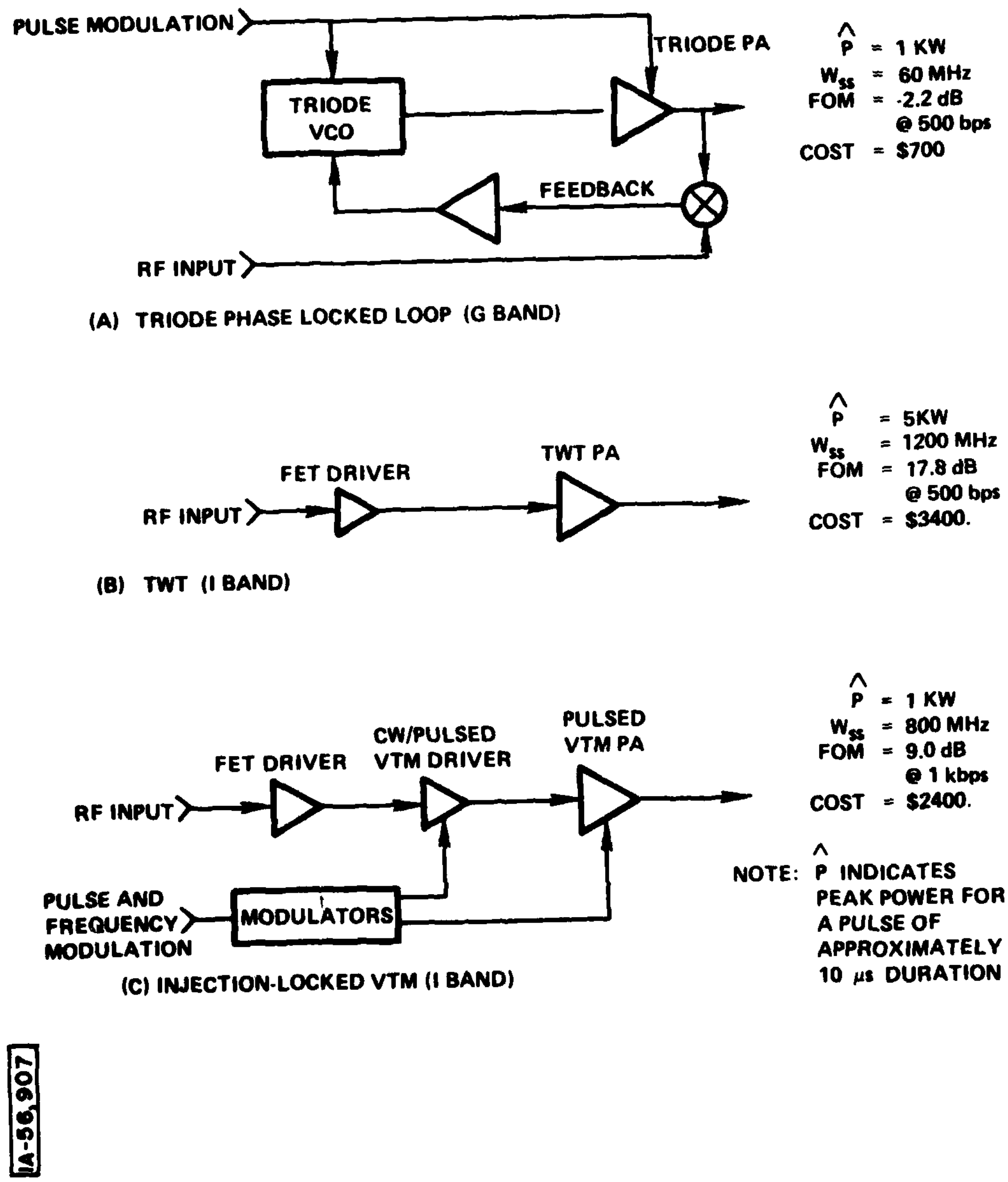


FIGURE A–1: CANDIDATE PULSE TRANSMITTERS

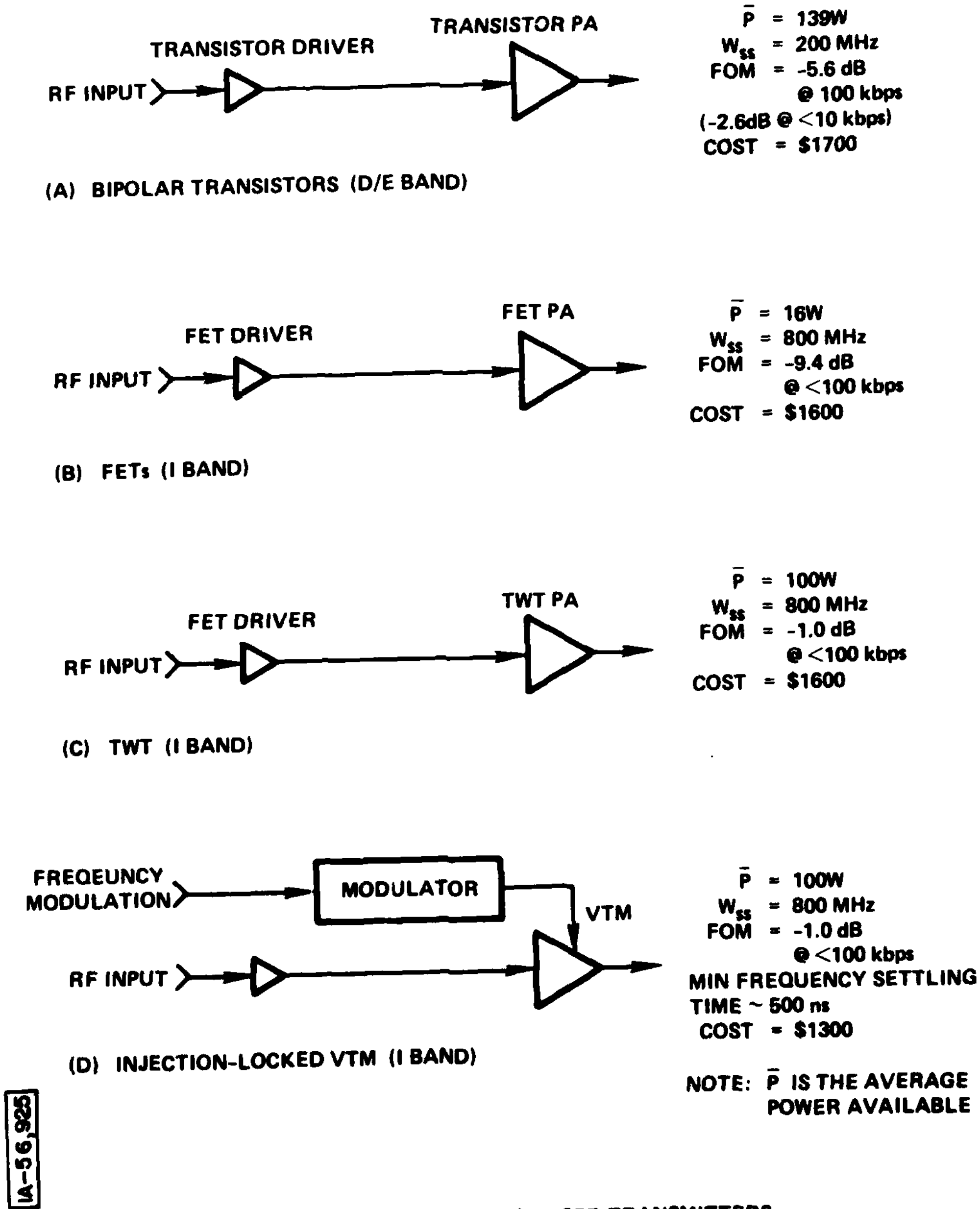


FIGURE A–2: CANDIDATE CW/PULSED TRANSMITTERS

Because of the low frequency of operation and the consequent bandwidth limitation, only a modest FOM (-2.2 dB) is possible. This transmitter might be applied economically in systems where the received jammer-to-signal ratio is sufficiently low.

b. The pulsed travelling wave tube (TWT) shown in design (b) offers the largest possible FOM (17.8 dB), but has a high cost with current production technology and is not amenable to substantial cost reduction with mass production.* The TWT offers simplicity and low drive power requirements, but it also requires high supply voltages and offers poor efficiency (typically 13%). To achieve this AJ capability, the cost of the data link terminal is further compounded by the higher costs of the T/R section components required by the high frequency operation (I Band). These additional costs are not included here.

c. The pulsed voltage-tuned magnetron (VTM) illustrated in design (c) appears to provide value in an FOM-versus-cost tradeoff. It also operates at a higher efficiency (typically 30%) than does the TWT. However, the FOM for the VTM is about 9 dB lower than that of the TWT. Adequate control of center frequency requires a high-voltage frequency control which somewhat complicates this approach. Since it cannot be used as an amplifier, the VTM operates as an injection-locked oscillator. Requiring a high-power drive (100 Watts), it is conveniently driven by another injection-locked VTM. The two tubes offer a natural means of multiple mode operation, which we will describe later.

* Research on low cost, low mean-power TWTs using printed circuit construction was carried out some years ago, with limited success. Further funding of similar research should be considered in the light of the data link system requirements.

Figure A-2 shows the main candidates considered for the CW (and low pulse power) transmitter. Both available transistor and tube technology are candidates here.

a. Bipolar transistors can be combined, as is common practice, to provide hundreds of watts of power with reasonable efficiency. Steady progress is being made toward providing greater power output per transmitter module containing several transistors. Improvements in the combining techniques result in greater ease of broadbanding the devices. The frequency limitations of such devices constrain the bandwidth to about 200 MHz, and the output power is limited by efficiency and the prime power available.

   A maximum output CW power of about 100 watts is considered practical. The typical transistor transmitter shown in design (a) has an FOM of -5.6 dB at 100 kbps. This does rise to -2.6 dB at 10 kbps, but it presents no advantage over the tube competitors.

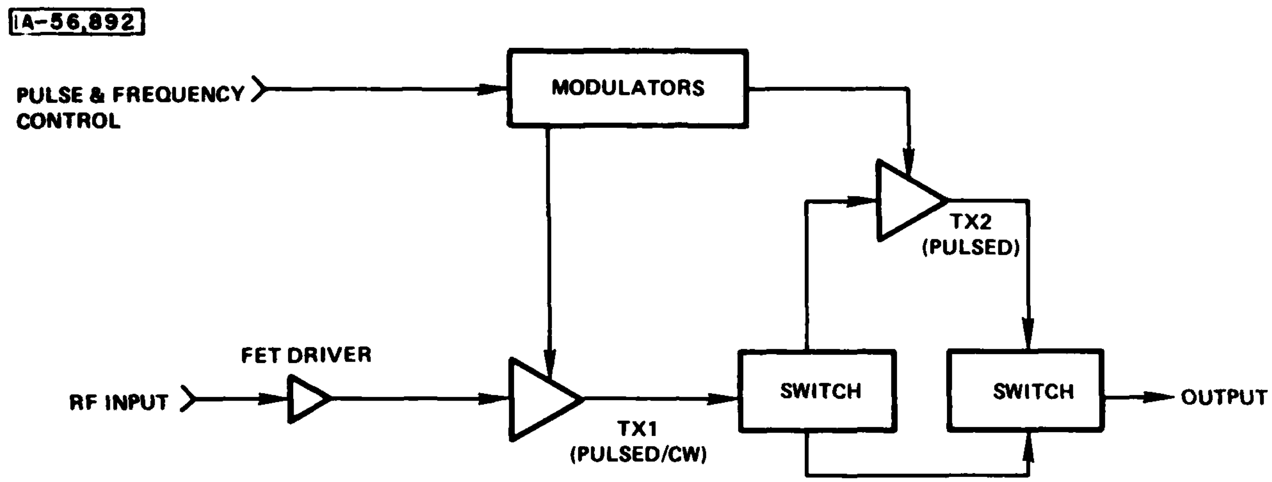
b. In view of the high cost and low output powers available at the high bandwidths and frequencies needed, FET transmitters, as shown in design (b), are presently not even competitive with bipolar transistors. Since FETs are undergoing a rapid development process, however, considerable cost reductions can be expected in the future, in addition to increased power output per module.

c. The pulsed/CW TWT, shown in design (c), has undergone extensive development for ECM applications. It offers both simplicity and an improved FOM of -1 dB at 100 kbps. It has a low efficiency (typically 15%), however, which increases its prime power demand.

d. Like the high-power pulsed VTM, the pulsed/CW VTM in design (d), requires a high-voltage frequency-tuning circuit. The cost of this circuit is difficult to estimate, since a fast-switching modulator must be specially developed. In the near term, frequency switching speed will be limited to about 500 ns, using current VTM designs. If a new design, tailored to our needs, can be developed in conjunction with the modulator, we consider the VTM the best future value in terms of the FOM per dollar. The VTM efficiency is reasonably good (typically 30%), which reduces prime power demands, or conversely allows a higher power tube to be used.

The best compromise approach for a transmitter which approximates the ideal curve $E \times W_{ss} \propto (\text{bit rate})^{-1}$ for a variable data rate with variable AJ protection, lies in the use of two transmitters as shown in Figure A-3. The low power pulsed/CW tube TX1 is used either directly as a transmitter or as a driver for the high pulse-power tube TX2. Three different options are listed. The proposed near-term option (2) uses a pulsed/CW TWT and a pulsed VTM.

The recommended long-term option (Figure A-3 #1) would use the specially developed pulsed/CW VTM described previously. This represents only slight degradation in performance over the near-term option. We expect to lower the frequency switching time to below 100 ns for fast frequency settling. The third option, (Figure A-3, #3), provides about a 9 dB improvement for low data rates at a considerable cost increase; it is probably not justified.



OPTIONS

TX1 (PULSED/CW)	TX2 (HIGH POWER PULSED)	FOM (dB) PULSE/CW	FOM (dB) HIGH POWER PULSE	COST $	NOTES
1. VTM 80W 800 MHz	VTM 900W 800 MHz	-2.0 @ 100 kbps	8.6 @ 1 kbps	2800	MIN. SETTLING TIME ≈500 NS. POSSIBLE LONG TERM SOLUTION
2. TWT 80W 1200 MHz	VTM 900W 800 MHz	-0.2 @ 100 kbps	8.6 @ 1 kbps	3200	PROPOSED FOR NEAR TERM
3. TWT 80W 1200 MHz	TWT 4500W 1200 MHz	-0.2 @ 100 kbps	17.3 @ 500 bps	<4900	MAX. AJ MARGIN

FIGURE A-3: MULTI-MODE TRANSMITTER

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