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# EVALUATION OF CSE* MODEM FOR AIRCRAFT TERMINAL MULTIPLE ACCESS TO COMMUNICATIONS/ NAVIGATIONAL SATELLITES

5-10255

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by

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*Certain aspects of this Company's coded signal extraction (CSE) modem development, for which a patent has been applied, have been included in this evaluation. Since the program did not include any substantial development of the technique but was limited to defining and demonstrating a specific application, no proprietary or patent rights are surrendered by the Company. This document contains proprietary information of Technical Communications Corporation, the further use or disclosure of which has not been authorized.

Contract No. NAS 5-10255

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## SUMMARY

The objective of this program was to analyze and construct experimental equipment for the application of "CSE" modulation to aircraft satellite communications. CSE modulation* is essentially linear swept FM, used as a subcarrier for multiple-access communications. Because of its time resolution capabilities, it has the potential to afford protection against multipath fading (such as is experienced by aircraft communications relayed by satellite) and, at the same time, to provide multiple-access to the satellite.

Because of the experimental nature of the program, the exact specifications of the satellite users were not defined, although the program was confined to binary communications. Consequently, the following two modes of operation were analyzed: (1) at 2400 bps (a typical vocoder rate); and (2) at 75 bps (a typical teletype rate). Both modes of operation used a 50 kHz bandwidth and a transmitter center frequency of about 150 MHz compatible with the ATS-1 VHF satellite.

Experimental equipment capable of 2400 bps transmission has been built and delivered to NASA, Goddard Space Flight Center, for further testing. The experimental modulator was designed to interface with a VHF transmitter at NASA's Mojave Ground Station, and the experimental demodulator was designed to interface with an aircraft receiver using an 18 MHz intermediate frequency (IF).

*CSE is a proprietary technique for which a patent is pending.

Analytical and laboratory results have confirmed the value of CSE modulation in dispersive channels and the feasibility of implementing it. Relatively insensitive to doppler frequency shifts and efficient in bandwidth utilization, it affords the same degree of multipath fading protection as more sophisticated techniques—for example, pseudonoise or frequency hopping—in a self-synchronizing mode of comparative simplicity.

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# EVALUATION OF CSE MODEM FOR AIRCRAFT TERMINAL MULTIPLE ACCESS TO COMMUNICATIONS/NAVIGATIONAL SATELLITES

## 1. INTRODUCTION

### 1.1 Objective of the Contract

The objective of the contract was to demonstrate the feasibility, construct engineering models, and evaluate the performance of a communication technique known as coded signal extraction (CSE). Equipment developed under this program will be used to conduct experimental voice-bandwidth data transmissions between aircraft and an ATS-VHF repeater, with the purpose of advancing the state of the art in communications/navigational satellite system design.

The specific program goals follow:

1. Evaluate and describe the expected performance of the CSE modem in enabling air traffic control over remote areas by satellite relay of communications and navigation aids.
2. Support such evaluation with analytical and experimental proof.
3. Show comparisons with the expected performance of competing approaches.

## 1.2 Program Outline

The major tasks of the program follow:

1. Define a CSE design configuration best suited for the VHF satellite-aircraft communications application.

2. Compare this technique with alternative techniques and develop adequate proof of the CSE's suitability for this application.

3. Build and deliver one engineering model modulator and demodulator for experimental purposes.

A design configuration of the CSE communication system that was considered to be optimum for experimental tests of the technique was selected. The receiver initially consisted of a passive filter realization of the compressive receiver for 2400 bps modulation and an active correlation receiver for a high TW product 75 bps portion. The selection of an active correlation receiver was based on an investigation into alternative methods of realizing passive compressive filters of various TW products. The investigation included an intensive literature search in an attempt to benefit from the work of others in this field. A summary and appraisal of alternative methods that can be used is presented in Section 4 of the first quarterly report for this contract.

In summary, most of the techniques for achieving dispersive delay lines would have required extensive development for this application. A technique developed by O'Meara (1959) appeared feasible for the 2400 bps mode. The higher TW product mode operating at 75 bps presented formidable delay line problems in maintaining both component tolerances and reasonable sizes. Thus, a correlation receiver rather than a passive receiver was selected for the 75 bps mode.

During the program the cost of implementing both the 75 and the 2400 bps modes appeared to be greater than the available funds. The modulator and demodulator were therefore modified and simplified to include only the 2400 bps mode. This decision was influenced by a greater concern for aircraft vocoded-voice capability than for aircraft teletype.

A major consideration in comparing the CSE modulation technique to alternative approaches was performance in a multipath environment. To combat the effects of multipath fading, it is necessary to use a high time-bandwidth product signal. The degree of protection is directly proportional to the effective transmitted bandwidth for a given data rate. Any spread-spectrum form of modulation can therefore resolve multipath signals, although some (notably pseudonoise with its $\frac{\sin x}{x}$ spectrum) require more rf bandwidth for the same result. The performance of swept FM is in many respects—such as immunity to thermal noise, impulse noise, and CW interference—comparable to that of pseudonoise. CSE modulation, like pseudonoise, can be employed in many different communication system configurations—for example, binary or m'ary signals, correlators or time invariant filters, and coherent or noncoherent detectors (although swept oscillators are not conducive to phase coherence over successive sweeps). An important difference between pseudonoise and CSE is found in their respective sensitivities to doppler shifts or frequency instabilities when time invariant filters are used. The output amplitude of the CSE dispersive line is relatively unaffected by frequency offsets, while the output of a pseudonoise multi-tapped delay line is seriously degraded by frequency mismatches.

CSE differs in some respects from pseudonoise in its instrumentation, particularly in the design of the time invariant filter. Unlike the tapped delay lines used as matched filters for pseudonoise, addresses or slopes of the dispersive delay line are not easily changed. Moreover, presently known implementations of dispersive delay lines make their use with high TW products inflexible. However, electron spin resonance techniques or breakthroughs in digital delay lines could in the future provide lightweight, compact, and flexible matched filters for the swept FM signals. Multiple-access applications of the chirp technique are compatible with controlled access, analogous to time-division multiplexing, and with random access. If, however, the random-access mode requires full use of the available bandwidth for every address, complicated sweep patterns are required, precluding the use of easily instrumented time-invariant filters. The use of swept FM with correlation receivers would nevertheless be applicable.

Subsequent sections of this report describe the theory of swept FM and its potential applications to communications, the problem of receiver synchronization, the configuration of the delivered modem, and the performance of the modem in a multipath environment. A more detailed description of the system, including the operator's instructions and circuit diagrams, is presented in the Instruction Manual.

## 2. THEORY OF OPERATION

### 2.1 Properties of Swept FM

The CSE modulation technique is based on swept FM (chirp). Chirp modulation was first developed in the 1950's at Bell Telephone Laboratories (Klauder et al., 1960) for radar applications. It has been used in radar and sonar because it provides high time resolution and unambiguous time intervals while maintaining a constant envelope transmitted signal. Moreover, passive (time invariant) FM chirp receivers are relatively insensitive to doppler shifts (Kramer, 1967). These properties, which make swept FM an attractive radar signal, also make it a candidate for communications applications (Holland-Moritz et al., 1964), particularly where multipath propagation and/or multiple users are primary considerations (Dayton et al., 1967).

The mathematical form of single chirp signal is:

$$s(t) = A(t) \cos 2\pi(f_o t + at^2) \qquad -T/2 < t < T/2 \qquad (1)$$

where

$A(t)$ = signal envelope

$f_o$ = center frequency

$a$ = frequency sweep rate

$T$ = sweep duration.

The signal power spectrum, as shown in Figure 1, is the same for both positive and negative sweeps. The chirp signal makes very efficient use of the spectrum if $1/T$ is at least an order of magnitude less than the swept bandwidth. As the frequency deviation becomes very much larger than $1/T$, most of the signal power is contained within the swept bandwidth. The spectrum becomes nearly rectangular when $A(t)$ is a rectangular pulse.

The optimum receiver or matched filter can be realized with a time invariant filter in the form of a dispersive delay line. Heuristically, the function of the time invariant filter is to delay each frequency component of the input signal by a different amount so that all components add in phase at some point in time. This produces a compressed pulse at an increased peak amplitude compared to the peak amplitude of the input signal. In a lossless dispersive delay line or compressive filter, the peak of the pulse is very nearly $\sqrt{TW}$ times the amplitude of the input. There is therefore a potential processing gain or improvement in the signal-to-noise ratio of approximately the time bandwidth product. The time invariant or compressive filter has an impulse response that is the time inverse of the swept FM signal. Its frequency response is therefore constant in amplitude and has a parabolic phase characteristic.

An alternative optimum receiver is a time variant filter or correlator. The correlator can be instrumented by using a frequency-swept local oscillator that removes the frequency modulation, followed by a band-pass filter matched to the resulting tone burst. The output of the correlator is shown by Eq. (2):

$$r(t) = \int_{-T/2}^{t} s(\tau) \cos 2\pi(f_1\tau + a\tau^2)d\tau \doteq \frac{1}{2} \int_{-T/2}^{t} A(\tau) \cos 2\pi(f_1 - f_2)\tau d\tau . \qquad (2)$$

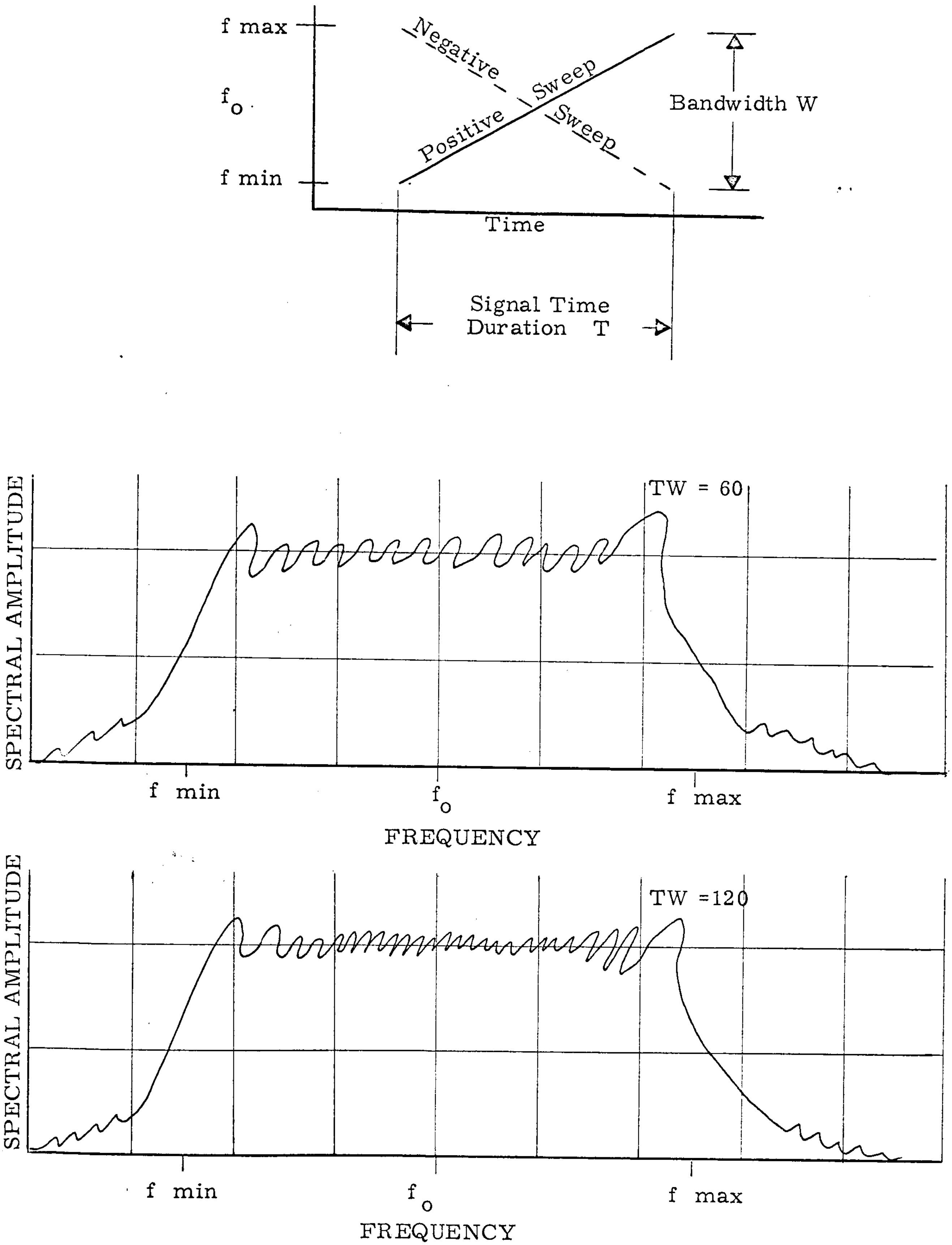



Figure 1. Typical Spectra of Swept FM Signals

In Eq. (3), it has been assumed that the sweep of the input signal and the sweep of the correlator are sychronized. It is important to note that a time shift of the signal results in a frequency shift at the correlator output. It can also be seen from Eq. (3) that the output of the correlator has a form similar to the Fourier transform of the signal envelope A(t). In fact, the output of the correlator as a function of the mismatch in frequency (or time) between the input and the local oscillator is the spectrum of A(t). If A(t) were a rectangular pulse, the correlator output as a function of the mismatch would have the form $\frac{\sin x}{x}$. A mismatch in frequency of $(f_1 - f_2) = \frac{1}{T}$ would correspond to the first null of the $\frac{\sin x}{x}$ function.

In radar, where there is uncertainty in the target velocity (doppler shift) and in its range (time delay), the effect of these mismatches is of prime importance. A useful tool in analyzing the effectiveness of a signal is its ambiguity function, which is the envelope of the matched filter output as a function of time and frequency mismatch (Woodward, 1953). A convenient pictorial representation of the signal ambiguity function is obtained by plotting surfaces of constant amplitude. The ambiguity function of the FM chirp waveform is shown in Figure 2, first in a three-dimensional plot and then in a more convenient two-dimensional plot.

An important property of swept FM is shown by the ambiguity diagram. A small shift in doppler $(\frac{1}{2T})$ with no timing mismatch will cause a great decrease in the filter output. Similarly, a small shift in timing $(\frac{1}{2\Delta f}$, where $\Delta f$ is the swept bandwidth) with no doppler shifts will also cause a considerable decrease in the filter output. However, the combination of a doppler shift and a timing shift can result in very little signal degradation. To a measurement system, this is an ambiguity; to a communications system, it is an opportunity to compensate for frequency changes by changing the receiver timing.

The ambiguity function is applicable to the time invariant matched filter as well as to the correlator. A time invariant or passive matched filter has an impulse response given by:

$$h(t) = \cos 2\pi (at^2 - f_o t) \qquad -\frac{T}{2} < t < \frac{T}{2} . \qquad (3)$$

The output of the matched filter for input s(t) is given by

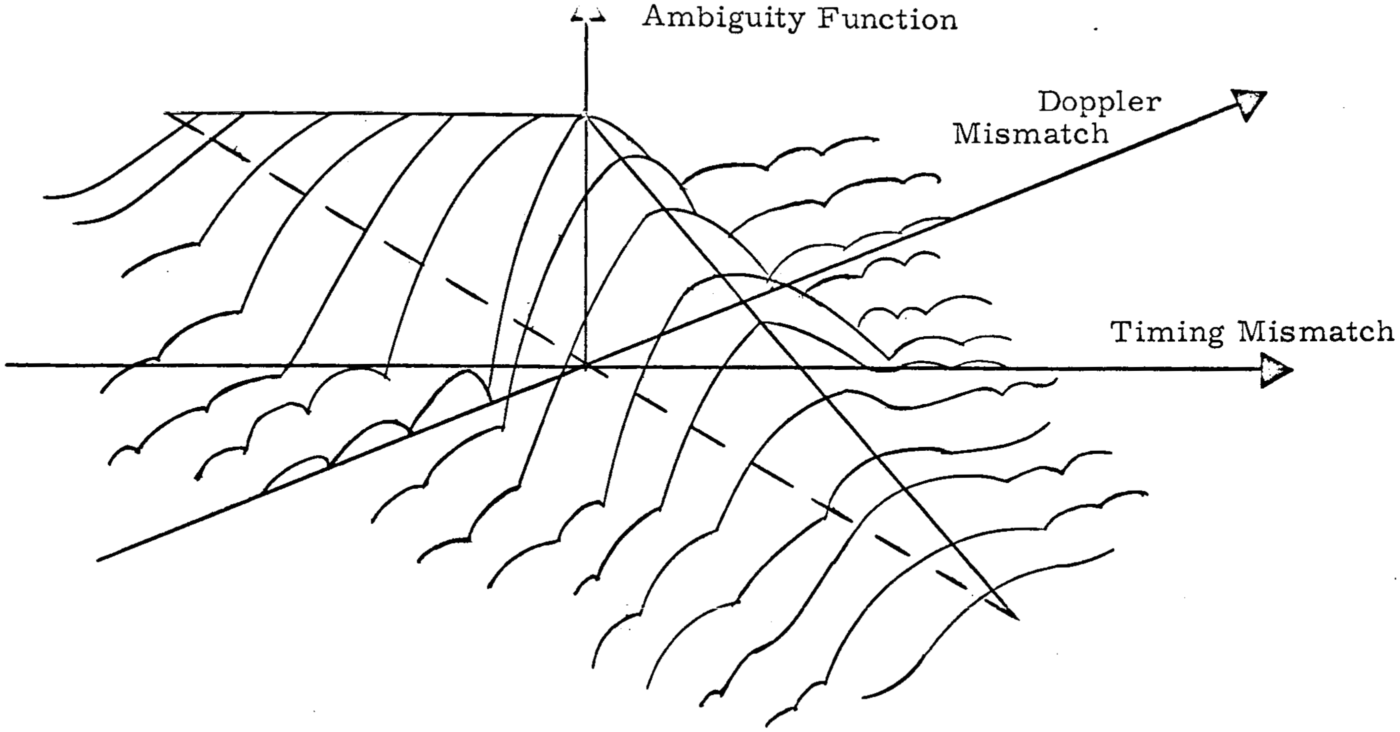



Figure 2a. Three-dimensional Representation

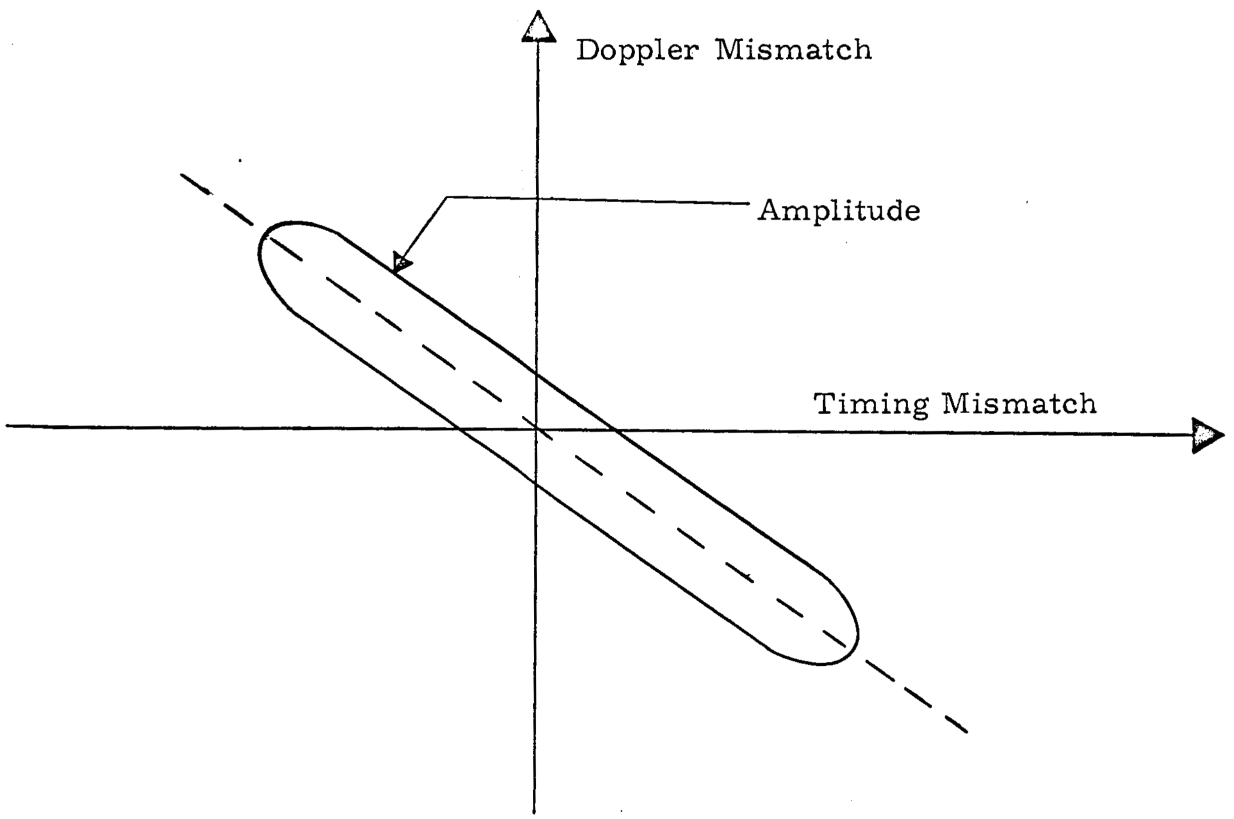



Figure 2b. Two-dimensional Representation

Figure 2. FM Chirp Ambiguity Function

$$r(t) = \int_{-\infty}^{\infty} h(\tau)\, s(t-\tau) d\tau \quad . \tag{4}$$

Carrying out the integration in Eq. (4) yields an envelope of

$$r(t) = \frac{\sin 2\pi\, a\, t^2}{2\pi\, a\, t^2} \qquad -T < t < T \quad . \tag{5}$$

The amplitude of the matched filter output as a function of time is a cross section (along the timing axis) of the ambiguity diagram. This function is shaped like the spectrum of A(t), the envelope of the transmitted signal. Thus, if A(t) were a rectangular pulse, the dispersive delay line output would be of the form $\frac{\sin x}{x}$ with its first null shifted $\frac{1}{\Delta f}$ from the peak. The ambiguity diagram indicates that a doppler shift will result in a similar pulse except that its timing will be shifted.

The analogy between the output of the dispersive delay line and the spectrum of A(t), the transmitted signal's envelope, is valuable in determining the effects of several types of mismatch between the transmitted signal and the receiver. For example, if the transmitter distorted the signal so that the envelope had a sinusoidal ripple, the matched filter output could easily be calculated. The spectrum of A(t) would have energy at the ripple frequency, and also at its negative frequency (Cook, 1962). The output of the dispersive delay lines would therefore have "paired echoes," smaller pulses preceding and following the main pulse separated by an interval proportional to the ripple frequency.

Similarly, imperfect filtering in the receiver can impose phase or amplitude modulation on the received signal. Its effect on the matched filter output is the same as its modulation effect on the spectrum of A(t). For example, receiver filters that round off the edges of A(t), making it resemble a gaussian pulse, will reduce the side lobes of the dispersive delay line output.

The effect of mismatched slopes ($\Delta f/T$) is important to communications applications of swept FM. Since the transmission of different slopes could represent information symbols and/or transmissions from different sources, a prime consideration is the distinguishability of these slopes— that is, how much of a mismatch is necessary between the slope of the signal and the slope of the filter before a significant reduction in the filter output occurs? This question can be answered by considering the mismatch as a parabolic phase or as linear frequency modulation on the signal envelope $A(t)$. The filter output can then be seen as analogous to the spectrum of $A(t)$ modulated by the mismatch. The output of the delay lines can be computed by using the Fresnel integrals used to calculate the transmitted signal spectrum.

## 2.2 Swept FM as a Communications Waveform

Two advantages of swept FM as a radar signal that make it attractive as a communications signal are its time resolution capability, which allows discrimination against multipath interference, and its insensitivity to doppler shifts. There are also two important differences between the radar and communications application of swept FM—first, in communications the transmitted signal must be modulated to convey information and, second, in communications the signal usually consists of a very long train as opposed to a single burst of energy.

The swept FM signal can be used to represent digital information in a number of ways. On-off keying or the use of two different slopes can be used to represent binary digits, while higher order alphabets can be implemented by using several different slopes or combinations of slopes. Moreover, the frequency sweeps need not be linear. The matched filter to the linear sweep is easily specified, but it is not necessarily easier to realize than a filter for a nonlinear sweep. The important requirement is to hold both generators and matched filters within close tolerances of the planned characteristic. Information can also be conveyed by the relative timing of successive signals. However, the multipath environment introduces serious problems in the demodulation of pulse position modulation.

Transmitting a train of swept FM signals as opposed to a single burst has two effects. The first, which is an advantage, is that synchronization is possible. By tracking the first few pulses of a long train, the timing of subsequent pulses can be derived. This is particularly important if correlation receivers are used as matched filters. Synchronization is discussed in greater detail in Section 2.3.

The second effect of transmitting a train of swept FM signals is the interaction between successive pulses. This interaction is noticeable if dispersive delay lines are used to distinguish positive slopes from negative slopes. The dispersive delay line matched to the transmitted pulse will produce a compressed pulse, but the delay line matched to the wrong slope will disperse the received signal. Consequently, its output will be spread nominally over two sweep durations and will thus introduce interpulse interference. This effect is noticeable when relatively low TW product signals are used since the level of this interference is inversely proportional to $\sqrt{TW}$.

A third basic difference between radar and communications application of swept FM is experienced when many users share the same communications channel. Interference from other subscribers' signals is often the limiting factor in multiple-access communications. In such applications, a common bandwidth is shared by a number of subscribers whose signals are characteristically partly interleaved and partly overlapped in time. Wherever overlaps occur, there is intersymbol interference proportional to the number of simultaneous transmissions and their data rates and inversely proportional to system bandwidth.

The FM chirp technique, like other spread-spectrum modulation forms, requires a bandwidth much greater than the information rate to realize its interference-resisting capabilities. In order to provide a transmitter with adequate bandwidth, it is often necessary to share the available bandwidth among many users. If such sharing can be done efficiently, it is possible to gain substantial improvements in error rates (in an interference-limited channel) without commensurate penalties in rf spectrum requirements. The total system bandwidth required is then a function of not only the needs of a single user, but also of the number of users in the system, of the number of simultaneous transmissions, and of the way in which sharing of the bandwidth is organized and controlled.

There exists a wide range of possible approaches to system organization. At the extremes lie random access, which requires very little user coordination, and centrally controlled access, where users are closely coordinated by a central control, switchboard, or operator.* The chirp modulation technique is compatible with either random access or controlled access.

*A "self-organizing" approach, where signaling among subscribers performs the system control function, has also been suggested. For purposes of this report, such techniques can be considered special cases of controlled access.

The objective of centrally controlled access, as opposed to random access, is to define a few signals or channels that have very little mutual interference and to assign them to users as their needs arise. Frequency division or time division may be used to define such channels. Time division can be implemented using short pulses, synchronized pseudo-noise codes, or synchronized FM sweeps. With swept FM, all users might use the same swept signal, but could be distinguished by the timing of their sweeps. Central control with time division requires user coordination on two levels. First, all receivers and transmitters must be mutually synchronized so that a signal intended for one receiver will not enter another in the time slot it is guarding. Second, during operation, time slots must be assigned to users who need access to the system at that moment. Because of the bandwidth limitation of most systems, there are usually far fewer available time slots than there are total subscribers; however, there must be at least as many time slots as there are simultaneous transmissions. By close coordination, the channel can be nearly filled without generating prohibitive mutual interference. The system bandwidth of time division is thus determined, primarily by the number of simultaneous users, regardless of the total number of system subscribers.

Following is an example of a centrally controlled time-division system using swept FM. Its major advantage is bandwidth efficiency and its major problem is synchronization. Superimposed on the teletype channels of the users is a high speed order wire used to control the making and the breaking of channels. A 50 kHz bandwidth has been assumed.

Figure 3 shows a typical time-frequency relationship among four simultaneous transmissions, three of which are operating at 75 bps and one order wire at 2400 bps. The user of channel 1 is required to start his frequency sweep within 208 $\mu$s of t=0.. The user of channel N would have to begin his sweep within 208 $\mu$s of t=(N-1) 208 $\mu$s. Thus, there are 64 available time slots in each 13.3 ms interval. Each of the teletype transmissions would sweep once in the 13.3 ms shown before repeating. The 2400 bps channel would sweep 32 times in the interval shown. The user's signal is not synchronized and could therefore begin at any point on the time axis. Each of the teletype channels would be orthogonal to all others. Since the higher speed channel is not synchronized, it will generate noise-like interference to each of the teletype channels.

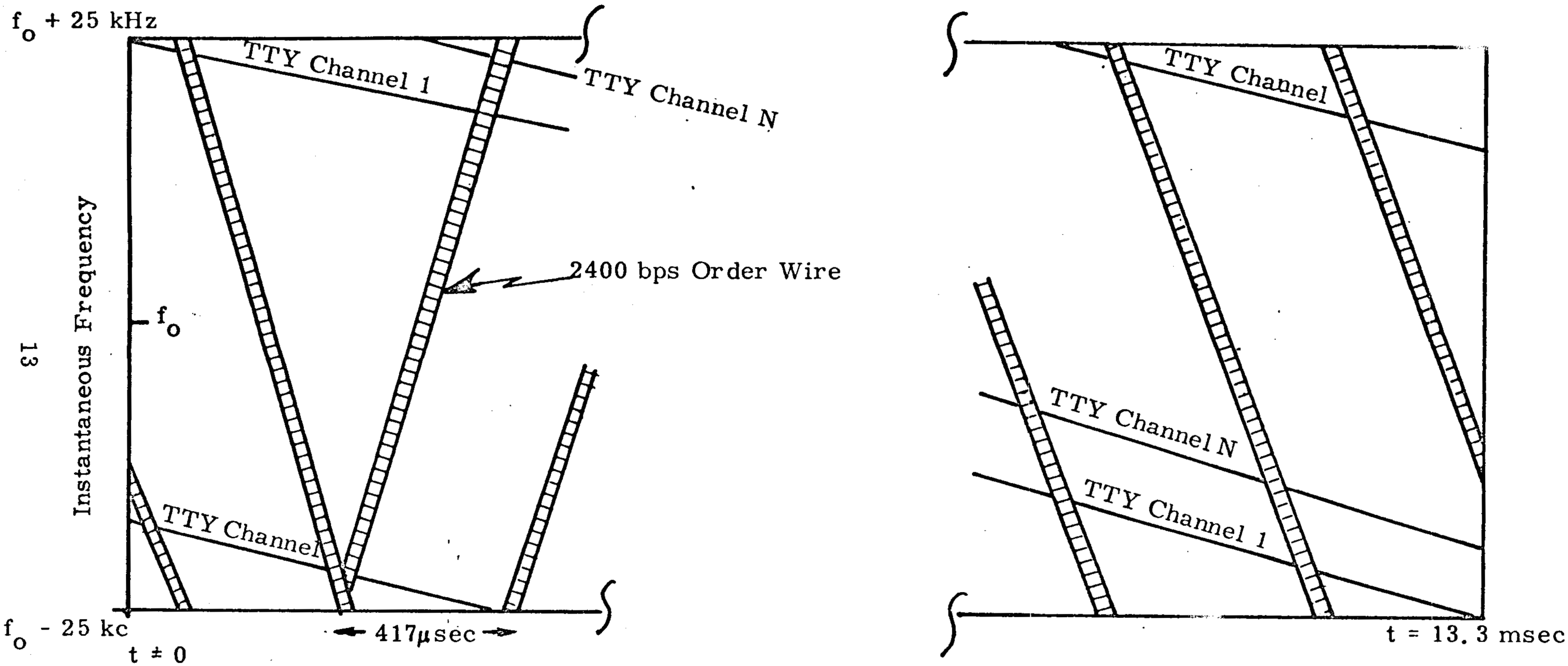



Figure 3. Typical Time-Frequency Relationships of Swept FM Signals in a Time-Division Multiple-Access System

The relative simplicity of the receiver and the relative immunity to intermodulation effects caused by nonlinear system elements are the primary advantages of a short pulse over the constant envelope modulations. However, in peak power limited cases or where jamming or pulse interference is a primary consideration, the constant envelope signals of synchronized pseudonoise or swept FM are superior. (The differences between synchronized pseudonoise and swept FM are primarily in their instrumentation.)

Random-access techniques are generally used when mutual interference is a secondary problem or when coordination among users is impossible. With random access, unrestricted and immediate access to the system is allowed all subscribers at all times with no controls. Each subscriber is assigned an address or channel. Unlike controlled access, usually so many addresses are required that it is impossible to assign an orthogonal signal to each address. Random-access systems are thus characterized by "graceful degradation"; as more users transmit signals, more interference is experienced by each. Channels or addresses can be distinguished by assigning each user a unique time-frequency matrix, pseudo-random code, or combinations of chirp slopes; usually, this signal or combination of signals would be transmitted with each bit. Binary information might then be sent as one or two addresses.

Ideally, one user will create no more interference to another user than an equal amount of noise power—that is, the addresses should be mutually uncorrelated. The signal vectors should be uniformly spaced to maximize the number of signals or addresses while minimizing the correlation among them. For chirp signals as with time-frequency matrices, a superimposed time vs frequency plot of all addresses should be uniformly filled; this is a fairly reasonable task with time-frequency matrices. Similarly, it is not difficult to conceive of a family of sweep patterns that fill the TW envelope uniformly.

The number of addresses that the random-access technique with swept FM can afford is a function of many system variables—such as the system bandwidth, the number of subscribers, and their duty factors, data rates and message lengths, transmitter powers, geographical dispersion of subscribers, and the allowable level of interference. The swept FM addressing techniques perform, equivalently in this regard, to pulsed time-frequency matrix techniques that have been studied exhaustively.* The most significant

*See ref. 9, 10, and 11.

difference is the detection procedure. The time-frequency matrix approach is based upon binary detection for each element of the matrix, whereas the swept FM receiver sweeps through the matrix and makes one detection per sweep; however, when many sweeps are used, this difference is negligible and the analysis of simultaneous users with time-frequency matrices can be applied to swept FM.

If it is required that all addresses sweep over the entire available bandwidth, time-frequency patterns such as those shown in Figure 4 are suggested. These patterns uniformly fill the time-frequency space and allow each user the entire bandwidth as protection against multipath. However, the design and instrumentation of time-invariant filters for such signals is complex and expensive. Consequently, in a multiple-access environment where complex sweep patterns are required, correlation receivers are preferable to the time-invariant matched filters. With correlation receivers, each receiver-transmitter pair must be synchronized, although system-wide synchronization (synchronizing all transmitters together) is not required.

## 2.3 Synchronization of Swept FM Demodulators

One of the most critical aspects of any communication modulation scheme is the problem of synchronization. Wideband techniques such as swept FM, by offering improved system performance via time resolution properties, require precise time alignment between the transmitter and the receiver. Procedures for effecting this adjustment range from the relatively simple problem of a single receiver with a time-invariant filter to the more complex problem of correlation receivers in a multiple-access environment.

The time-invariant matched filter or dispersive delay line is relatively easy to synchronize because a compressed pulse is produced for any swept FM signal of proper sweep rate and center frequency, regardless of its time of arrival. Synchronization is required only to sample the compressed pulse at or near its peak. Once initial timing has been acquired, synchronization can be maintained by automatically tracking the compressed pulses using track techniques such as the split-gate tracking loop.

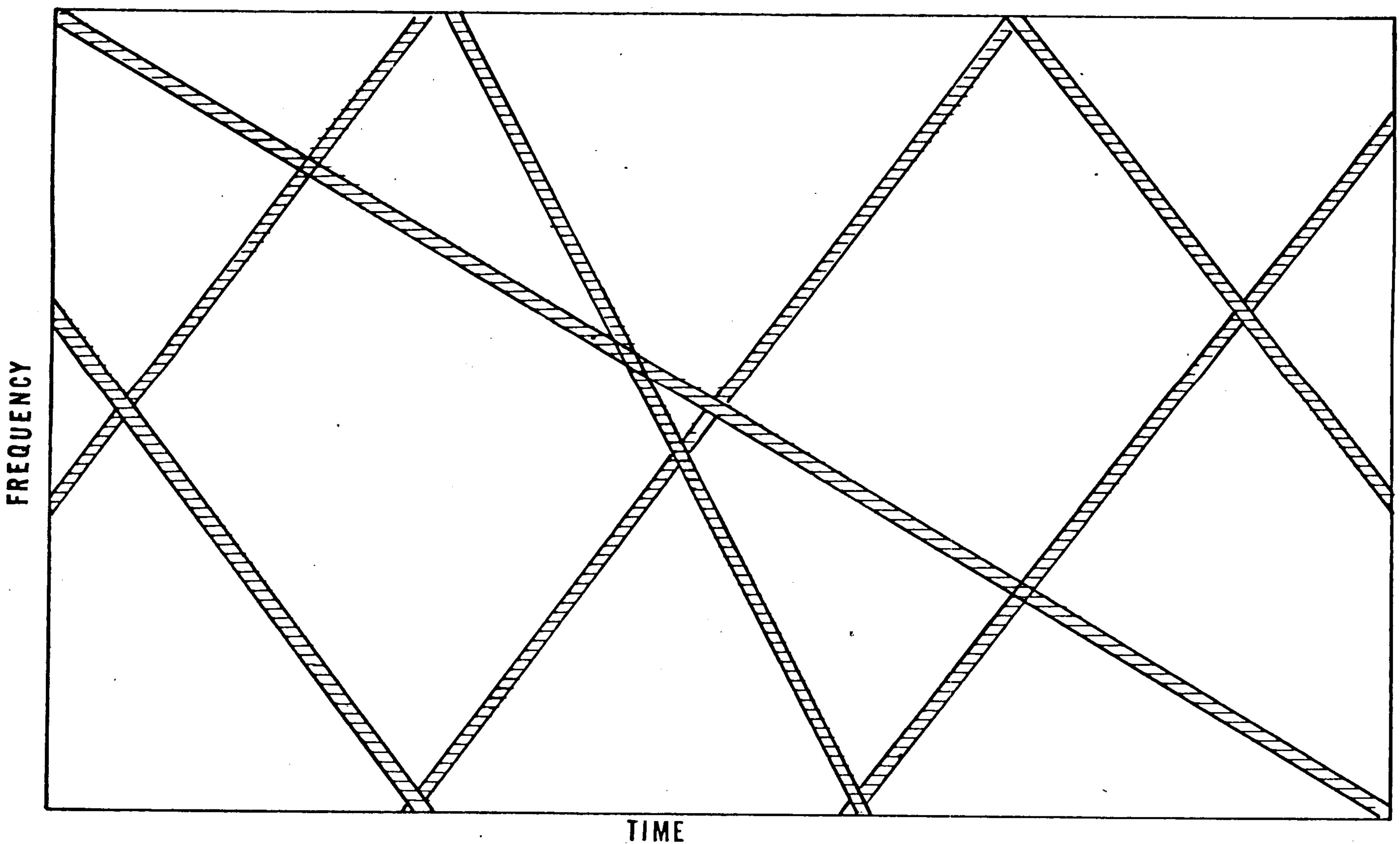



Figure 4. Time Frequency Diagram of Random-access Waveforms using Constant Bandwidth

Figure 5 is a block diagram of a split-gate loop that was instrumented at TCC. The input signal s(t) is an IF pulse with an envelope approximated by a $\frac{\sin x}{x}$ function. This signal is rectified, as shown in Figure 6, and applied to a gate that is the entrance to the timing loop. In order to lock on to the center of the pulse, an error function is needed to indicate how closely the loop is tracking the center of the pulse. Preferably, this error function will have a value of zero when the loop timing coincides with the input, will be positive when the loop timing should be advanced, and will be negative when the loop timing should be retarded. Such an error function is produced by the gating operation followed by low pass filtering or integrating. This type of loop, which is often used by radar receivers, is sometimes referred to as an early-gate late-gate (EGLG).

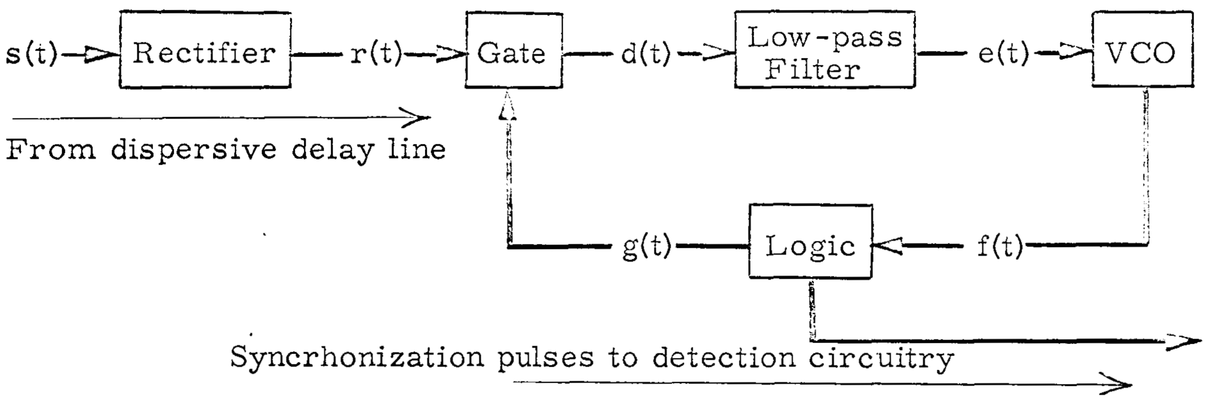
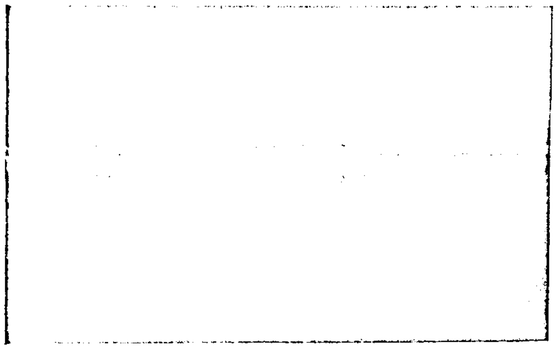



Figure 5. Split-Gate Loop

Figure 7 illustrates the typical waveforms of the loop. The gating function g(t) is shown to consist of a positive pulse (early-gate) followed by a negative pulse (late-gate). This function effectively multiplies the incoming rectified pulse, r(t), resulting in d(t). The form of d(t) will vary according to the relative position of the center of the pulse and the center of the gating waveform. Whenever these centers coincide ($\tau=0$) the positive and negative samples of r(t) are the same as long as r(t) is symmetrical about its peak. If the gate is early, the positive sample is smaller than the negative sample.

The output of the low-pass filter or integrator is a function of the relative sizes of the positive and negative samples. When the positive sample is smaller than the negative sample, the filter output becomes



Superimposed input and output of dispersive delay line

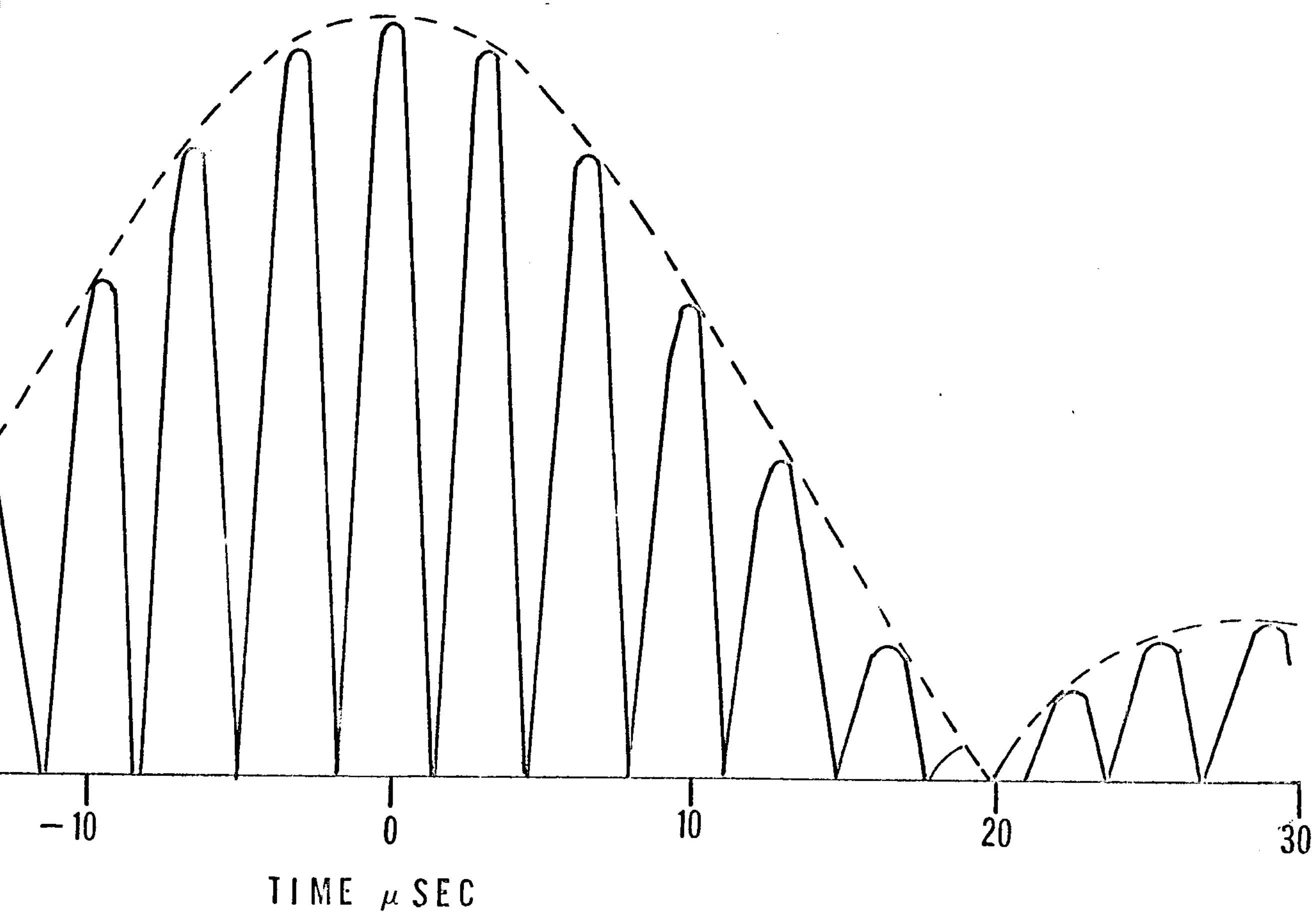



Figure 6. Rectified Output of Dispersive Line

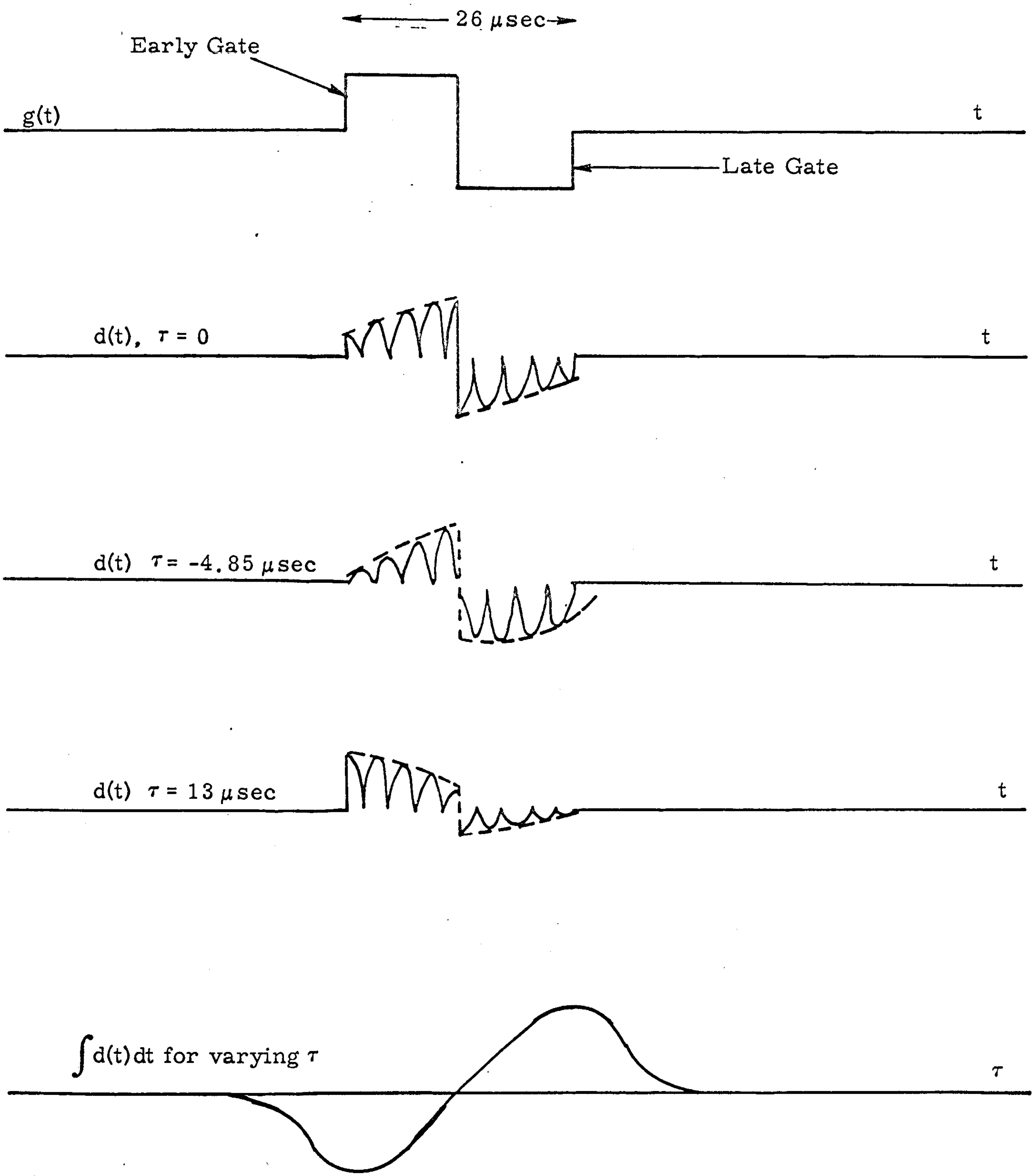



Figure 7. Typical Waveforms of Split Gate Loop

negative. This signal, which drives the VCO, decreases the sampling rate when it is negative; therefore, when the gate samples too soon, a negative signal applied to the VCO slows down the sampling rate until the gate coincides with the center of the incoming pulse r(t).

Conceptually, acquisition of timing for the correlation receiver is similar to that of the time-invariant matched filter. The analogy to the compressed $\frac{\sin x}{x}$ output of the dispersive delay line is obtained by varying the timing of the swept local oscillator. The magnitude of the matched filter output will of course be a maximum when the timing of the local oscillator sweep coincides with the input signal. As the timing between the swept signals is varied, the output pulse amplitude will vary according to

$$\frac{\sin 2\pi\Delta f\tau}{2\pi\Delta f\tau}$$

where $\Delta f$ is the sweep deviation and $\tau$ is the delay between the input and the local oscillator sweep. Like the dispersive delay line, the timing resolution and consequently the synchronization accuracy requirement of the correlation receiver is inversely proportional to the bandwidth.

One problem in acquiring synchronization is that of distinguishing between the direct and reflected signals, since both produce compressed pulses at the matched filter output. In most instances, these can be distinguished by selecting the earlier pulse. Even if multipath delays are comparable to the bit duration, appropriate synchronization signals can be chosen to distinguish the earlier arrival.

To automate the search for the peak of the matched filter output, similar techniques are appropriate for both correlation and time invariant receivers. The multipath signal can cause an ambiguous output in both cases. The time to acquire synchronization is considerably longer for the correlation receiver, however, since each measurement of the search procedure usually requires an entire bit duration.

The problem of maintaining synchronization is more difficult with the correlation receiver than it is with the dispersive delay line, which can track the compressed pulse output of the delay lines. One method that has been used is based upon measuring the phase shift through the integrate and dump matched filter. With linear swept FM, a timing error between the input signal and the swept local oscillator is reflected by a proportionate shift in frequency of the matched filter input (Figure 8). Since the matched filter has a linear phase characteristic as a function of frequency, the phase shift through the filter provides a measure of the timing error between the input signal and the local oscillator.

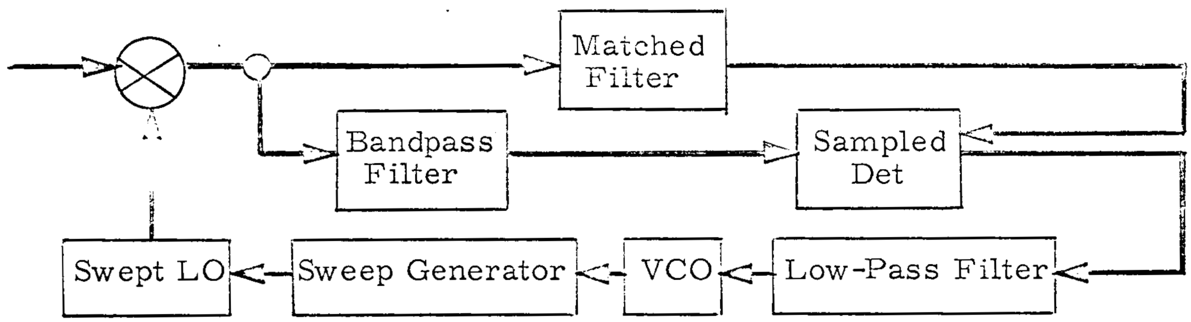



Figure 8. Synchronization Circuitry of Correlation Receiver

To compensate for this phase shift, a phase lock loop can be used. A relatively wide bandpass filter with linear phase characteristics in the band of interest is used to filter the unwanted harmonics caused by mixing and to reduce the noise bandwidth. The phase of this filter's output is then compared to the phase of the matched filter's output. The phase detector is sampled toward the end of each bit duration when the signal output of the matched filter is near its maximum. This phase measurement produces an error signal that drives a VCO and thereby controls the timing of the swept VCO. This loop behaves much like the split-gate tracking loop.

The swept FM receivers are relatively insensitive to doppler shift or any other source of frequency shift as long as it is small compared to swept bandwidth. Generally, the linear delay vs frequency characteristic of a dispersive delay line can be extended beyond the swept bandwidth so that frequency shifts merely change the timing of the output pulses. Similarly, the timing of a correlation receiver can be varied to compensate for frequency changes of the input signal. Consequently, small doppler shifts can be tracked by the synchronization loops previously described.

In an environment where doppler shifts or other frequency uncertainties are comparable to the swept bandwidth, the implementation of synchronization circuitry presents formidable problems. If frequency shifts carry the input of the dispersive delay line beyond its range of linear delay, the output pulse suffers serious distortion. Such frequency shifts must be detected and compensated for (perhaps by adjusting a local oscillator frequency) ahead of the delay line. The correlation receiver must conduct a two dimensional search in time and frequency to acquire synchronization. This process is conceptually easy to visualize as an ordered search for the maximum of the swept FM ambiguity function. Nevertheless, tracking problems are difficult because changes in frequency cannot be easily distinguished from changes in timing.

In a multiple-access environment, the difficulty of receiver synchronization depends greatly on the system organization. In a random-access system, where different slopes are used as addresses, there is not a high degree of correlation between signals intended for different receivers. Consequently, a receiver is not likely to lock on to the wrong signal when it is acquiring synchronization. In a system that assigns channels based on time slots, however, several synchronization problems are introduced.

Designating channels by assigning users different start times but the same frequency sweep requires system-wide synchronization—that is, each transmitter as well as each receiver must be synchronized to a system clock. In many instances such as where propagation times between users vary significantly or where a high degree of coordination among users is difficult, system-wide synchronization is impractical. There are certain satellite applications, however, where system-wide synchronization is feasible.

To achieve system-wide synchronization for satellite application, a system clock either located in the satellite or supplied by a ground terminal would be used to synchronize all the receivers in the system. The major synchronization burden would be placed on the transmitters, the signals of which must be synchronized with the system clock when they pass through the satellite. Consequently, in addition to tracking the system clock, each transmitter would be required to measure his range to the satellite to compensate for propagation time to the satellite.

If ephemeris data were inadequate to determine the range to the satellite, each transmitter would be required to transmit a ranging signal to measure the propagation delay (Griffiths et al., 1966).

## 3. SYSTEM DESCRIPTION

During this program, a modulator and a demodulator were constructed for test purposes. The modulator was designed to interface with a NASA transmitter located at their Mojave Ground Station. Test signals could then be transmitted to the ATS-1 satellite over the Pacific Ocean and relayed to a receiver. The demodulator was designed to interface with an aircraft receiver.

To test the modem's capability in various multipath environments, measurements would be made with the satellite at different elevation angles with respect to the aircraft. Laboratory testing of the modem is provided by a back-to-back interface frequency that allows the modulator output to be fed directly to the demodulator.

The modem was designed to operate at 2400 bps with a swept bandwidth of nominally 50 kHz, resulting in TW = 20. Information is conveyed by representing a binary "1" by a positive frequency sweep and a binary "0" by a negative frequency sweep. The matched filters used by the demodulators are dispersive delay lines (based on an O'Meara design). The outputs of the delay lines are envelope detected and compared after every sweep. Receiver synchronization is achieved by a radar technique called an "early-gate late-gate."

The modulator is packaged in two full ATR boxes designated T-1 and T-2; Figure 9 illustrates their functions. The binary message stream is converted to a stream of sweep signals (positive and negative sawtooth waves) in T-1. In T-2, the sweep signals control a VCO, the output of which is a linear FM signal. The output of the VCO is heterodyned from its 250 kHz center frequency to 74.61 MHz, which is the Mojave transmitter input frequency. In the transmitter the signal frequency is doubled to 149.22 MHz. The VCO output is also applied to a frequency doubler that provides the back-to-back interface through the demodulator 500 kHz input. (See Figures 10 thru 15 for photographs of the modulator and demodulator.)

The demodulator is packaged in two full ATR boxes and one half ATR box designated R-1, R-2, and R-3. Figure 16 illustrates the functions of R-1 and R-2; R-3 is the receiver dc power supply. In R-1 the 18 MHz input signal is heterodyned to a 500 kHz IF, which is also the back-to-back interface frequency. From 500 kHz the signal is heterodyned in 2 ways to 155 kHz, which is the center frequency of the dispersive delay lines. By mixing with the 655 kHz local oscillator, the resultant 155 kHz signal has its sweep reversed. By mixing with the 345 kHz local oscillator signal, the sweep of the received signal is unchanged. Thus, two identical dispersive delay lines can be used—one matched to the down-swept signal (without its sweep reversed) and the other matched to an up-swept signal (with its sweep reversed).

In R-2 there are two independent synchronization loops using "early-gate late-gate." Two loops are required because frequency shifts such as doppler effects will advance the timing of one channel while retarding the timing of the other. Sample pulses, which center on the compressed pulse output of the dispersive delay lines, are derived from these loops. These pulses are used to drive detection circuitry that reconstructs the message stream.

In the normal mode of operation, the modulator will accept a binary (±6 V) message stream from an external source, which can be driven from a 2400 bps ±6 V clock output of the modulator, and generate a swept FM signal at 74.61 MHz with a swept bandwidth of ±12.5 kHz. The receiver will accept a swept FM signal at a frequency of 18 MHz and a swept bandwidth of ±25 kHz and, after semiautomatic synchronization, will reconstruct the binary message stream.

Alternate modes of operation allow changing the output frequency by using an external local oscillator, using externally generated timing and/or

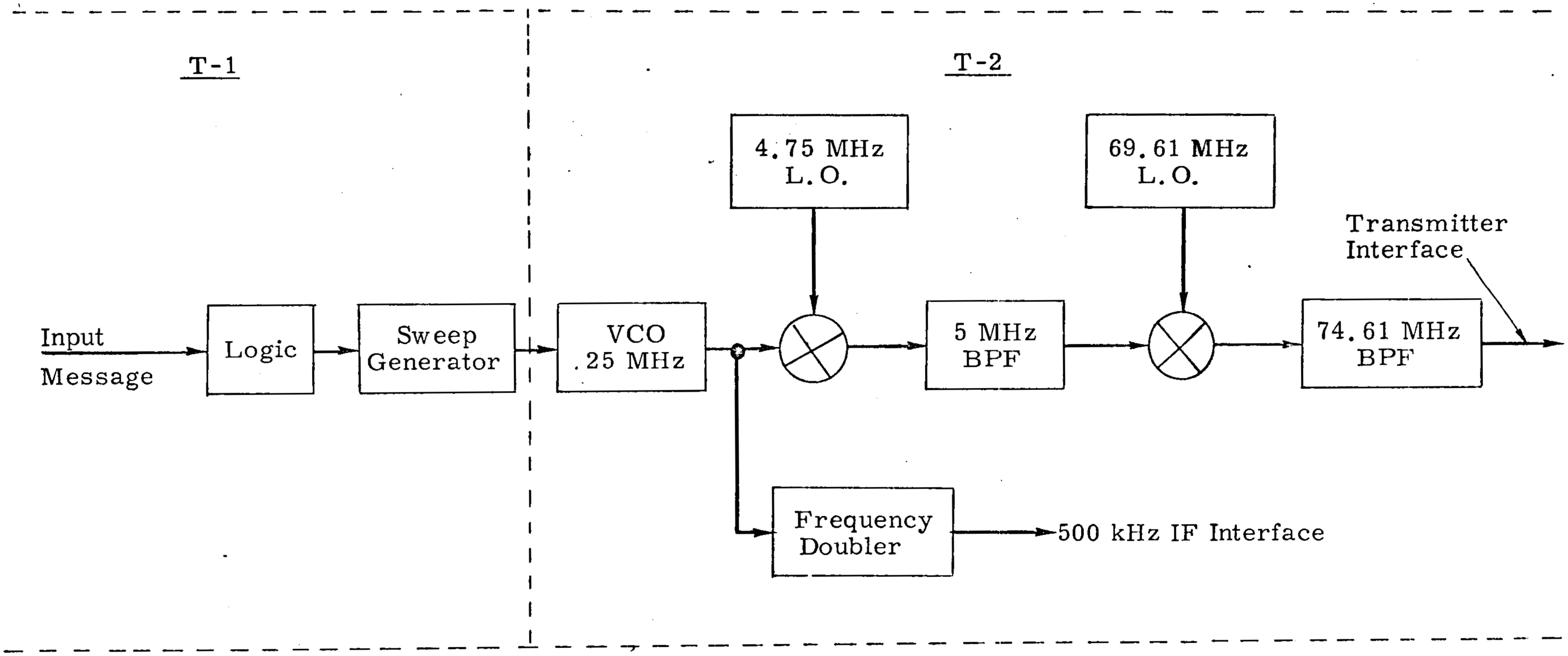



Figure 9. Simplified Block Diagram of Modulator

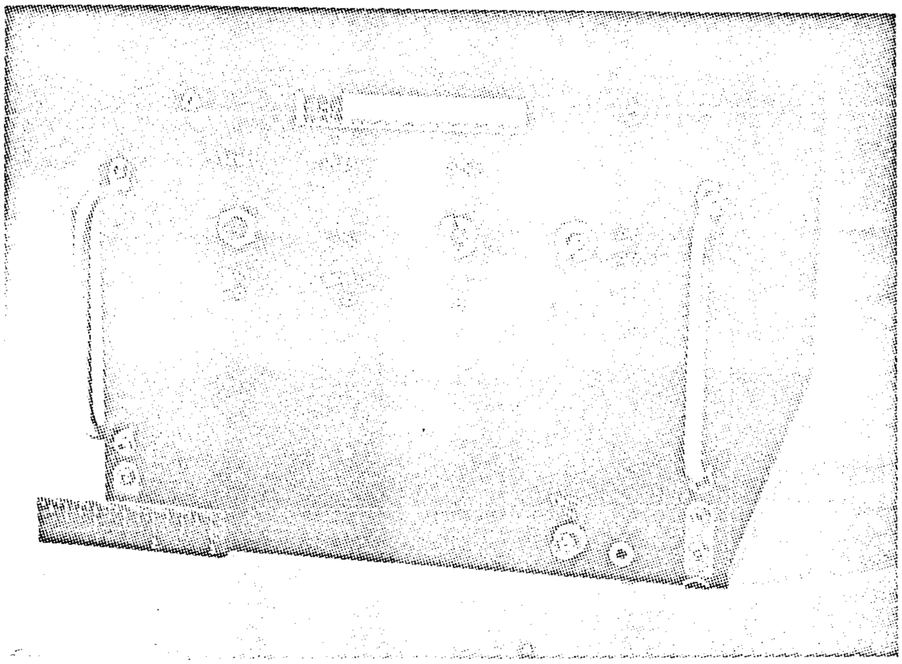


Figure 10. Photograph of Modulator Box T-1

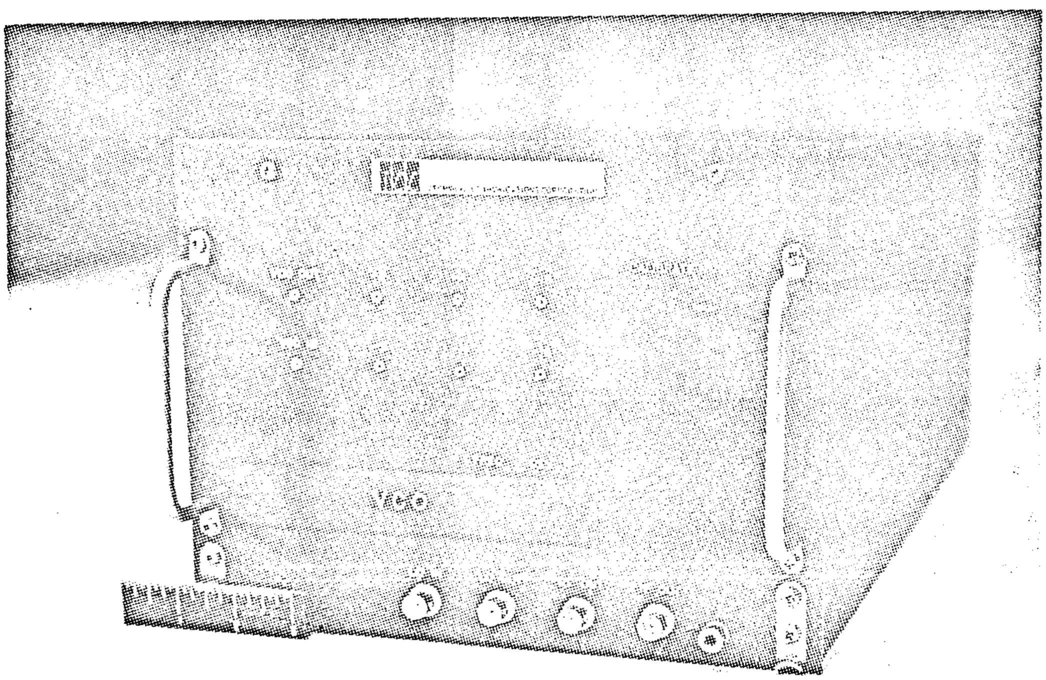


Figure 11. Photograph of Modulator Box T-2

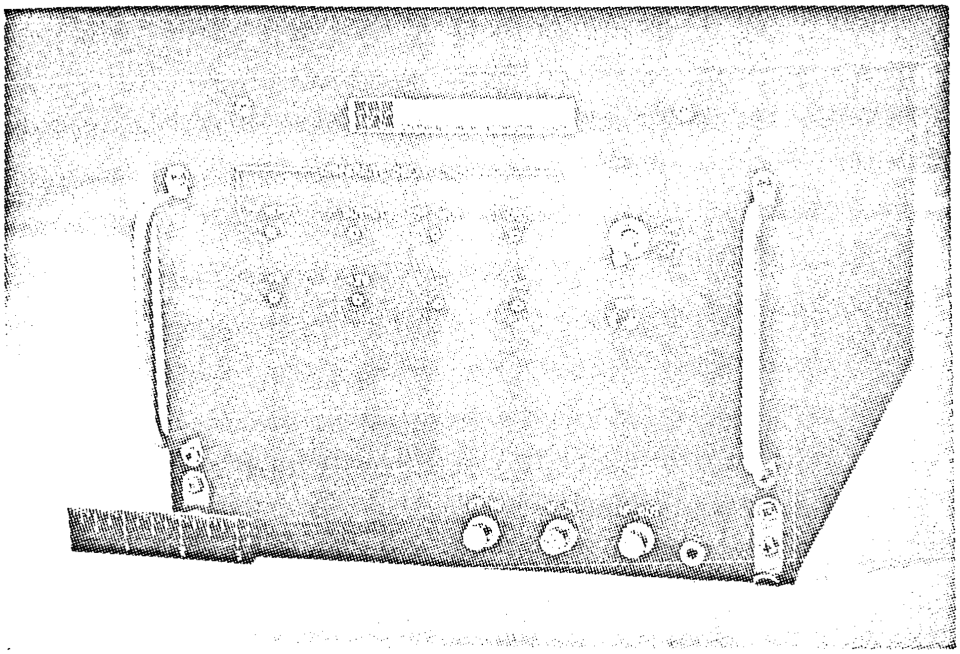


Figure 12. Photograph of Demodulator Box R-1

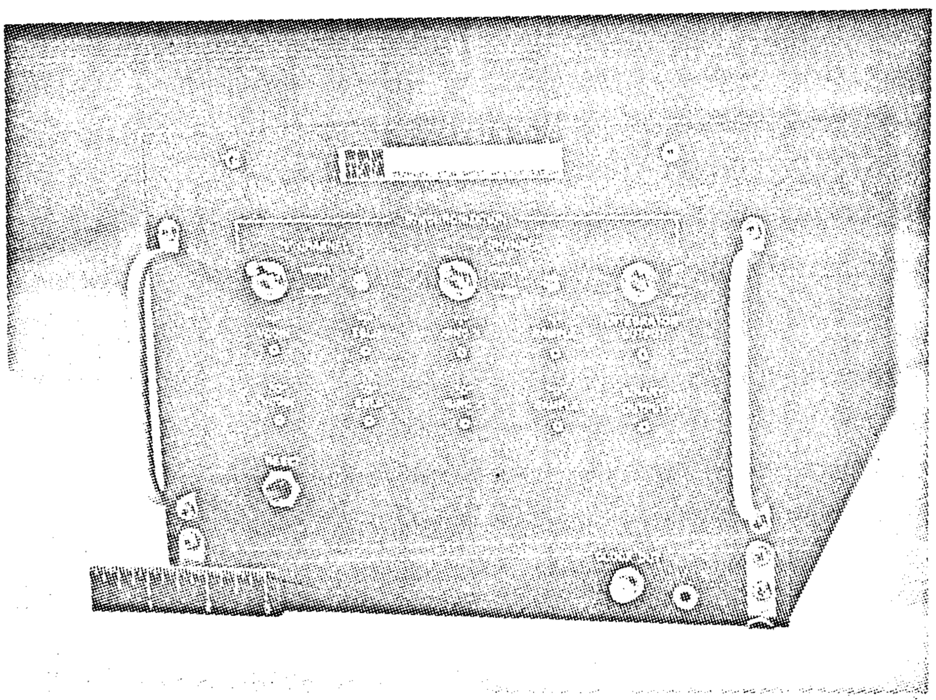


Figure 13. Photograph of Demodulator Box R-2

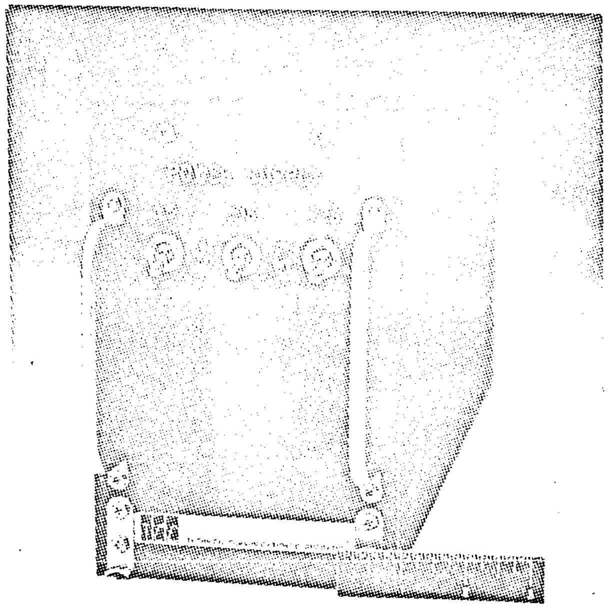


Figure 14. Photograph of Demodulator Power Supply

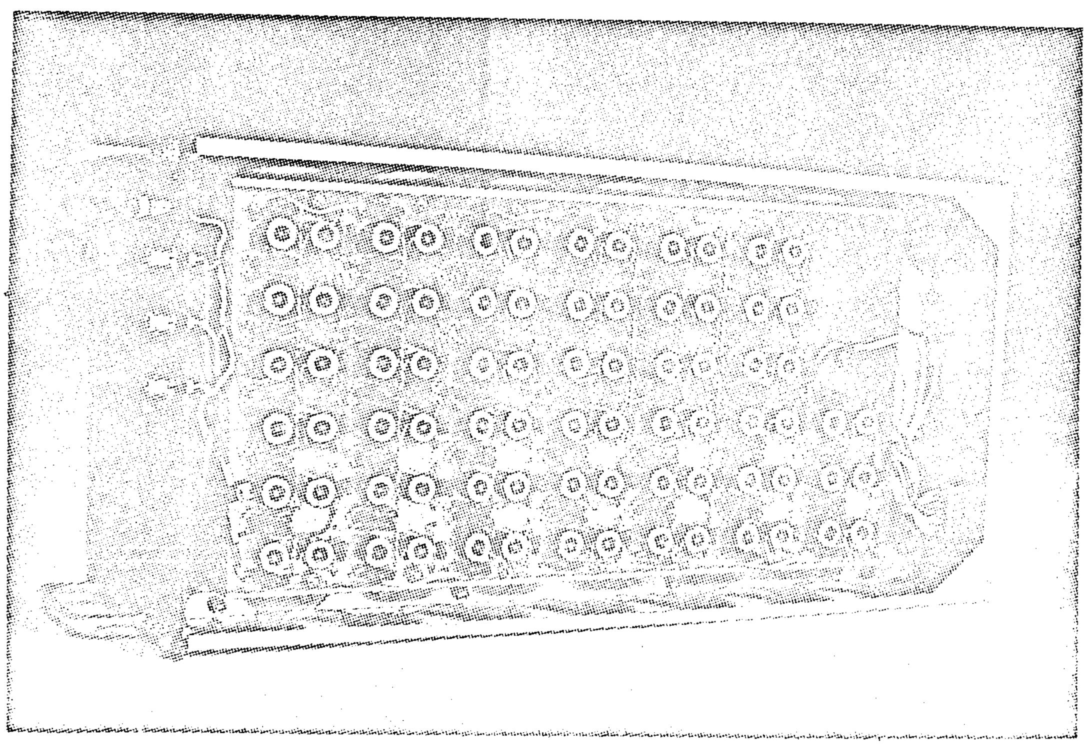


Figure 15. Dispersive Delay Line Mounted in R-2

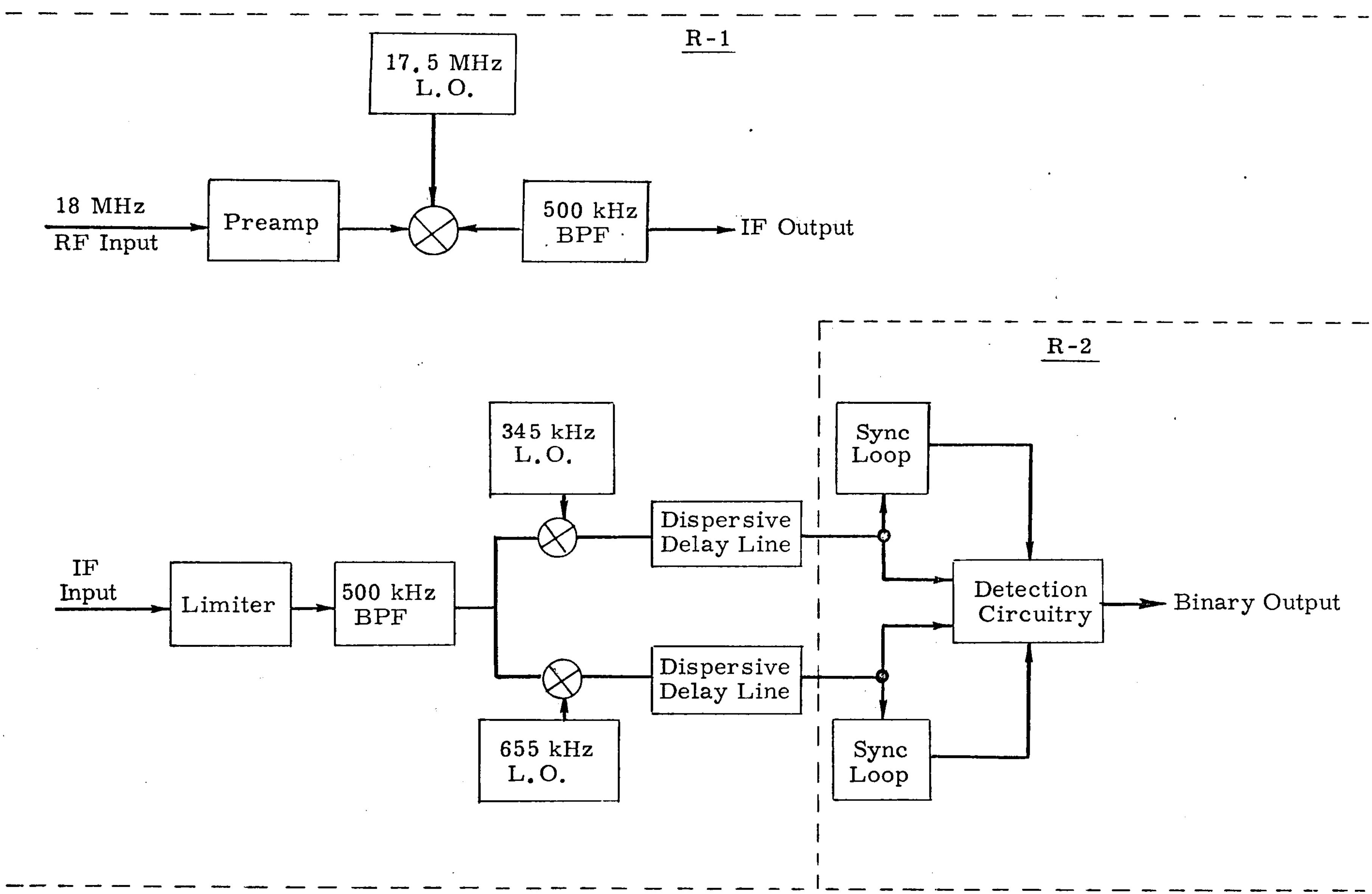



Figure 16. Simplified Block Diagram of Demodulator

internally generated message signals and back-to-back operation. In the back-to-back operation, the modem can disengage the synchronization circuitry and maintain timing by driving the modulator and demodulator with a common timing source. A detailed description of the modem, its operating procedures, packaging, and circuitry is presented in the Instruction Manual (TCC, 1967).

# 4. SYSTEM PERFORMANCE IN A MULTIPATH ENVIRONMENT

## 4.1 Multipath Model

A primary detriment to efficient use of an aircraft-satellite communications system is the multipath problem—that is, multiple arrivals of the same signal due to reflections. The following two dominant effects are caused by the multiple arrivals:

(1) intersymbol interference; and

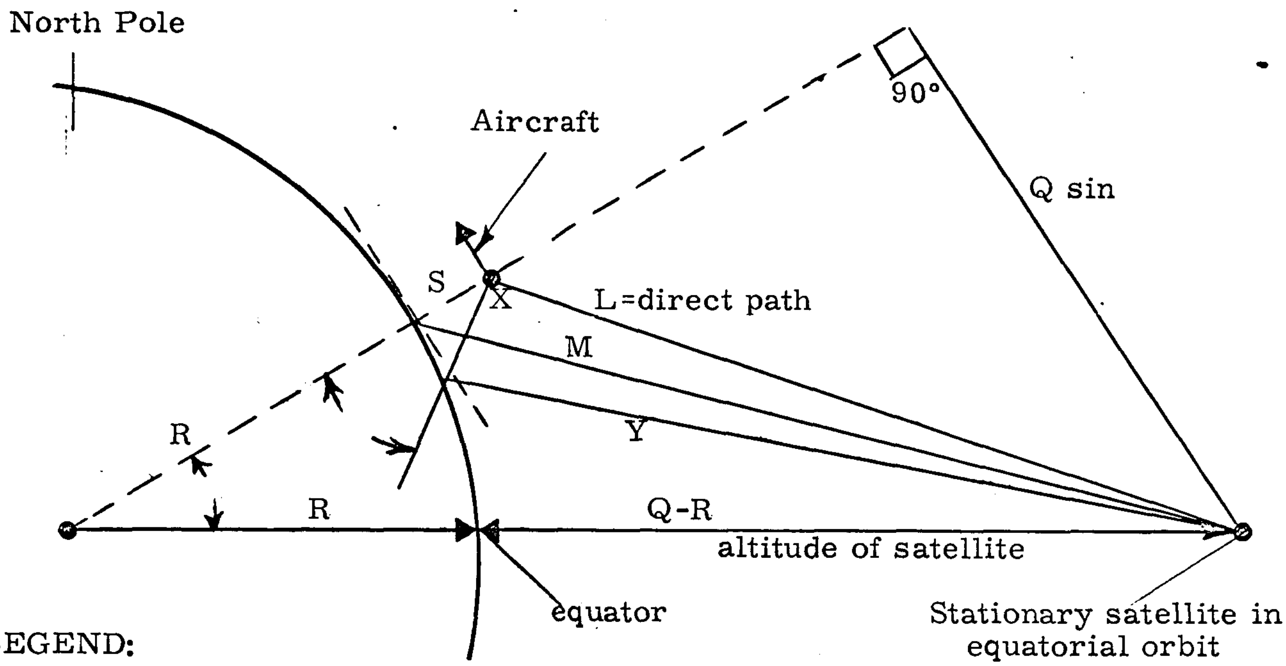
(2) coherent signal fading caused by equal-amplitude direct signals and secondary path signals cancelling each other.

This problem is particularly severe for aircraft operating at VHF because of the restrictions on antenna gains which might otherwise discriminate against the reflected signals. If the reflecting surface (such as sea water) does not greatly attenuate the multipath signal, it will be approximately as strong as the direct signal, differing only in its relative time of arrival. The time delay between the direct and reflected signals is a function of aircraft altitude and the relative positions of the satellite and the aircraft. This parameter is emphasized because of its effects on measures to combat multipath.(Second Quarterly Report of this contract.)

A geometrical model for aircraft-satellite multipath is shown in Figure 17. All pertinent distances are shown on this diagram along with a summary of important relationships.* A model of this type is known as a "two-ray" model—that is, there are only two possible paths:

(1) a direct path from the satellite to the aircraft; and

(2) a single specularly reflected path from the satellite to the earth to the aircraft.**




LEGEND:

| | |
|---|---|
| $Y \cong M - \frac{SQ}{M} \sin\theta \tan\phi$ | $D = X + Y - L = \frac{S}{\cos\phi} \left[1 + \frac{Q\cos(\theta+\phi) - R\cos\phi}{M}\right]$ |
| $L$ = direct path | $S$ = aircraft altitude |
| $Q \cong 6.5\,R$ for synchronous altitude | $X = \frac{S}{\cos\phi}$ |
| $M = \sqrt{R^2 + Q^2 - 2QR\cos\theta}$ | $D$ = path difference |
| $L \cong M - \frac{S}{M}[Q\cos\theta - R]$ | |
| $Y + X$ = approximate reflected path | |

Figure 17. Geometrical Model of Multipath

*This model, along with other technical data to follow, has been derived from AIAA Paper No. 66-294 entitled, "Fading and Multipath Considerations in Aircraft/satellite Communications Systems," by F. E. Bond and H. F. Meyer of Aerospace Corporation, El Segundo, California.

**Unpublished data based on measurements made at MIT's Lincoln Laboratories with their LES 3 and LES 4 experimental satellites indicate that multipath computations based on this model are extremely accurate except at very low aircraft altitudes.

Table 1 shows the time difference between a direct signal and its specular reflection from the earth as a function of aircraft altitude and latitude with respect to the satellite. It is assumed that the satellite is in an equatorial orbit and the aircraft is at the same longitude. If the aircraft is at a different longitude from that of the satellite, the effective latitude is increased.

Table 1. Multipath Time Dispersion as a function of Elevation and Altitude

| Altutide (ft) | Latitude (deg) | $\Delta t (\mu\ sec)$ | $\frac{1}{2\Delta t}$ (kHz) |
|---|---|---|---|
| 1,000 | 80 | 0.13 | 1430 |
| 1,000 | 60 | 0.7 | 710 |
| 1,000 | 40 | 1.4 | 360 |
| 1,000 | 20 | 2 | 250 |
| 5,000 | 80 | 0.7 | 710 |
| 5,000 | 60 | 4 | 125 |
| 5,000 | 40 | 7 | 71 |
| 5,000 | 20 | 9 | 56 |
| 10,000 | 80 | 1.3 | 143 |
| 10,000 | 60 | 7 | 71 |
| 10,000 | 40 | 14 | 36 |
| 10,000 | 20 | 18 | 28 |
| 30,000 | 80 | 4 | 125 |
| 30,000 | 60 | 22 | 23 |
| 30,000 | 40 | 42 | 12 |
| 30,000 | 20 | 55 | 9 |

In many typical aircraft applications, the information in Table 1 is not accurate for aircraft positions at very different longitudes from satellite longitude; for example, with a satellite positioned at 60°W longitude, New York is more than 45° away and London is 60° away. To present a clearer picture of expected multipath delays, an "effective" latitude has been computed which represents the angle subtended by lines from the satellite to the equator (at 60°W longitude) and from the satellite to a point on the earth at some latitude and longitude different from those of the satellite—that is, the position of the aircraft. Figure 18 is a plot of effective latitude as a function of true latitude and longitude with the satellite as a reference. Multipath delay can be found by entering Figure 18 at the aircraft latitude and longitude to obtain effective latitude. The total delay can then be obtained from Table 1 for the particular aircraft altitude.

When the reflected signal has a coherent (constant phase) component of approximately equal amplitude and opposite phase to the direct signal, the signal input to the aircraft receiver is cancelled.* This is normally a dynamic situation since the phase of the reflected signal cannot remain constant because of reflecting surface changes and/or satellite or aircraft motion. Only very small changes in multipath delay are required to shift the relative phase between direct and reflected signals such that phase addition rather than phase cancellation occurs. This point is illustrated in Figure 19, which shows a plot of relative received signal strength vs multipath delay for a carrier frequency of 150 MHz. Only 6.67 nsec of delay separates a peak from a null.

Figure 19 shows a plot of received signal strength vs carrier frequency for a multipath delay of 20 $\mu$sec. While one frequency is experiencing a multipath null, another frequency 25 kHz distant is experiencing a peak; therefore, a frequency diversity of 25 kHz is required to eliminate the possibility of a total loss in received power due to multipath fading.

## 4.2 Communications System Constraints

The previous discussion of multipath referred to the following two distinct effects:

(1) intersymbol interference; and

(2) fading.

The first of these effects is caused by the multipath of one symbol interfering with subsequent symbols; the second effect is when the multipath of a symbol interferes with the direct path of that same symbol. Modulation techniques

*A cancellation is highly probably when the reflection is off the ocean surface since a fairly calm surface tends to appear rather smooth to a signal at VHF; ground reflections are seldom specular because the terrain is almost always rough.

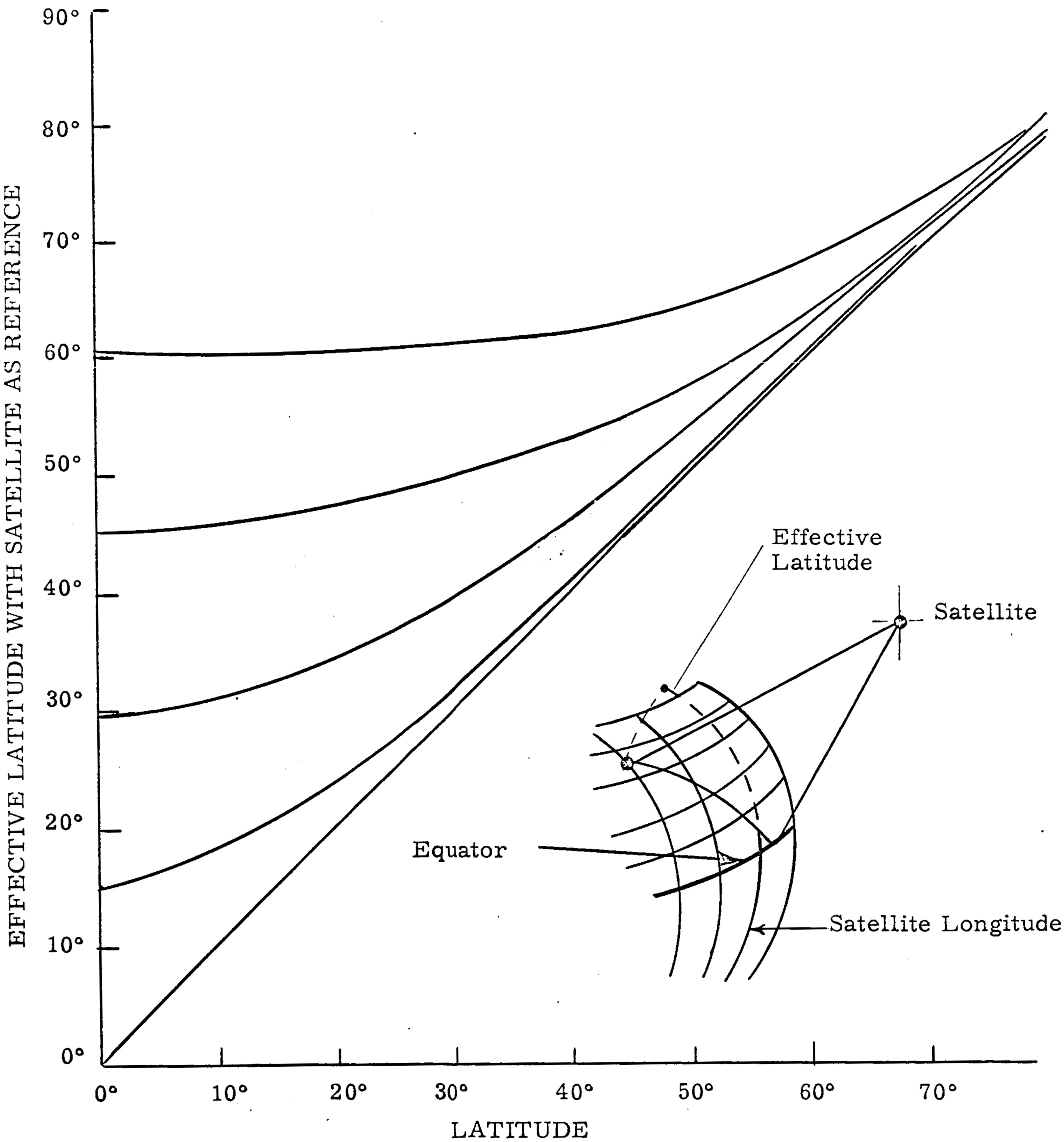



Figure 18. Effective Latitude as a Function of Aircraft Latitude and Longitude

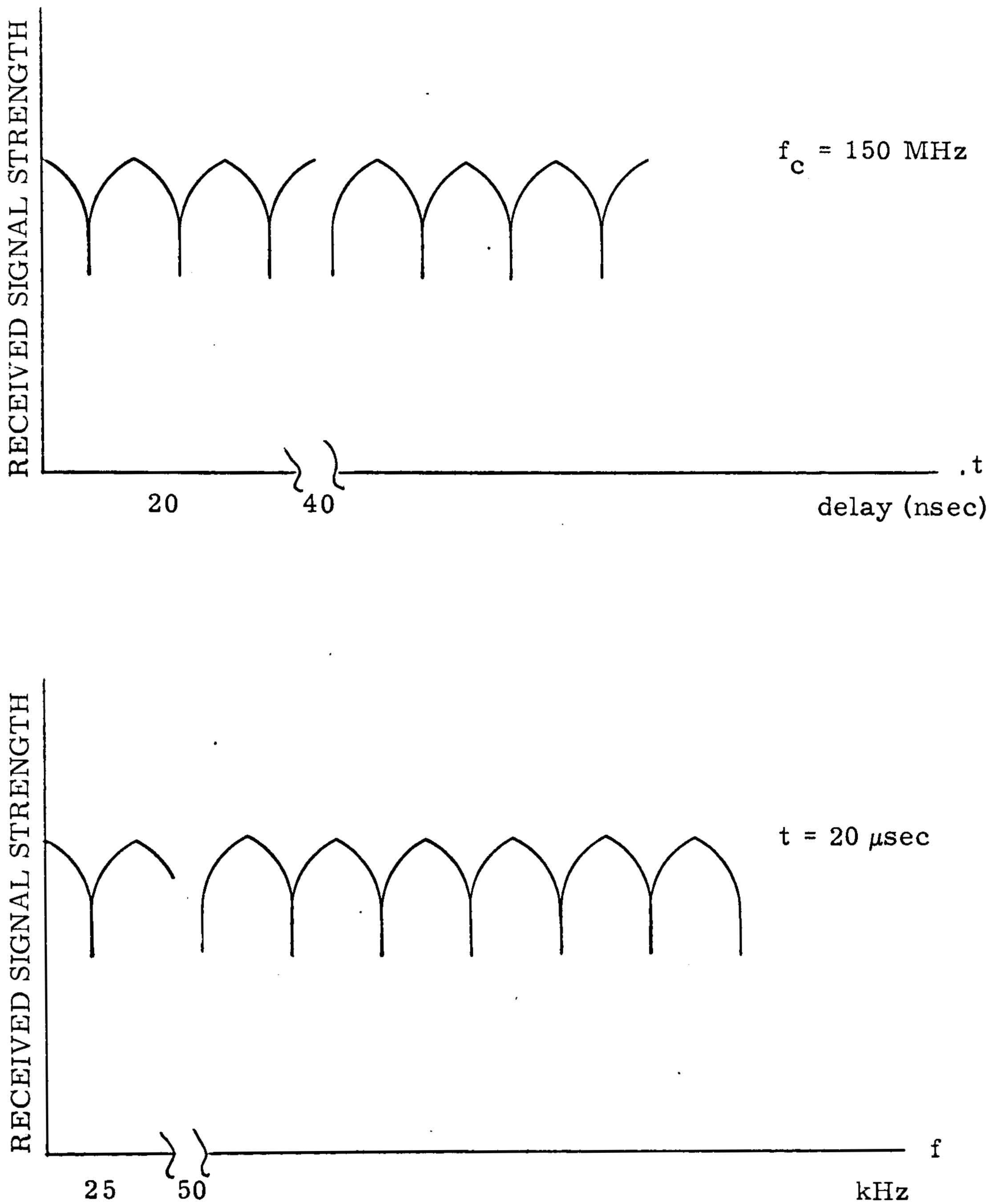



Figure 19. Multipath Interference Pattern

that combat these effects require a high time bandwidth product—that is, a bandwidth greater than that normally required to support the information rate. The required bandwidth and the maximum information rate are determined by the delay between the direct and reflected paths.

Figure 20 shows the effect of a 50 kHz bandwidth limitation on aircraft position (altitude, latitude, and longitude) for the Atlantic Ocean area. The curves are loci of points with a constant multipath delay of 20 $\mu$sec. Curves are shown for aircraft altitudes of 15,000, 20,000, and 30,000 ft. Within the area (southward), an aircraft at the given altitude experiences a multipath delay greater than 20 $\mu$sec and can therefore resolve the direct and multipath signals. Outside the area (northward), more bandwidth is required to distinguish between the two paths. Figure 21 shows the same information for the Pacific Ocean area with a satellite centered at 150°W longitude.

For example, consider an aircraft that flies at an effective latitude between 20 and 60 deg and at an altitude of from 10,000 to 30,000 ft. The delay between the direct path and the specularly reflected path would then be expected to vary from 7 to 56 $\mu$sec. If short pulses were used to discriminate against the multipath fading, they would have to be less than 7 $\mu$sec long, requiring a minimum bandwidth of about 142 kHz. Moreover, the pulse would have to be separated so that the multipath of the trailing edge of one pulse would not overlap with the leading edge of the next. Considering only one multipath signal (up or down link) and a bandwidth limit of $1/7\mu$sec, the maximum pulse repetition rate would than be $(7+5.6\ \mu\text{sec})^{-1}$ or 16,000 bps. If the same multipath is experienced on both up and down links, this rate is approximately halved.

If the above-mentioned aircraft used multiple-tone frequency diversity to combat multipath, the requirements could be similarly derived. With frequency diversity, the approach is to transmit frequencies separated sufficiently so that they do not all fade simultaneously. The bandwidth requirement is approximately the difference between the highest and lowest frequencies employed. Figure 19 shows the multipath interference pattern as a function of frequency. It also shows that if one frequency is in a deep fade, another separated 1/2t will be at peak. For the hypothetical aircraft, the minimum delay might be as little as 7 $\mu$sec, requiring a distance between frequencies of about 70 kHz or half the bandwidth required by time discrimination; depending on the sharpness of the nulls, even less distance between frequencies might be required. One further aspect of frequency diversity should be cited,

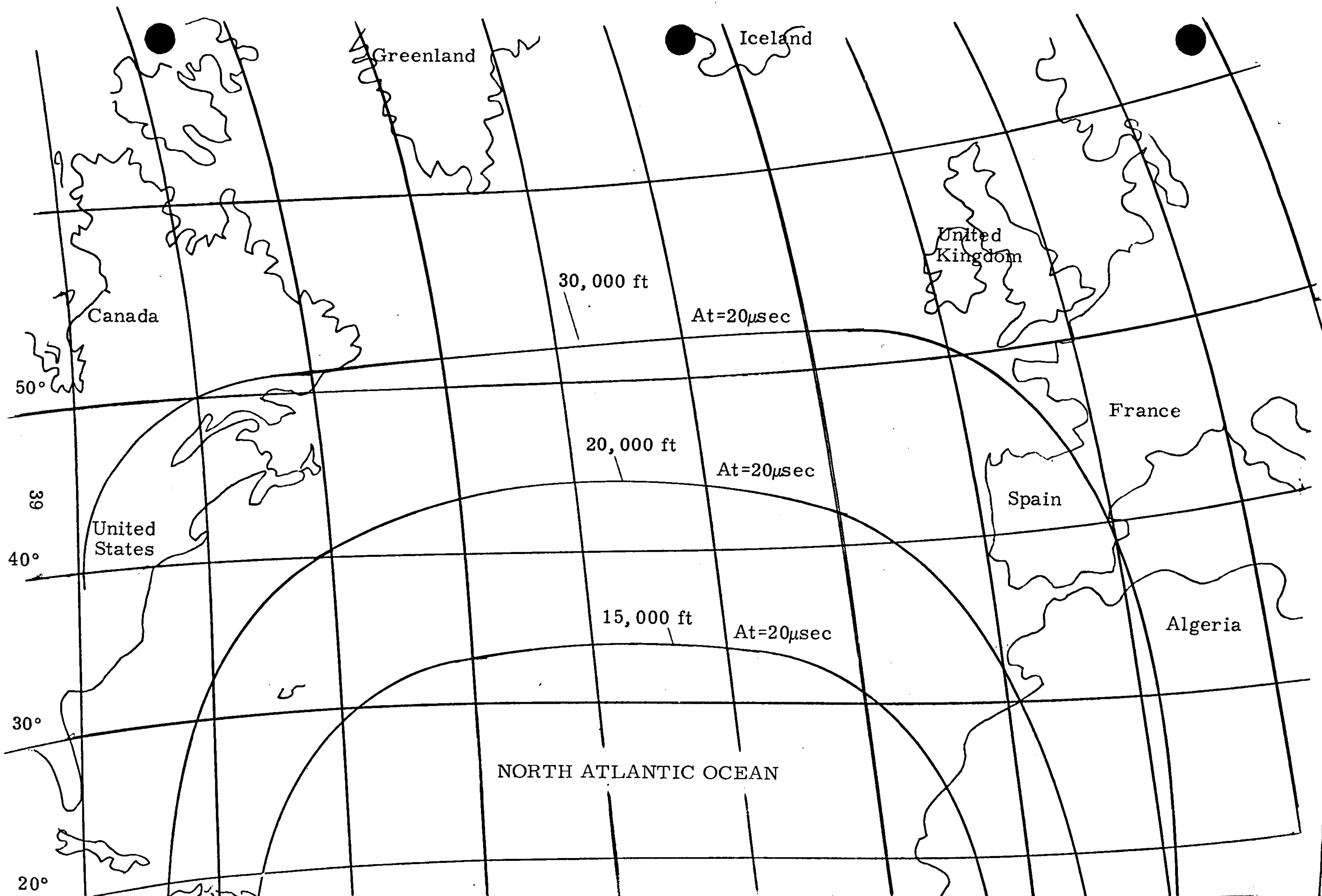



Figure 20. The Effect of a 50 kHz Bandwidth Limitation on Aircraft Position (altitude, latitude, and longitude) for the Atlantic Ocean Area

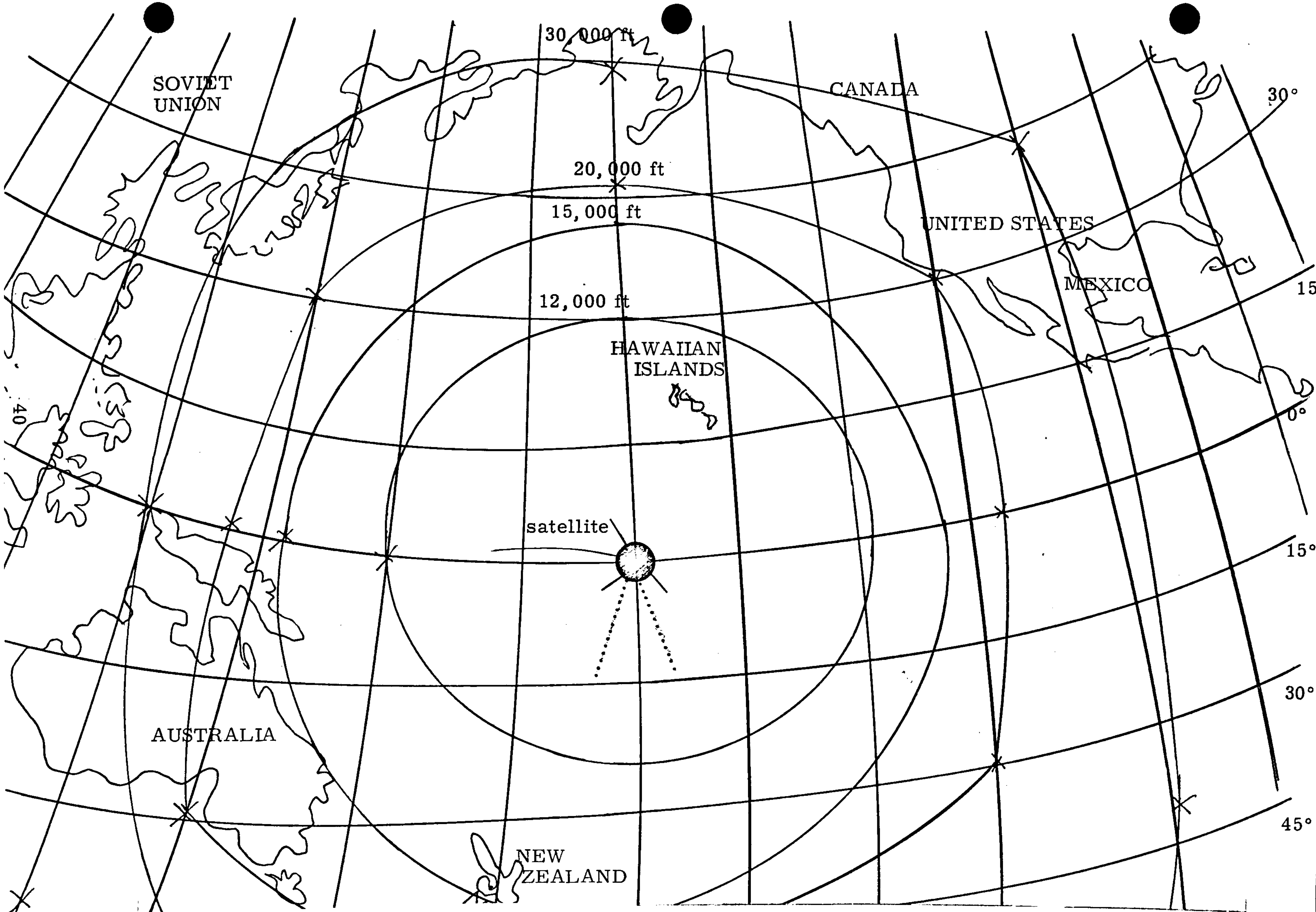



Figure 21. The Effect of a 50 kHz Bandwidth Limitation on Aircraft Position (altitude, longitude, and latitude) for the Pacific Ocean Area With a Satellite Centered at 150°W Longitude

although it does not have a first order effect on the bandwidth. If only 2 frequencies separated by 70 kHz were used, they would be in deep fades at multipath delays that are even multiples of 7 $\mu$sec—namely, 14, 28, 42, and 56. The addition of another frequency halfway between the other two frequencies would eliminate fades at delays of 14 and 42 $\mu$sec; a fourth frequency between the first and second or the second and third frequencies would eliminate the two remaining possible fades. Consequently, the shorter the multipath delays, the greater is the bandwidth requirement; and, the wider the range of delays, the greater is the number of frequencies required. Like time discrimination, the pulse rate of frequency diversity is limited by the maximum multipath. The longer the multipath delay, the more closely spaced are the frequencies and the longer must be the pulses to distinguish the frequencies. For this example, the maximum pulse rate would be 18,000 bps.

### 4.3 Performance of CSE in the Multipath Environment

A third approach to combat multipath interference with high time-bandwidth signals is swept FM, which is the basic waveform of CSE. By sweeping a carrier frequency across the band, frequency diversity is obtained. The width of the frequency sweep must be 1/2t, which is 70 kHz for the example, and the maximum number of sweeps per second is 16,000 (like the time discrimination example). At a rate higher than 16,000 sweeps per second, a maximum delay would cause interference in the compressive filter of the receiver.

The level of the interference caused by multipath is also a function of the pulse shape at the input to the detector. For the 2400 bps mode, the pulse envelope of both the direct and the multipath signals would ideally be of the $\frac{\sin x}{x}$ form with the first null occurring 20 $\mu$sec from the peak. The effect of the interference will also vary with its phase relationship to the direct signal. When the direct and reflected signals are out of phase, fading can occur. If the multipath is delayed at least 20 $\mu$sec, its amplitude will be at least 13 dB lower than the direct signal; neglecting attenuation during reflection, this level would occur at the first side lobe of the multipath pulses corresponding to about a 30 $\mu$sec delay. Subsequent lobes corresponding to delays 50, 70, 90, etc. microseconds would at least be 18, 21, and 23 dB below the direct signal. For delays less than 20 $\mu$sec, the interference effects can be severe. At 15 $\mu$sec it is about 10 1/2 dB below the direct signal; and at 10 $\mu$sec it is about 7 dB below the direct signal.

These calculations have assumed that the direct signal is sampled at its peak. However, as the multipath signal approaches the direct signal the effective geometric center of the combined pulse will be shifted towards the multipath (see Figure 22 ). Thus, the synchronization (based upon on early-gate late-gate (EGLG) that calculates this center) will experience an error compounding the effect of the interference.

A multipath signal of equal amplitude to the direct signal will shift the center of the detected pulse by half its relative delay. Figure 22 shows the average envelope of the direct signal combined with a multipath signal delayed 10 and 15 $\mu$sec. For the 15 $\mu$sec delay, the EGLG would lock to a point delayed 7.5 $\mu$sec relative to the direct signal. At this point, the signal is only three quarters its maximum value. When the multipath signal is exactly out of phase with the direct signal, complete cancellation can occur. Whenever the multipath is within 20 $\mu$sec of the direct signal and has a commensurate amplitude, the system is subject to deep fades.

CSE's ability to discriminate against multipath is comparable to two other wideband modulation techniques, time-division multiplexing and phase-shift code division. In all three cases, time resolution (and consequently multipath discrimination) is determined primarily by the inverse of the transmitted signal's bandwidth.

The CSE modulation technique differs from time division in two respects. Since the CSE employs continuous transmission, its average power is not affected by duty factors as is the pulse transmission of time division. Moreover, the channels afforded by the frequency sweep slopes are compatible with either controlled-access or random-access system organizations, whereas time division is more suited for controlled access techniques. The major difference between the CSE approach and phase-shift code modulation is the latter's phase coherence. While coherent systems have a better error rate vs signal-to-noise ratio characteristic, there is associated with coherent detection the problem of phase locking on a highly mobile platform. The major advantage of swept FM, compared to pseudonoise, it its tolerance to doppler shifts.

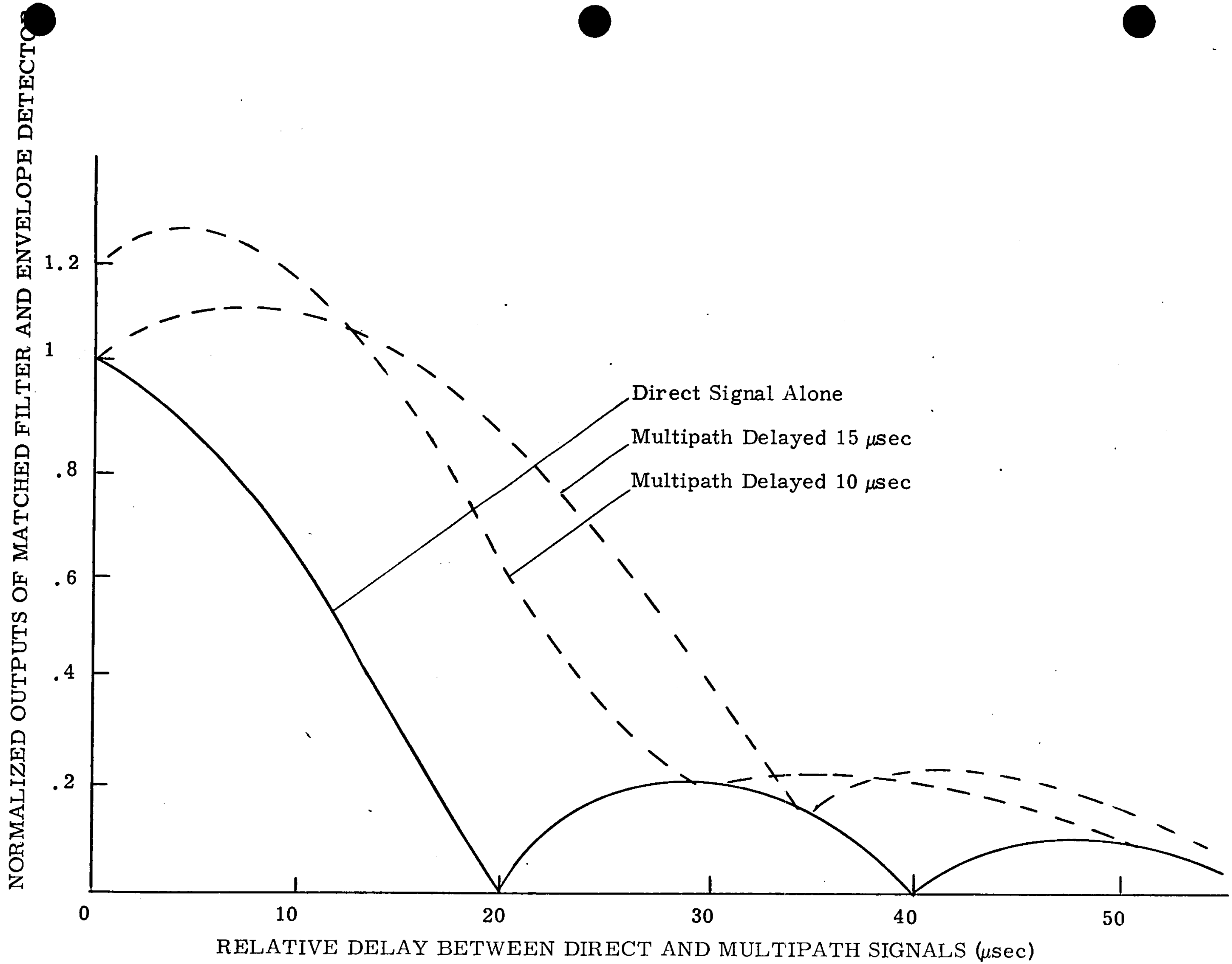



Figure 22. Average Envelope of Direct and Multipath Signals of Equal Amplitude

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