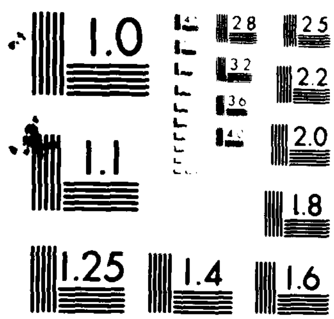
AD-A108 748 SRI INTERNATIONAL MENLO PARK CA F/G 9/5
DESIGN OF A TRANSPARENT-TRANSPONDER SYSTEM.(U)
NOV 80 D T MAGILL, A A BURNS, D J BARNES DNA001-78-C-0380
UNCLASSIFIED DNA-5722F NL

END
DATE
FILMED
1 82
DTIC


1.0
2.8
2.5
3.2
2.2
3.6
2.0
1.1
1.8
1.25
1.4
1.6

LEVEL (12)

DNA 5722F

AD A108748

DESIGN OF A TRANSPARENT-TRANSPONDER SYSTEM

D. T. Magill
A. A. Burns
D. J. Barnes
SRI International
333 Ravenswood Avenue
Menlo Park, California 94025

1 November 1980

Final Report for Period 1 August 1978—30 June 1980

CONTRACT No. DNA 001-78-C-0380

DTIC FILE COPY

APPROVED FOR PUBLIC RELEASE;
DISTRIBUTION UNLIMITED.

THIS WORK SPONSORED BY THE DEFENSE NUCLEAR AGENCY UNDER RDT&E RMSS CODE B322078462 I25AAXHX63804 H2590D.

Prepared for
Director
DEFENSE NUCLEAR AGENCY
Washington, D. C. 20305

81 12 22 046

410.14

Destroy this report when it is no longer needed. Do not return to sender.

PLEASE NOTIFY THE DEFENSE NUCLEAR AGENCY, ATTN: STTI, WASHINGTON, D.C. 20305, IF YOUR ADDRESS IS INCORRECT, IF YOU WISH TO BE DELETED FROM THE DISTRIBUTION LIST, OR IF THE ADDRESSEE IS NO LONGER EMPLOYED BY YOUR ORGANIZATION.

6 3 " 5 TT

UNCLASSIFIED

SECURITY CLASSIFICATION OF THIS PAGE (When Data Entered)

REPORT DOCUMENTATION PAGE		READ INSTRUCTIONS BEFORE COMPLETING FORM
1. REPORT NUMBER DNA 5722F	2. GOVT ACCESSION NO. AD-A108	3. RECIPIENT'S CATALOG NUMBER 748
4. TITLE (and Subtitle) DESIGN OF A TRANSPARENT—TRANSPONDER SYSTEM		5. TYPE OF REPORT & PERIOD COVERED Final Report for Period 1 Aug 78—30 Jun 80
		6. PERFORMING ORG. REPORT NUMBER SRI Project 7744
7. AUTHOR(s) D. T. Magill A. A. Burns D. J. Barnes		8. CONTRACT OR GRANT NUMBER(s) DNA 001-78-C-0380
9. PERFORMING ORGANIZATION NAME AND ADDRESS SRI International 333 Ravenswood Avenue Menlo Park, California 94025		10. PROGRAM ELEMENT, PROJECT, TASK AREA & WORK UNIT NUMBERS Subtask I25AAXHX638-04
11. CONTROLLING OFFICE NAME AND ADDRESS Director Defense Nuclear Agency Washington, D.C. 20305		12. REPORT DATE 1 November 1980
		13. NUMBER OF PAGES 86
14. MONITORING AGENCY NAME & ADDRESS(if different from Controlling Office)		15. SECURITY CLASS (of this report) UNCLASSIFIED
		15a. DECLASSIFICATION/DOWNGRADING SCHEDULE

16. DISTRIBUTION STATEMENT (of this Report)

Approved for public release; distribution unlimited.

17. DISTRIBUTION STATEMENT (of the abstract entered in Block 20, if different from Report)

18. SUPPLEMENTARY NOTES

This work sponsored by the Defense Nuclear Agency under RDT&E RMSS Code B322078462 I25AAXHX63804 H2590D.

19. KEY WORDS (Continue on reverse side if necessary and identify by block number)

Transionospheric Radio Propagation
Channel Sounding
Spread-Spectrum Signals
Coherent Bandwidth

20. ABSTRACT (Continue on reverse side if necessary and identify by block number)

This report describes and discusses the engineering specifications of a proposed system of great flexibility and general applicability for directly determining the effects of disturbed plasmas on a wide variety of actual or proposed signal waveforms. Both rocket-borne and satellite applications and missions are possible, and numerous signals in the VHF-UHF range (to about

DD FORM 1473 1 JAN 73 EDITION OF 1 NOV 65 IS OBSOLETE

UNCLASSIFIED

SECURITY CLASSIFICATION OF THIS PAGE (When Data Entered)

1.5 GHz) can be dealt with simultaneously without undue mutual interference. The system operates by retransmitting back to earth signals that originally had been modulated onto the lower sideband of an 8-GHz uplink carrier. If they passed through regions of naturally occurring or man-made plasma disturbances, the downlink signals would suffer degradations similar to those produced by the high-altitude nuclear environment. Even when the uplink transmitter is collocated with the downlink receiver (a convenient arrangement), the rather high frequency of the uplink signals ensures that plasma effects will be minimal on the uplink. The flexibility to readily change the types of downlink signals and their parameters makes this system particularly attractive for a satellite application because of the long-term nature of many satellite missions. Alternative system configurations and design tradeoffs are considered in detail.

TABLE OF CONTENTS

LIST OF ILLUSTRATIONS

LIST OF TABLES

I INTRODUCTION

Part of the design study reported here was concerned with determining which system alternatives would be the most effective for attaining the goals of a proposed, general-purpose system for measuring the effects of disturbed ionospheric plasmas on communications, surveillance, and position-location systems. The heart of the proposed general-purpose test and measurement system is a "transparent" transponder capable of passing a large variety of wide-bandwidth signals, several of which could be passed through simultaneously. Thus, channel sounding signals, channel diagnostic signals, simulated or actual systems signals, and ECM-type signals can be dealt with at the same time. The transponder has been designed to be as simple as possible to reduce unit cost to a minimum. This makes it feasible to employ it in a rocket application as well as in a satellite. The rest of this report describes the system, its subsystems, and the specifications developed for each of the elements.

The ground-based terminal equipment and system elements are conceived to have three functions. The first is to perform channel sounding and diagnoses. The second is to provide the capability for simulating generic types of signaling waveforms, with emphasis on spread-spectrum techniques. The final function is to supply an interface that would allow flexible and convenient use of actual signals from existing and prototype systems. More emphasis has been placed so far on the first two of these functions than on the third.

A brief description of how the transponder works is appropriate in this section. The desired downlink signals from the transponder are in the VHF-UHF range, and are transmitted up to the transponder as lower sidebands of an 8-GHz carrier (see Figure 1). Figure 2 illustrates the frequency relationships. The main unit of the transponder down-converts those sideband signals to baseband, where they are filtered out and transmitted back to the surface using whatever power amplifiers are

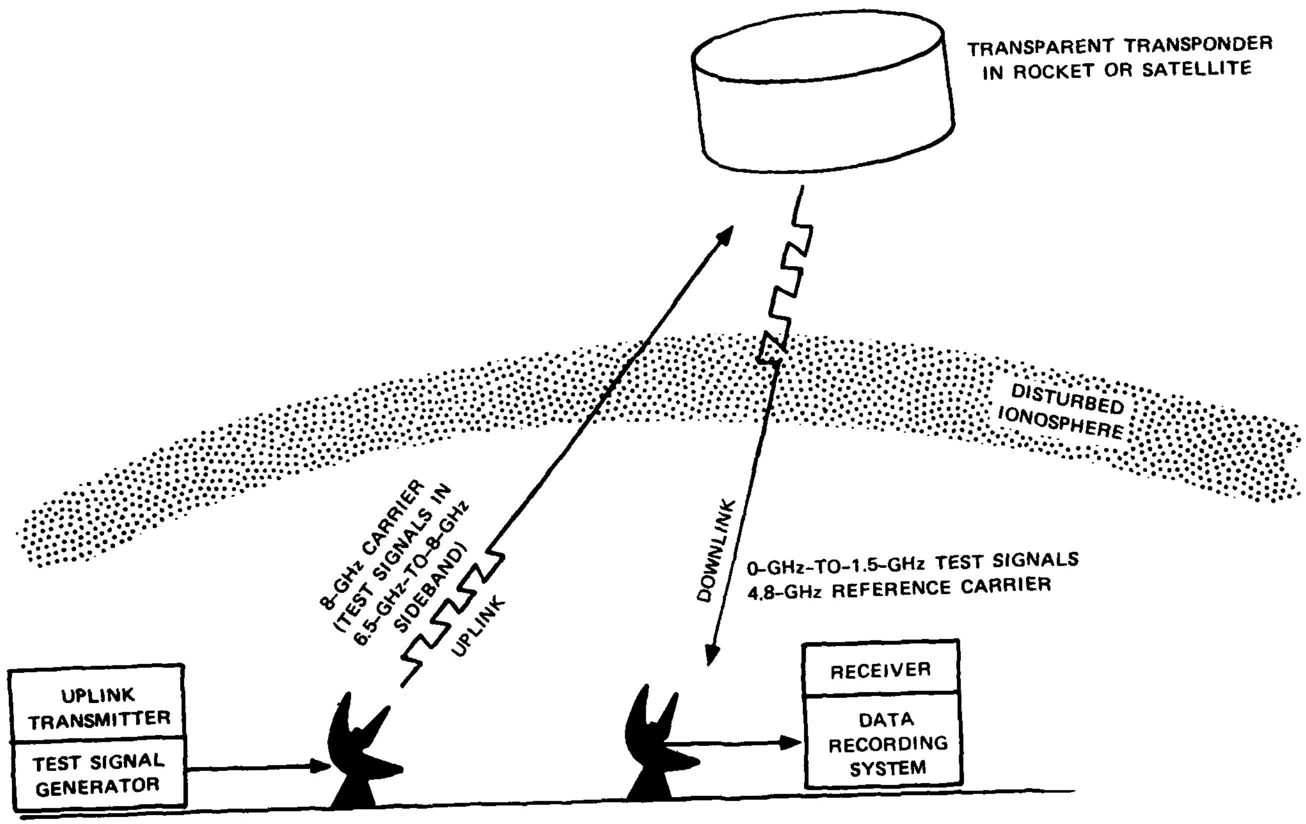


FIGURE 1 TRANSPARENT TRANSPONDER APPLICATION

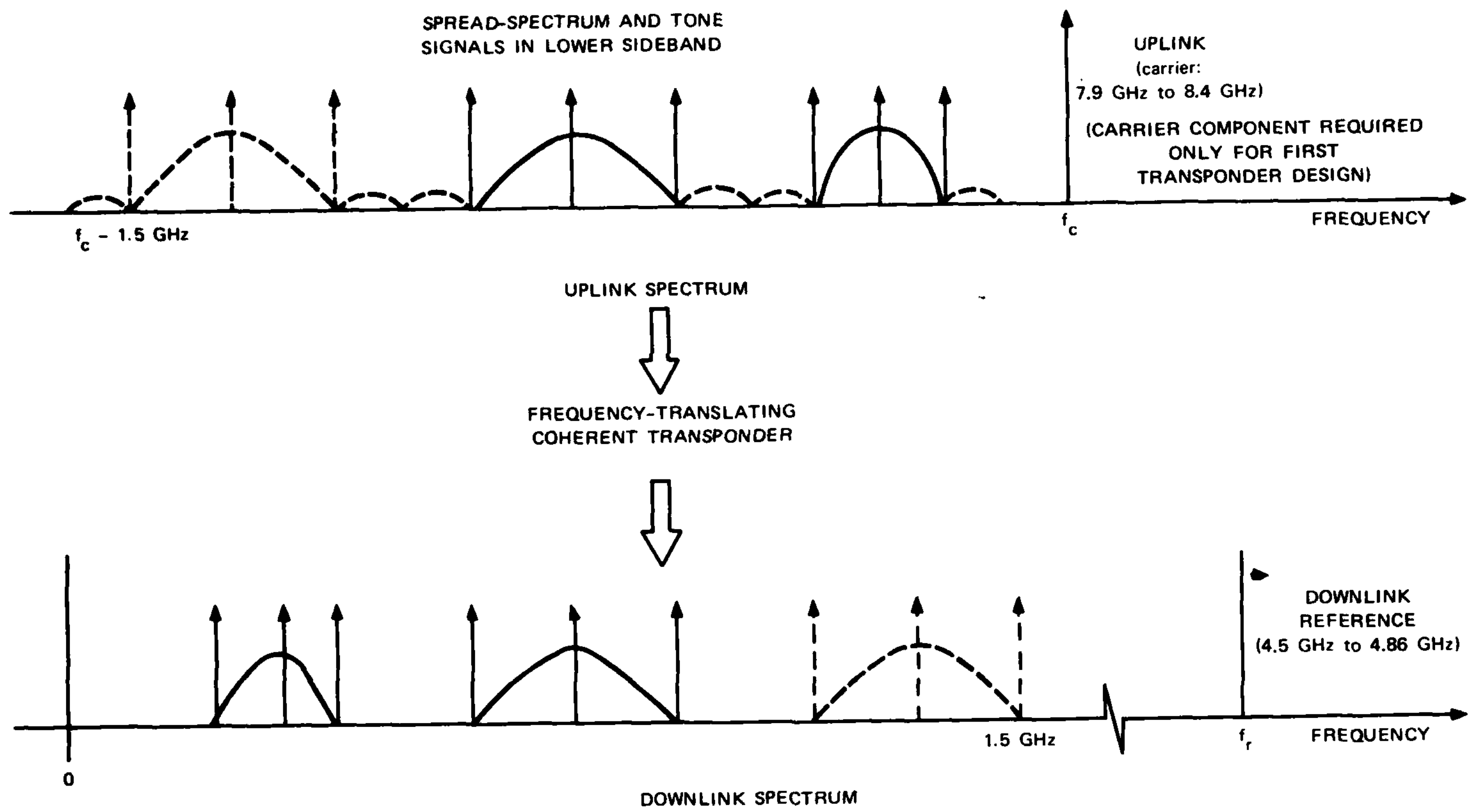


FIGURE 2 UPLINK AND DOWNLINK SPECTRA OF SPREAD-SPECTRUM TECHNIQUES PROPAGATION EXPERIMENTS

needed for the application. An SHF downlink reference signal is also generated in the transponder. A detailed description of the transponder and its operation appear in Section IV-B. Its design went through several iterations; the final design is superior because of its relative simplicity.

This effort was concerned almost totally with general hardware specifications and design configurations. Very little will be said here about ionospheric radio propagation under degraded conditions, which is well-covered in numerous DNA reports by various organizations, nor will specific applications for the transparent-transponder system be considered beyond the general remarks here.

We have already mentioned that the unit cost of the transponders is low enough to be worthwhile in rocket applications. Those applications include flights in conjunction with barium releases such as STRESS, PLACES, and SECEDE, and flights involving the naturally disturbed equatorial and auroral ionosphere. All those environments possess characteristics similar to those produced by high-altitude nuclear detonations, although obviously not so severe. Although the initial costs would be larger, the longer lifetime afforded by carrying the transparent transponder in a satellite would greatly enhance the payoff. We note that most of the wideband satellite tones can be passed through the transponder (which provides an even higher frequency phase reference), so there is compatibility with that ground-station equipment as well.

In early 1979 we were requested to study the possibility of using this system during the then upcoming PLACES test series. A subset of the transparent-transponder system was developed and suggested for that particular application. However, a decision was made to use special-purpose equipment, and we thereafter pursued design of the more flexible general-purpose system described here.

II MEASUREMENT OBJECTIVES

A. Channel Characteristics

Unlike many time- and frequency-spread channels, such as typical terrestrial links, the transionospheric channel is inherently dispersive as well. In the undisturbed ionosphere, dispersion is not usually a problem unless the fractional bandwidth is large and the center frequency is relatively low. Although most simple treatments of dispersion predict symmetrical time-spreading, pulse dispersion is actually asymmetric,[1*] and the emergent pulse will have a stretched-out "tail" if there is appreciable dispersion. It may be difficult to distinguish between this situation and a strictly time-spreading channel, which also produces an asymmetric impulse response with a tail due to off-line-of-sight scattering. Even though dispersion does not result in coherent-bandwidth reduction as in the case of frequency-selective fading, the resulting signal distortion can seriously degrade the performance of a system unless some kind of compensation is applied, such as including a dispersive filter having the opposite phase-shift-versus-frequency characteristic.

Dispersion also affects frequency-selective fading produced by scattering from ionospheric irregularities under disturbed conditions. A discussion of the practical or theoretical effect of having frequency-dependent scattering (this channel seems to be rather unique in this) is beyond the scope of this work. We do note that the condition of wide-sense stationarity[2] for the channel in the frequency domain is violated, which means strictly that characterization in terms of a channel delay profile (or its Fourier transform, the coherent bandwidth), may not be very accurate. It may suffice, however, to model the channel as a cascade of dispersive and frequency-selective fading elements.

---

*References are listed at the end of this report.

The transionospheric channel is usually characterized by its ability to support rather wide-bandwidth signals, even under disturbed conditions. For a given set of conditions, however, the fractional coherence bandwidth exhibits a frequency dependence as rapidly varying as f^3, which means that it shrinks rather quickly as the center frequency is reduced. Thus a choice of a fairly low operating frequency might provide a good simulation of a higher-frequency system in the more severe nuclear environment. Because the measure of dispersion has an f^{-3} dependence, what might be called the fractional "dispersion-limited bandwidth" varies as f^{-2}, which is less rapid than the coherence bandwidth. This says that dispersion will be eventually dominant as the frequency is raised, but both effects may be negligible by that point.

At the wavelengths of interest, the ionosphere is also birefringent due to the geomagnetic field. This is not much of a problem if care is taken to maintain good antenna-polarization circularity, and propagation paths nearly transverse to the field are avoided. Because the two circularly polarized modes have different group velocities, Faraday rotation will be seen only on the overlapping part of an originally received linearly polarized pulse.

B. Channel Sounding Objectives

One objective of this work is to correlate measured bit error rate (BER) on representative modems with the measured channel response. The measurements will be performed in both the time and frequency domains, thereby characterizing the time-variant channel more accurately than previously.

This effort is directed toward measuring the effects of disturbances affecting the transionospheric satellite communication (and position-determination) channel. The disturbance may occur on either the uplink or the downlink, or both. While many satellite communication modems employ spread-spectrum modulation--e.g., the URC-55 and -61 and the USC-28, and could be used for time-domain channel sounding--their construction is such that it is not convenient to perform the desired channel sounding measurements. A channel sounding (time domain) spread-spectrum unit, which is optimized for the channel sounding function, will have, in general, different parameters--e.g., phase-lock loop noise bandwidths--than the satellite modem. Consequently, it is desirable to simultaneously measure the BER with the satellite communication modem and sound the channel. The channel should be sounded in both the frequency and the time domain.

Our system design, in fact, does permit all three signals to be present. The design is such that any data modem with a 70-MHz interface may be used with the system, subject to the constraint (discussed later in the report) that excessive mutual interference does not occur between the three signals. Simulated ECM (jamming) signals can also be passed through the system. Note that the system design is very flexible and it is not necessary to package the data modems for a spacecraft or rocket environment as would be the case if the transmitters were located in the vehicle.

C. Error-Rate Determination

Determination of error rate requires the use of an external data modem and bit-error-rate measurement equipment. The modem must be connected to the auxiliary parts of the experimental equipment. Further discussion of this arrangement is presented in Section IV-A-4.

The error rate in future experiments needs to be measured at an appropriate data rate. If the data rate is too low, it will be impossible to accurately measure the BER in a time period during which the propagation conditions are fixed. If the data rate is too high, too much uplink and downlink power will be required. Also, the data modem may be incapable of transmitting at rates in excess of 10 or 100 kbits/s. Due to these constraints we do not expect to be able to accurately identify small changes in BER as they relate to particular channel characteristics. However, we expect after many experiments to correlate higher probability of bit error with particular channel characteristics. That is, certain channel characteristics may consistently increase the BER from 10^{-3} or

10^{-5} to 10% to 50%. We expect to be able to identify such situations with the proposed experimental equipment.

III SYSTEM CHARACTERISTICS

A. System Description

In this section we present some general considerations and choices in the selection and design of the proposed experimental system. We begin by considering general alternative systems and proceed to frequency selection tradeoffs.

1. Alternative Systems

Measurement of man-made or natural ionization can be accomplished by a variety of methods. We assume that measurements are to be made on the ground as opposed to measurements by spacecraft or a rocket package because ground-based measurements offer greater flexibility, accessibility, and lower cost. Given that channel measurements are performed on the ground, there are only two potential locations for the transmitter or signal source: (1) a spacecraft or rocket package, and (2) the ground. The basic premise underlying all of our work is that only a ground-based source provides the flexibility needed to accommodate different circumstances and objectives.

If the transmitter is located on the ground, it requires a transponder in a spacecraft or a rocket for relaying the transmitted signal back down through the signal-degrading medium. Thus, an uplink signal is required that must be of sufficiently high frequency that this signal does not encounter measurable distortions.*

The transmitter may be either collocated with the receiver (monostatic) or placed in separate locations if logistics or a

*Alternatively, one could "sound" the disturbed media on the uplink and return the signal for ground reception on a very much higher frequency that does not encounter the distortion. But this approach exacerbates the problem of coping with interference.

frequency-management problem so dictates. Collocation, when feasible, offers significant operational advantages. The recommended system is configured so that it may be operated either monostatically or bistatically.

Another alternative pair of choices is sounding the channel in the frequency or in the time domain. With the former approach a set of coherent sinusoids (i.e., δ-functions in the frequency domain) is transmitted and their relative amplitudes and phases are measured. Because these quantities are samples of the channel time-varying transfer function, the channel is directly characterized in the frequency domain. For a linear channel this frequency-domain characterization can be readily converted to a time-domain representation by using an inverse Fourier transform. Alternatively, a time domain characterization can be measured directly in the time domain. With this approach a short* pulse [or, equivalently, a spread-spectrum signal generated by a maximal-length sequence (M-sequence)] is used to measure the (time-varying) channel impulse response. Since the channel output is the convolution of the time-varying impulse response and the input signal, this technique will strictly yield a good estimate of the impulse response only to the extent to which the effective input signal is a δ-function. In practice, it is really only necessary for the bandwidth of the input signal to be as great as that of the widest-bandwidth signal contemplated. This is equivalent to bandlimiting with a filter, and it is important to control the rolloff so as to minimize time sidelobes, which can confuse interpretation. (The same bandlimiting considerations apply for the multiple-tone approach as well.) Assuming a linear channel, this time-domain characterization can be converted to a frequency-domain representation by usin a Fourier transform.

Numerous Fourier-transform relations exist between the various channel measures as well;[2] here we only note that because of the uncertainty relation, a long-duration channel impulse response or delay profile

*Compared to the duration of functions such as the channel delay profile.

implies a narrow coherence bandwidth, which occurs when RF energy is scattered along paths sufficiently far (depending on the circumstances) from the direct path.

Both alternatives in principle provide the same information. The real differences between them reside in practical considerations, such as the number of well-calibrated receiver channels required with the multiple-tone approach. The multiple-tone approach concentrates the RF energy into a series of narrow spectral slots, which allows interference to be filtered out. Using a PN sequence overcomes the problem of a very high peak-to-average power that would happen with straight time-domain probing.

Frequency-domain sampling (with only one tone) is the natural approach for a Doppler-spread channel that produces no significant time spread. A single impulse would suffice for a time-invariant channel that did have a finite-width delay profile. The channel of interest here exhibits both time- and frequency-scattering. However, even for rapidly moving satellites the time spread or delay-profile extent will be much less than the correlation time (submicroseconds versus milliseconds). This means that the channel needs to be sampled at a high rate but only in relatively infrequent bursts. If 2F is the widest bandwidth of interest, and T is the greatest duration of the delay profile, either the impulse response needs to be sampled 1/2F s apart or the transfer function needs to be sampled 1/2T Hz apart, so the total number of sampling taps and tones required per burst is the same for both approaches--namely, 2TF. This quantity depends on the particular situation; for the wideband satellite it is 7 (see Reference 3), which must be near the practical minimum number. The burst repetition rate is of the kilohertz order, and is also the required (two-sided) bandwidth of the individual tone-approach receivers.

Because the multiple-tone approach requires a fairly large number of carefully matched receivers and a complex frequency synthesizer, it is not a very attractive approach in most circumstances. It may be reasonable in some applications to step a single receiver (or transmitter-receiver pair) rapidly through a series of tones. For example, suppose we need to support a system bandwidth of 2F = 300 MHz and the channel delay profile has to be measured out to T = 100 ns, which implies a coherence bandwidth of 10 MHz. This means that 30 tones minimum are required to cover the possibilities. So, if the channel coherence time is 1 ms, say, it would be necessary to step through the 30 tones in times short compared to 1/30 ms. This in turn implies that the instantaneous receiver bandwidth must be in excess of 30 kHz to allow the output time to settle down. Because that bandwidth is much smaller than the tone spacing of 10 MHz, stepping is a more reasonable approach than providing 30 receiver channels. There would be a 20-dB SNR penalty because of the wider bandwidth required for stepping compared with continuous multiple tones. Furthermore, a very complex frequency synthesizer(s) would be needed to ensure that phase data are accurately obtained.

Time-domain sounding is attractive because it directly obtains the channel impulse response and because it is relatively easy to implement the 2TF sampling taps. Alternative methods of correlation are the use of sliding delays, multiplexed correlators, and matched filters. Advantage can be taken, as discussed above regarding frequency-stepping, of the 3rd to 4th order of magnitude difference between the channel coherence time and its delay-profile extent. That is, it is not necessary to sample at all tap delays on every pulse. The main problem with time-domain sounding is achieving and maintaining synchronization, and data acquisition is more difficult than in the multiple-tone method because much higher sampling (burst) rates are required. However, the synchronization problem can be overcome in various ways, and time-domain sounding is the preferred approach for all but the smallest values of 2TF.

Actually, none of these approaches is precluded by the transponder, which has been designed, after all, to provide exactly that flexibility. The impact is on the ground-station equipment. It is necessary to decide what equipment to construct that will provide basic channel-sounding capability with enough parameter flexibility to be useful in different applications.

Some consideration was given to the possibility of a hybrid system that combined both time- and frequency-domain sounding. The greatly increased complexity of even the most straightforward implementations of such a system completely outweighs the slight advantage of reducing the number of frequencies needed. There are, however, some good reasons why a minimum CW tone capability should be included. First, nearly all ionospheric data have been collected using CW techniques; a CW signal would allow directly relating new results to old. Also, the ionosphere is intrinsically dispersive; both group delay and dispersive signal broadening occur even when there are no irregularities, so some kind of measure of electron content must be made. Finally, time-domain probing tends to deemphasize the higher frequencies in the channel (e.g., the triangular impulse of the self-convolved or matched-filtered PN code has an ideal sinc-squared spectrum); CW tones at the band edges would be useful when the correlation bandwidth is wide.

Thus, our recommended system includes both frequency- and time-domain channel sounding. We propose using three sinusoids per band of interest. Two sinusoids would be located near the upper and lower band edges while the third would be located near the band center. The PN sequency period (and receiver integration time) must be long enough to provide the required signal-to-noise ratio (SNR) and processing gain, but be sufficiently short to be responsive to the time variations in the propagation media.

One may wish to simultaneously characterize the propagation media in two or three frequency bands simultaneously. If the sounding signals in these bands are coherently related, it permits to some degree a characterization of the total channel. For example, if the channel is not too complex, measurements in the vicinity of 100 MHz and in the range of 200 to 300 MHz may serve to characterize the channel from 100 to 300 MHz. Here we assume that some physical laws dictate that measurements in selected portions of the entire band are adequate to characterize the entire band.

2. System Tradeoffs

System tradeoffs fall into two different categories: (1) system parameter choices, and (2) equipment design choices. The former includes such things as the carrier frequency, the number of bands, and the PW code lengths and rates. The latter includes questions such as the desired tradeoff between transmitter power and antenna gain. The subsequent sections consider these tradeoffs in detail.

B. Frequency-Selection Criteria

Three separate frequency choices are required and will be addressed separately as: (1) the downlink channel sounding frequency, (2) the downlink pilot or reference frequency, and (3) the uplink frequency. In general, as the frequency is increased, the available bandwidth is increased, and there is a tendency for components to be more costly. However, it should be noted that the principal component saving is obtained by selecting frequencies used commercially or militarily so that products, such as power amplifiers, are readily available.

1. Downlink Channel Sounding

Use of as low a downlink frequency as reasonably possible is desirable because the effective channel bandwidth is inversely proportional to the cube of the carrier frequency. A 5-MHz bandwidth at 100-MHz carrier represents a nominal value. Use of lower frequencies makes it more difficult to find an available band. For example, 100 MHz is centered in the commercial FM band. Consequently, the choice of downlink channel sounding carrier frequency must reflect the availability of a channel. The channel must be wide enough to adequately sound the disturbed media.

It should be noted that the available frequencies depend on the location of the experiment. For example, there may well be no FM stations in the middle of the South Pacific, while at Eglin Air Force Base, contention with the FM band is a definite problem.

Furthermore, the difficulty of the frequency allocation problem depends on the relay transponder vehicle. If it is a rocket that is operational for only a few minutes for occasional experiments, a temporary license may be relatively easy to obtain. By contrast, a synchronous satellite is a relatively permanent item that may require much greater effort in obtaining the desired frequency allocation. A non-stationary satellite may encounter even greater difficulties. Clearly, any satellite-borne transponders must be capable of being turned off when not actively participating in an experiment.

2. Downlink Reference Frequency

The downlink reference frequency must be chosen high enough that it encounters no significant attenuation or phase shift due to the disturbed media. Any frequency above several gigahertz should be adequate. The reference channel, being a single sinusoid, requires only enough bandwidth to accommodate one- or two-way Doppler (depending on the transponder configuration) and the frequency instability of the basic crystal source.

The downlink reference or beacon frequency provides a tracking signal that is useful in antenna pointing and initial acquisition. For cost reasons it is desirable to use manual as opposed to automatic tracting. Thus, the ground receive antenna beam must not be too narrow or initial acquisition and manual tracking will be too difficult. Since link losses require a fixed receive capture area, the downlink reference frequency should not be too high. Additionally, generation of power is more difficult at the higher frequencies. Consequently, the downlink reference frequency should be as low as possible without encountering significant distortion.

3. Uplink Frequency

The uplink frequency must be chosen sufficiently high that no significant attenuation or phase distortions take place on this link in the event it should pass through a disturbed medium. Any carrier frequency above several gigahertz should be adequate. The frequency selected must be capable of supporting the bandwidth of the transmitted signals, which, of course, is identical to the bandwidth of the downlink-channel sounding signal.

Since both the ground transmitter and the transponder front end (i.e., low-noise amplifier) are expensive items, it is desirable to choose a commonly employed frequency band. Readily available components should have a significant impact on cost reduction. We recommend the use of the uplink military satellite communication band--that is, the frequency range of 8.2 GHz.

4. System Considerations

Selection of the three frequency bands must be coordinated because they have a significant impact on the cost of the transponder. Since the transponder is most likely expendable it has a great impact on the overall system cost. Consequently, these cost factors must be carefully reviewed before the three frequency bands are selected. Further consideration of this matter is found in Section IV-B-1, which is devoted to alternative transponder designs.

For monostatic situations another system consideration is that it is highly desirable that the same antenna reflector suffice for the uplink and downlink pilot signals. For bistatic situations it is also desirable (from an interchangeability or flexibility viewpoint) that the same antenna reflector handle both the uplink and the downlink reference signal.

Since higher-order--e.g., second- or third-order--harmonics are frequently created in power amplifiers, the uplink frequency band should be higher than the downlink reference frequency band. In this case it is impossible for harmonics created in the ground power amplifier to interfere with the reception of the downlink reference frequency.

5. General Frequency Plan

The general frequency plan selected is illustrated in Figure 2. The uplink spectrum includes potentially four major components consisting of the carrier that serves as a reference for phase-locking purposes, and two or possibly three channel sounding bands. Normally we would expect to use only two of these bands, but the highest band may prove useful in some circumstances. Note that the highest band on the downlink, which is shown in dotted lines, is the lowest frequency or sideband on the uplink.

Within each band four signals are shown--three tones and a PN sounding signal. The three tones required for frequency domain sounding are uniformly spaced, with the center tone lying in the center of the band and the upper and lower tones lying in the nulls of the PN direct-sequence channel sounding signal. With this spacing of the tones we uniformly and fully cover the band. Thus, the relative phase shift between upper and lower tones should be a very sensitive measure of channel disturbance.

The PN-channel sounding power spectrum is illustrated for each band as the familiar sinc-squared form. In practice, the filtering required to avoid adjacent-channel interference will greatly alter the sidelobes with respect to the levels shown in the illustration. Note that the PN clock rate must be set exactly equal to the tone spacing if the tones are to lie in the spectral nulls of the PN-channel sounding signal. This condition, which should not be difficult to achieve, is desirable for two reasons. First, in this case the upper and lower tones do not interfere with the PN signal. Thus, the interference power is three times lower than indicated in Eq. (1) of Section IV-1-D. Similarly, the upper and lower tones encounter no interference power from the PN sounding signal. However, the center tone does encounter the interference given in Eq. (3) of Section III-D.

Note that Figure 2 is not drawn to scale. The bandwidth of the second (and of the third, if desired) band will be considerably larger, perhaps twice that of the first or lower frequency band.

C. Link Computations

In this section we evaluate the capacity quotient of three links: (1) the uplink, (2) the downlink reference or pilot signal, and (3) the channel sounding signals, which may exist in one, two, or three separate frequency bands.

We must determine the SNR in the transponder passband for the maximum range. This is significant because it determines how effectively we use the transponder power. Also, we need to determine the SNR in the transponder phase-lock loop closed-loop bandwidth. This SNR is significant because it limits the system's ultimate accuracy in making phase measurements.

It is necessary to determine the SNR in the downlink reference phase-lock closed-loop, equivalent-noise bandwidths. This SNR also determines, after suitable scaling, the accuracy of our phase measurements.

With respect to the downlink-channel sounding signals it is necessary to compute the SNRs for the channel-sounding PN signal and the three tones in each of the two frequency bands. Note that it is also necessary to calculate the SNR for the data signal, which may be in one or both of the frequency bands. These SNR computations must reflect the power sharing that takes place between these three signals in the transponder power amplifier.

It should be noted that the ground-recorder phase-lock loop (PLL) need not have an SNR as high as the transponder PLL. Section V-C analyzes the phase jitter effects.

Table 1 presents the nominal values for the uplink. The transmitter antenna gain of 24 dB corresponds to a 1-ft dish, which is 50% efficient. At the 8-GHz carrier frequency the half-power bandwidth is 8.4°. Thus, the tracking problem should not be very severe, and open-loop control should be easily accomplished.

The RF bandwidths for the low and high bands are assumed to be 25 and 50 MHz, respectively. These bandwidths are significant in that it is desirable to have a transponder SNR greater than 6 dB in this bandwidth. In SNR lower than this value would result in significant wasted downlink power. That is, an excessive amount of reradiated noise would be present on the downlink. This inefficiency could only be combatted by the transponder transmitting at a high power level, which requires a larger, heavier, and more expensive transponder. That is, if half of the power

Table 1

NOMINAL UPLINK PARAMETERS

Carrier frequency = 8 GHz	
Transmitter antenna gain = 24 dB	
Maximum range = 500 km	
Receiver noise temperature = 1000 K	
Receiver antenna gain = 6 dB	
Miscellaneous losses = 3 dB	
Transponder bandwidths	
Low band	= 25 MHz
High band	= 50 MHz
Reference PLL	= 50 kHz
Desired signal-to-noise ratios	
Low band	= 6 dB
High band	= 6 dB
Phase-lock loop	= 30 dB
Minimum transmitter power	
Low band	69.2
High band	= 139.4
Reference	= 34.6
Total	243.2 watts

in the AGC amplifier bandwidth is noise power, then the transponder power amplifier power output must be increased by a factor of two to provide the desired signal downlink power level.

It should be noted that these link computations assume that there are separate AGC or limiting amplifiers for each band. It is more likely that there will be only one AGC system in the transponder. While a single AGC poses no significant problem there is a question of what is the SNR in the IF passband of the AGC amplifier. For example, if separate 25- and 50-MHz bandwidth filters are provided, the required powers will be identical to those cited in Table 1. However, if one designs the AGC on the basis of one wide bandwidth covering both bands, the equivalent noise bandwidth of the receiver might easily be 400 MHz. In this case the ground transmitter power would need to be nearly 7.3 dB higher than the minimum values cited in Table 1.

The desired bandwidth for the phase-lock loop is estimated to be about 50 kHz. This bandwidth is required to achieve rapid acquisition in the presence of the expected Doppler and frequency uncertainties. The bandwidth could be reduced if a more complex phase-lock loop, such as a hybrid AFC/AGC loop, is used to speed acquisition. However, a somewhat more complex transponder would result.

We estimate that the required SNR in the phase-lock loop is 30 dB (see Section V). A lower value would result in a degraded measurement of the channel phase and amplitude distortion. A precise analysis of the performance is presented in ection V.

The effective noise temperature of the transponder is assumed to be 1000 K and the receiver antenna gain is chosen to be 6 dB. Miscellaneous losses are assumed to be 3 dB. With these assumptions the minimum transmitter powers (for the maximum range) are 69.2, 139.4, and 34.6 watts for the low and high bands and the reference signals, respectively. The total power is 243.2 watts. Note that the envelope is not constant because the transmitted signal is composed of three different signals. Consequently, the peak power capability of the ground transmitter must be considerably greater than 243 watts if IM cross products in the transmitter power amplifier are to be avoided. In fact, we find in Section IV-A-3 that the peak power requirement is 676 watts.

Nominal downlink parameters are presented in Table 2. The downlink carrier frequencies are assumed to be 0.2, 0.4, and 4 GHz. The transmitter antenna gain is assumed to be 0 dB for all three bands. For the high and low bands the receiver antenna gain is assumed to be 8 dB, corresponding to readily available helices. At 4 GHz we assume a 2-ft dish also having a gain of 24 dB. This assumes that we separate transmit and receive antennas in either a bistatic or monostatic configuration. Alternatively, if a single 2-ft antenna is employed for both transmission and

Table 2

NOMINAL DOWNLINK PARAMETERS

Carrier frequencies	
Low band = 0.2 GHz	
High band = 0.4 GHz	
Reference band = 4 GHz	
Transmitter antenna gain (all bands)	= 0 dB
Receiver noise temperature (all bands)	= 1000 K
Receiver antenna gain	
Low band = 8 dB	
High band = 8 dB	
Reference band = 24 dB	
Maximum range = 500 km	
Miscellaneous losses = 3 dB	
Required signal-to-noise ratio	
SS signals in 200-Hz bandwidth	= 40 dB
Tones in 200-Hz bandwidth	= 40 dB
Reference signal in 1-kHz bandwidth	= 20 dB
Required transmitter power	
Low band = 0.67 watts	
High band = 2.82 watts	
Reference band = 90 milliwatts	
Total prime power (10% efficiency)	= 36 watts

reception, it will be necessary to point the transmit antenna twice as accurately as implied by the 8.4° beamwidth cited earlier.

As for the case of the uplink, we assume that the maximum range is 500 km and that there are 3 dB of miscellaneous losses.

It is assumed that both the spread-spectrum signals and the tones require a 40-dB SNR as measured in a 200-Hz bandwidth. We assume that the reference tone must be received with at least a 20-dB SNR as measured in a 1-kHz passband. The significance of the SNR on the overall performance is assessed in Section V.

Assuming that zero margin is required, the downlink powers for the low, high, and reference bands are 0.67, 2.82, and 0.089 watts, respectively. If one assumes that the efficiency of the transponder in converting prime power to RF power is 10%, then approximately 36 watts of prime power are required. If a link margin of 3 dB (at the maximum range) is desired, then the prime power requirement is increased to 72 watts.

D. Mutual Interference

Three signals may share each of the two channel sounding frequency bands. These signals are: (1) the PN, or time-domain sounding signal, (2) the three-tone, or frequency-domain sounding signal, and (3) the data signal (for BER measurement). The SNR computations for each of these three signals must be modified to reflect the interference terms contributed by the other two signals.

The PN-channel sounding signal has an inherent processing gain against the three tones and the data signal. However, the total power of the tones should be significantly greater than the power spectral density of either the PN ranging or data signals. Consequently, these two signals with nominally low power spectral densities may be treated as additive background noise in addition to the receiver noise. Frequently the data signal will be a spread-spectrum signal. In this case it will have processing gain with respect to the other two signals. The magnitude of the processing gain will depend on the selected bit rate of the data modem. Lower bit rates are to be preferred because they offer greater processing gain.

If the data modem is not a spread-spectrum modem, then no processing gain advantage is possible. However, if the other signals are sufficiently low in power (not power spectral density), the data modem performance may still be acceptable. That is, since the other two signals have either large processing gain or very narrow detection bandwidths, their power levels may be low enough that they do not create significant interference with respect to the data signal.

The allowable level of additive interference to the data signal must be sufficiently low that it does not prevent one from observing the degradation caused by the channel disturbance. Nominally the total additive interference, including receiver noise and mutual interference terms, should be at least 10 dB below the desired signal levels. Higher interference levels will tend to mask the degradations introduced by the disturbed medium.

Now consider in detail the SNR at the detector inputs for the three subsystems. Let P_D, P_S, and P_T be the powers (in any one band) of the data signal, of the sounding signal, and of each of the three tones, respectively, as received at the ground-based receiver. Let W_D and W_S be the RF bandwidths occupied by the data and sounding signals, respectively, and b_D and b_S be the detection bandwidths, respectively. The detector bandwidth for the tone receiver is denoted b_T.

The SNR at the data detector input is given by

$$SNR_D = P_D[N_O + P_S/W_D + 3P_T/W_D]b_D \quad , \tag{1}$$

where N_O is the one-sided noise power spectral density ratio. The SNR at the sounding signal detector is given by

$$SNR_S = P_S/[N_O + P_D/W_S + 3P_T/W_S]\ b_S \quad . \tag{2}$$

The SNR at the tone detector is given by

$$SNR_T = P_T/[N_O + P_S/W_S + P_D/W_D]\ b_T \quad . \tag{3}$$

Joint occupancy of the same spectrum can yield satisfactory performance if: (1) the spread-spectrum processing gains W_D/b_D and W_S/b_S are sufficiently large, and (2) the power spectral densities P_S/W_S and P_D/W_D are sufficiently small in comparison to N_O. We require

$$(b_D/W_D)\ (P_S + 3P_T) \le 0.2\ B\ N_0 b_D \quad , \tag{4}$$

$$(b_S/W_S)\ (P_D + 3P_T) \le 0.2\ N_0 b_S \quad , \tag{5}$$

$$P_S/W_S \le 0.1\ N_0 \quad , \tag{6}$$

and,

$$P_D/W_D \le 0.1\ N_0 \quad . \tag{7}$$

If these criteria can be met, the multiple-access interference degradation can be kept below 0.8 dB.

The feasibility of meeting the criteria of Eq. (4) through (7) can be established by example. Let the nominal bandwidths for the sounding and tone detectors be 200 Hz. Assume that the data modem may desire to operate at a bit rate as high as 50 kbits/s. Further assume that both the sounding and the data (spread spectrum) signals occupy the full available RF bandwidth of 5 MHz and that all detectors desire a 20-dB SNR. If we normalize our design so that $P_S = P_T = 1$, then $P_D = 25$. Clearly, the data power represents the most significant interference power. In fact, it may be shown that it is necessary to increase the sounding and tone signal powers to 2 to maintain a 20-dB SNR for these signals. This increase has a negligible system impact because the data modem has adequate processing gain to render these signals insignificant, and because the total signal power, $P_D + P_S + 3P_T$, is only slightly affected by P_S and P_T when P_D is large. When P_D is comparable to P_S and P_T, there is no need to increase these latter two powers to combat the multiple-access interference.

Higher data rates could be accommodated by giving the data signal proportionally more power. This would require a proportional increase in the power of the sounding and tone signals, which, again, would have little system impact.

E. Loop Bandwidths

Tracking loop bandwidths must be adequate to track the channel-induced distortions. These distortions may be due either to the disturbed media or to the transponder position variations. Two types of loop bandwidths are involved--carrier tracking bandwidths and code tracking bandwidths. We expect that the code tracking bandwidths will be significantly narrower than the carrier tracking bandwidths. The latter must respond to small changes in the carrier phase, which can change an order of magnitude (or more) more readily than a code phase if the percentage bandwidth is 10 or less.

Selection of appropriate closed-loop bandwidths for these various loops is a particularly sensitive design issue for loops encountering additive noise, such as the transponder and ground reference receiver phase-lock loop. In such cases optimizing the closed-loop bandwidth is important. Bandwidths that are too narrow will provide inadequate transient response, while broad bandwidths will accept too much noise power. Prior experience has indicated that the bandwidths of phase-lock loops designed for tracking the sounding tones should be approximately 200 Hz. The bandwidths of the other phase-lock loops should be considerably broader to handle Doppler and frequency stability uncertainties. In fact, adaptive bandwidths may be required to provide rapid frequency acquisition times yet adequate steady-state phase noise.

F. Antenna Pointing

The antenna pointing requirements, which assume manual or preprogrammed tracking, are a function of the transponder trajectory. Rapidly moving (angularly) transponders are more difficult to track, and broader antenna beams may be required to handle this case.

Nominally we expect an antenna 3-dB beamwidth of 1° or 2° to represent a reasonable lower limit for the beamwidth with the expected trajectories. Use of narrower ground transmitter beams could reduce the required ground transmitter power but could necessitate the use of expensive automated tracking techniques such as monopulse. Larger ground receiver antennas could reduce the transponder power requirement, but again would require expensive automatic angle tracking. For rough preliminary design purposes, one may assume that the same ground antenna size is used for transmission and reception, because a monostatic experimental setup may be used in many situations. We have conservatively assumed 8.4° 3-dB beamwidths in our nominal link computations for the uplink and downlink. This means that the tracking problem is not severe. Alternatively, we could double or quadruple the antenna sizes and lower the ground and transponder transmitter powers. We have not selected this alternative because we believe the former approach is operationally more attractive.

The antenna gains possible on the transponder package are minimal because steering is not economically feasible. The vehicle housing the transponder will probably be spin-stabilized. Nominally, we expect gains in the range of 0 to 6 dB with capture areas related to the frequency of operation.

IV SUBSYSTEM DESIGN

In this section we present the detailed designs for the three major subsystems: (1) transmitter, (2) transponder, and (3) receiver.

A. Overall Transmitter System

Figure 3 is a block diagram of the overall transmitter system. The operator may enter mission-peculiar information for the microprocessor through a CRT terminal. The microprocessor will then point the 1-ft parabolic dish in the proper direction and bias the crystal VCO to the correct frequency range for the mission (and possibly in response to the measured frequency of the transponder). The crystal VCO produces the time base, which drives the coherent synthesizer.

The coherent synthesizer generates seven frequencies for transmission, one frequency for coherent up-conversion, and two code chipping rates. The seven transmission frequencies are two sets of three tones plus the uplink carrier frequency. By making all these frequencies coherent, the receiving system can choose to make as much use of the coherency as desired. Normally one would use the coherency within each of the bands independently, but it is possible to make use of interband coherency. For example, one could measure the relative phase shifts over the frequency range covered by both bands. If the channel disturbance is not too great, a satisfactory representation may be obtained in spite of the "unsounded" region between the two bands.

Separate spread-spectrum modulators are required for each of the two bands. The outputs from these modulators, which are on carriers generated by the synthesizer, are linearly combined and up-converted with a coherent reference signal. The lower sidebands (and carriers) are filtered, amplified, and transmitted. A 700-watt power amplifier (TWTA) is adequate to provide the required uplink power.

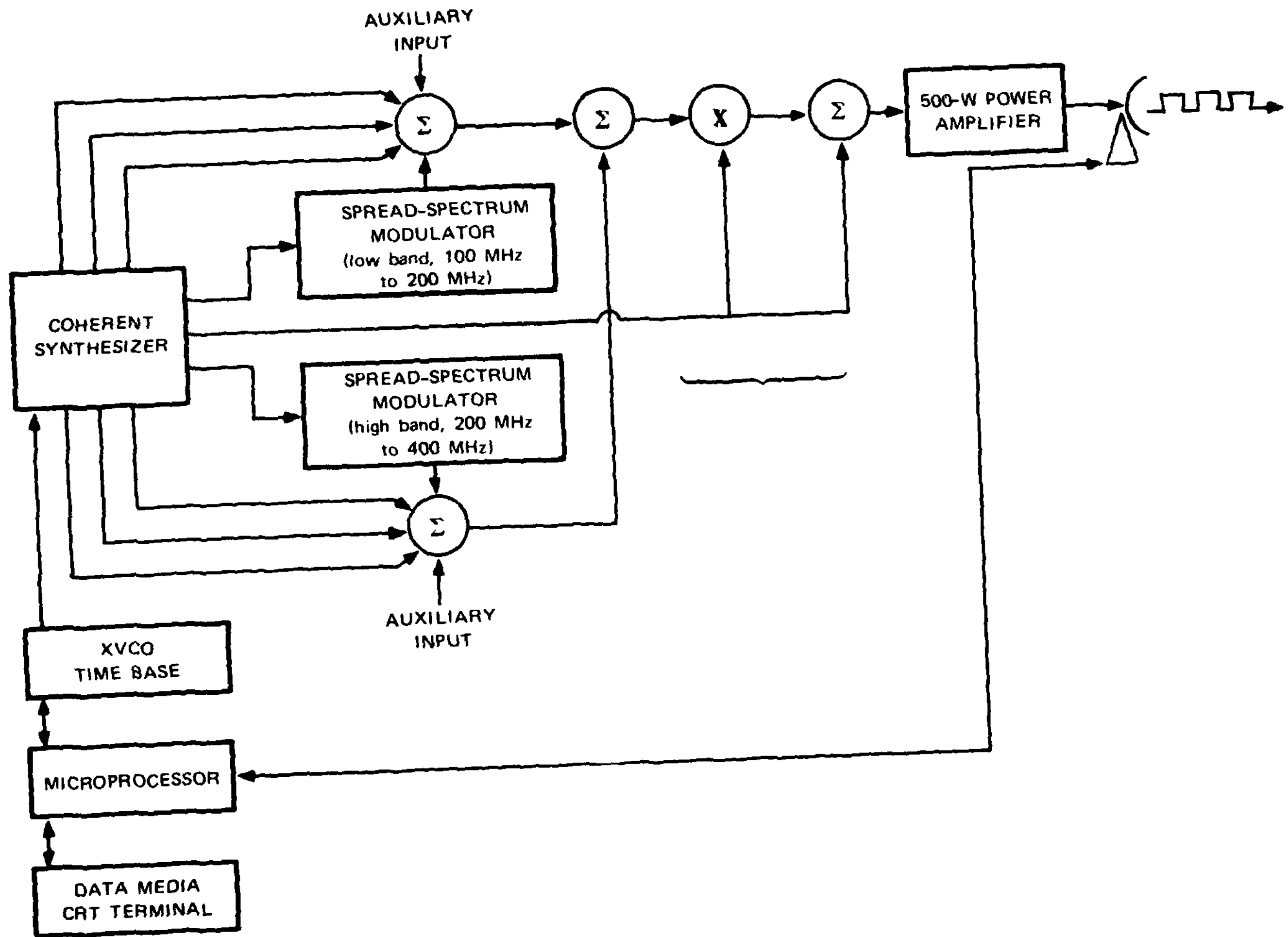


FIGURE 3 OVERALL TRANSMITTER BLOCK DIAGRAM OF *SPREAD-SPECTRUM* TECHNIQUES PROPAGATION EXPERIMENTS

A 1-ft, 8-GHz parabolic dish is required to provide adequate uplink power. This dish is controlled by the microprocessor.

1. Spread-Spectrum Signals

a. PN Spread Spectrum

The PN code structure selected is dependent on the receiver design and the maximum rocket speed, which we assume to be no more than 8,000 km/hr. For illustrative purposes we assume that the correlation receiver will measure the impulse response over 30 chips of delay with two taps per chip in both the in-phase (I) and quadrature (Q) channels. As a result, in any channel measurement interval, T_m, 120 data points are gathered.

Table 3 lists important parameters--i.e., first-null-to-first-null bandwidth, W, code length, T_p, channel measurement interval, T_m, and the maximum distance, D, that the rocket can move during the measurement interval as a function of chip rate and code length. One desires this transponder movement expressed in nanoseconds of round-trip delay to be substantially less than one chip duration for accurate measurement. Note that we assume that the correlation receiver is time-shared, gathering four data points as described above in one code period. The correlator is then delayed by one chip and the process repeated. As a result the channel measurement interval corresponds to 30 code periods.

Note that with a high chipping rate of 20 MHz and a long code of 1023 the effective motion of the rocket corresponds to a 22-ns shift for the case of a chip duration of 50 ns. For this high chipping rate a code length of 255 is much less likely to produce misleading results.

For channel sounding purposes there is no need to use quadriphase modulation. Consequently, we propose implementing a biphase PN system based on maximal-length (M) sequences. The chip rate will be variable at rates of 5, 10, 20, and 40 MHz, and the code lengths will be selectable between 2047, 1023, 511, 255, and 127.

Table 3

MEASUREMENT TIME AND OTHER PARAMETERS AS A FUNCTION OF CODE RATE AND CODE LENGTH

Code Length, L (No. of chips)	Code Rate, R (megachips/s)	Bandwidth, W (MHz)	Code Length, T_p (ms)	Measurement Period, T_m (ms)	Distance, D (m)
1023	2.5	5	0.4	12	26.7
	10	20	0.1	3	6.7
	20	40	0.05	1.5	3.3
511	2.5	5	0.2	6	13.4
	10	20	0.05	1.5	3.32
	20	40	0.025	0.75	1.68
255	2.5	5	0.1	3	6.71
	10	20	0.025	0.75	1.68
	20	40	0.013	0.37	0.85
127	2.5	5	0.05	1.5	3.35
	10	20	0.013	0.37	8.53
	20	40	0.006	0.18	0.40

b. Frequency-Hopping Spread Spectrum

Spread-spectrum signaling can be accomplished through the use of frequency-hopping (FH). However, as normally implemented, coherence between the hopping frequencies is not exploited. As a result, FH spread spectrum has inferior channel sounding properties (by several orders of magnitude) as compared to phase-chipping (direct sequence) spread spectrum. While in principle a coherent FH system with equal resolution could be constructed, its complexity is much greater than that of the direct sequence approach which we recommend and propose to employ.

An FH spread-spectrum modem can be used at the auxiliary ports of the transmitter and receiver. In this way the channel may be sounded by the direct sequence signal while the errors of the FH modem are recorded.

2. Tones

The method of generating the required tones for a two-band system is shown in Figure 4. Since three tones with equal spacing are required in each band, this arrangement can conveniently be realized through the use of amplitude modulators. Note that some selective filtering may be required to equalize the amplitude of each of the three components. Six line components are created by the coherent synthesizer and then shifted to the desired RF uplink frequency by mixing with another coherently related frequency. Note the requirement for five integer harmonic multipliers and one frequency divider.

The integers n_1, n_2, n_3, n_4, n_5, and n_6 must be selected to provide the desired frequencies illustrated in Figure 2. Note that the uplink reference signal (carrier) is not only used for the up-conversion but also added in as shown in Figure 4. If the simplified transponder design (see Section IV-B-3) is used, the uplink reference signal need not be present and, therefore, does not have to be added to the coherent synthesizer output.

The spread-spectrum signals (low and high band) are added to the tones in the appropriate bands. These signals are then processed by

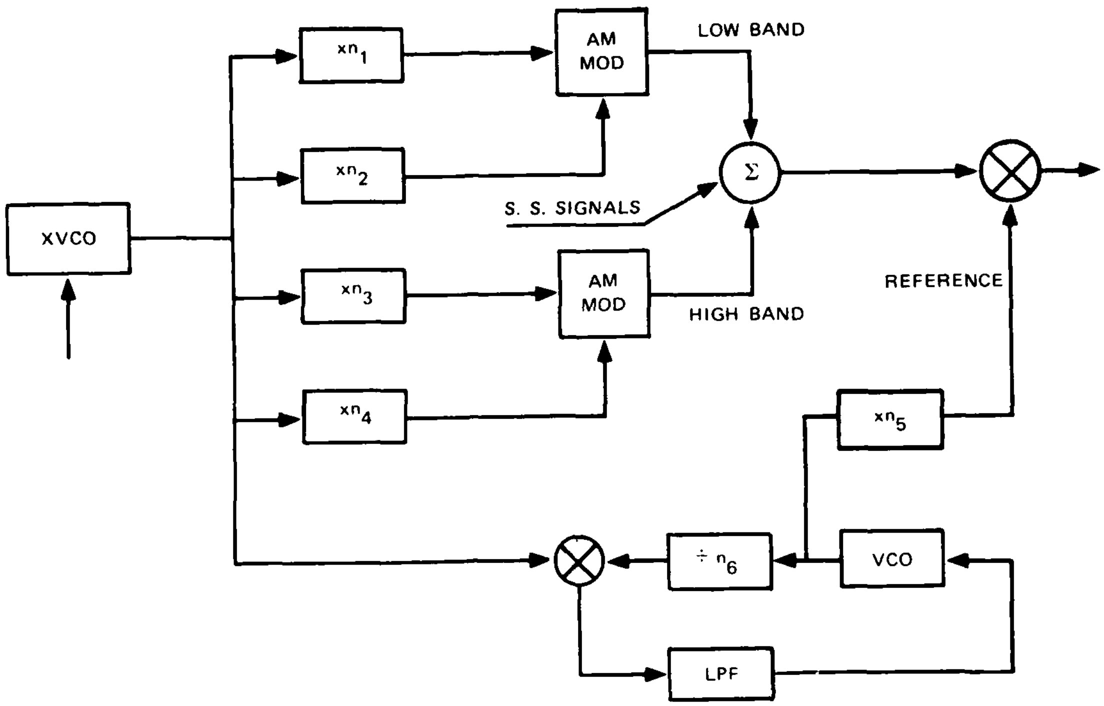


FIGURE 4 BLOCK DIAGRAM OF COHERENT SYNTHESIZER

the coherent synthesizer because they are up-converted by the same reference signal as used by the tones. Further discussion of this process is presented in Section IV-A-4.

3. Transmitter Peak Power Requirement

Since the ground transmitter must transmit three different signals, the peak power requirement is higher than the average power requirement. Taking the uplink example of Table 1, the peak power requirement is 676 watts or 4.4 dB higher than the 243 watts of average power.* The exact character of the transmitter input/output characteristic must be known to properly evaluate the IM cross products even for signals falling below the transmitter peak power limits.

*Here we assume that each band contains one constant envelope signal. With multiple signals, e.g., three tones plus a sounding signal, per band the peak power requirement is higher and it is not practical to meet it. Instead, some nonlinear distortion must be tolerated.

4. Choice of IF

The block diagram of the proposed transmitter and ground receiver assume that the IF used to transmit and detect the sounding tones is the same center frequency as the center frequency of the downlink bands. Thus, if a frequency band between 83 and 88 MHz is selected for sounding, the center frequency would be selected to be 85.5 MHz. Note that this is not a common or standard IF frequency. Consequently, if a military spread-spectrum modem were connected to the auxiliary input port this signal would not be located symmetrically about the frequencies being measured by tones and the direct sequence sounding signal. In general, it is desirable to symmetrically locate the tones and PN sounding signals about the data modem center frequency; however, it may not be necessary to do so.

There are two solutions to the problem of achieving symmetry in the frequency domain. First, the tones and the direct sequence signals (for both bands) could be generated at 70 MHz and then translated to the desired operating band. Alternatively, one could leave these signals at the downlink sounding frequencies and translate the frequency (IF) of the military modem to lie in the desired band. We favor the latter approach because it appears to be simpler, and in many, if not most, situations an external modem may not be used.

There is one difficulty with the proposed approach. The translation in frequency may be less than the channel bandwidth. In this case filtering cannot be used to separate upper and lower sidebands that occur in the frequency translation process. Consequently, the phase shift method of frequency translation illustrated in Figure 5 must be employed. While this approach to frequency conversion is more complex than the standard filtering approach, it does avoid the need for complex filters.

We propose to use this method only if a slight frequency translation is required. Implementation of broadband phase shifters is a non-trivial problem that is best avoided. Consequently, we recommend two approaches. First, if possible, we recommend that the lower sounding

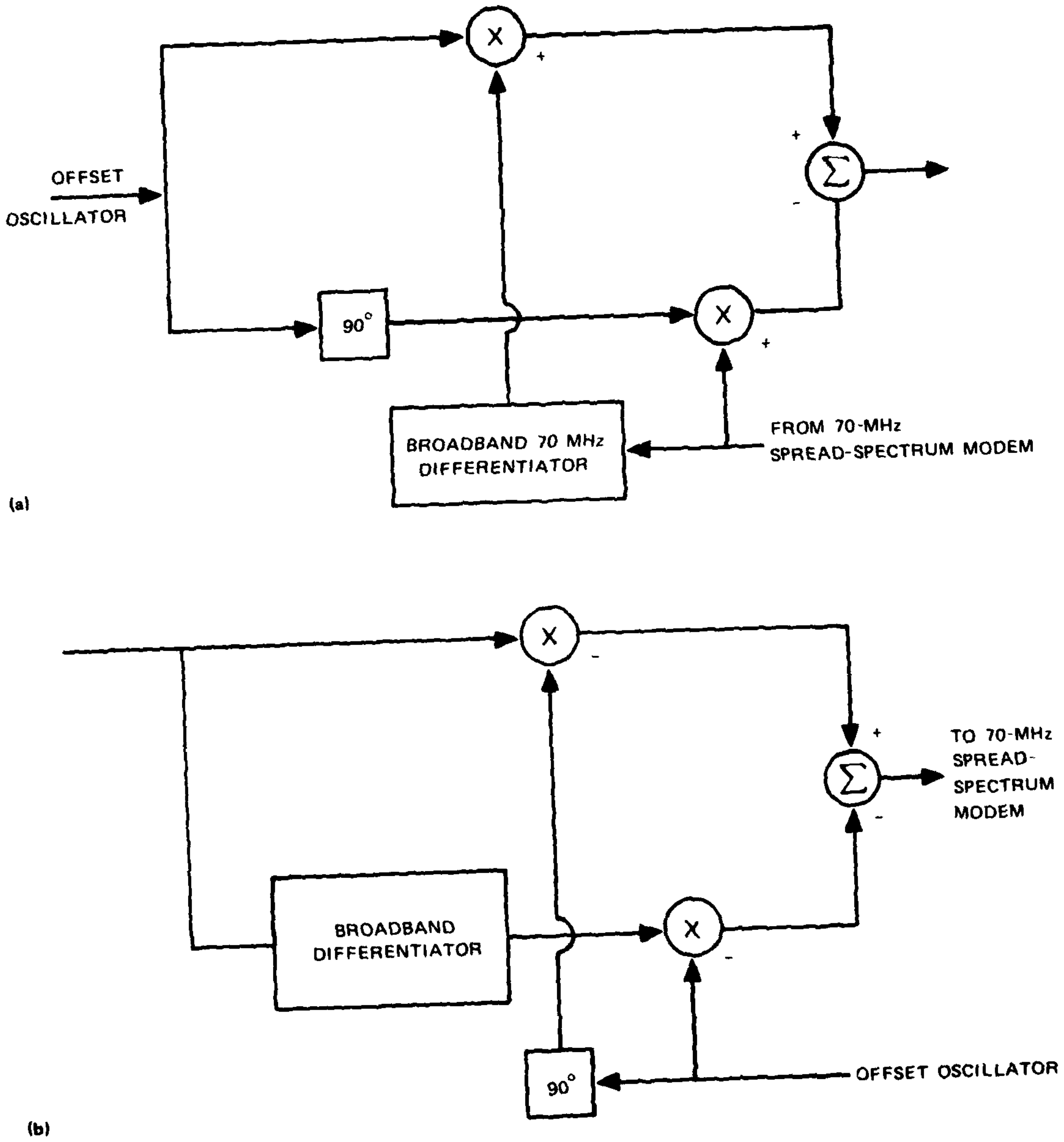


FIGURE 5 BLOCK DIAGRAMS OF THE PHASE SHIFT TYPE OF (a) UP-CONVERTER, AND (b) DOWN-CONVERTER

band be symmetrically placed about 70 MHz. If this cannot be done, then the lower sounding band should be located more than twice its bandwidth above 70 MHz. Since the code chipping rate can be reduced to 2.5 megachips/s, the lower sounding band may be centered as low as 85 MHz while using the conventional filter type frequency translators based on linear filtering to remove the undesired sideband.

B. Transponder Design

During the course of the project several transponder designs were studied. Rather than present each one of them in detail, we will present two rather different approaches in some detail and also discuss some of the fundamental design choices.

The gain control could be accomplished by one of two methods: automatic gain control or hard-limiting. The latter is advantageous if one expects jamming or significant interference. In the expected mode of operation neither of these situations should arise. We have chosen to use AGC to avoid the intermodulation (IM) cross products that would be created in the limiter approach. Due to the coherent multicarrier nature of the uplink signal, these IM cross products could seriously degrade the accuracy of the phase measurements on some of the tones.

A coherent AGC derived from the transponder phase-locked loop is recommended. A noncoherent AGC is undesirable due to the difficulty of measuring power in the entire IF passband, which may be as great as 1.5 GHz. If a noncoherent AGC were to be used, it would be necessary to use bandpass filters to measure the power in one or possibly all of the bands. This requirement would increase the weight, size, and cost of the transponder with respect to the coherent AGC design.

1. Transponder Alternatives

The major choice is what source provides the carrier basic frequency. SRI's initial thoughts were that this frequency would be provided by the ground transmitter. Later we considered an alternative design in which the transponder package set the downlink carrier frequencies. We believe that this alternative approach significantly simplifies the transponder while not significantly affecting the overall performance. Consequently, we recommend the alternative transponder design of Section IV-B-3.

2. Primary Transponder Design

Figure 6 is a conceptual block diagram of the primary transponder design. The signal is received, amplified with a low-noise amplifier, and coherently down-converted to an IF of approximately 1.6 GHz--i.e., the sidebands lie between 0.1 and 1.6 GHz. After IF amplification and gain control an 0.8 GHz VCO is doubled and phase-locked to the uplink carrier (now at 1.6 GHz). At the phase detector output the spectrum is inverted but the desired sidebands still lie in the 0-to-1.5-GHz region. The two bands are split by filters and linearly amplified separately to the desired power levels for downlink transmission.

The coherent down-conversion is achieved through the use of a phase-locked multiplier chain. The output of the 800-MHz VCO is divided by N and provides a reference for the multiplier PLL. The multiplier PLL VCO, which lies in the 900-to-970-MHz range, depending on the integers selected for N and M, is divided by M and phase-detected with the Nth subharmonic of the 800-MHz VCO. The multiplier VCO output is split and one line is multiplied by 7 to provide the reference signal, in the 6.3-to-6.8-GHz region, for the first down-conversion stage. The other output of the splitter is multiplied by 5 to provide the C-band downlink reference signal or beacon, which lies in the 4.5-to-4.86-GHz frequency range.

The portion of the transponder or coherent translator shown enclosed within the outer dashed line box would be common for all applications except that different integers might be employed to realize a particular frequency plan. Different applications would be accommodated by different filters, power amplifiers, and antennas. Thus, the most complex subsystem is common to all transponders. Consequently, the proposed design is cost-effective for a variety of applications. For

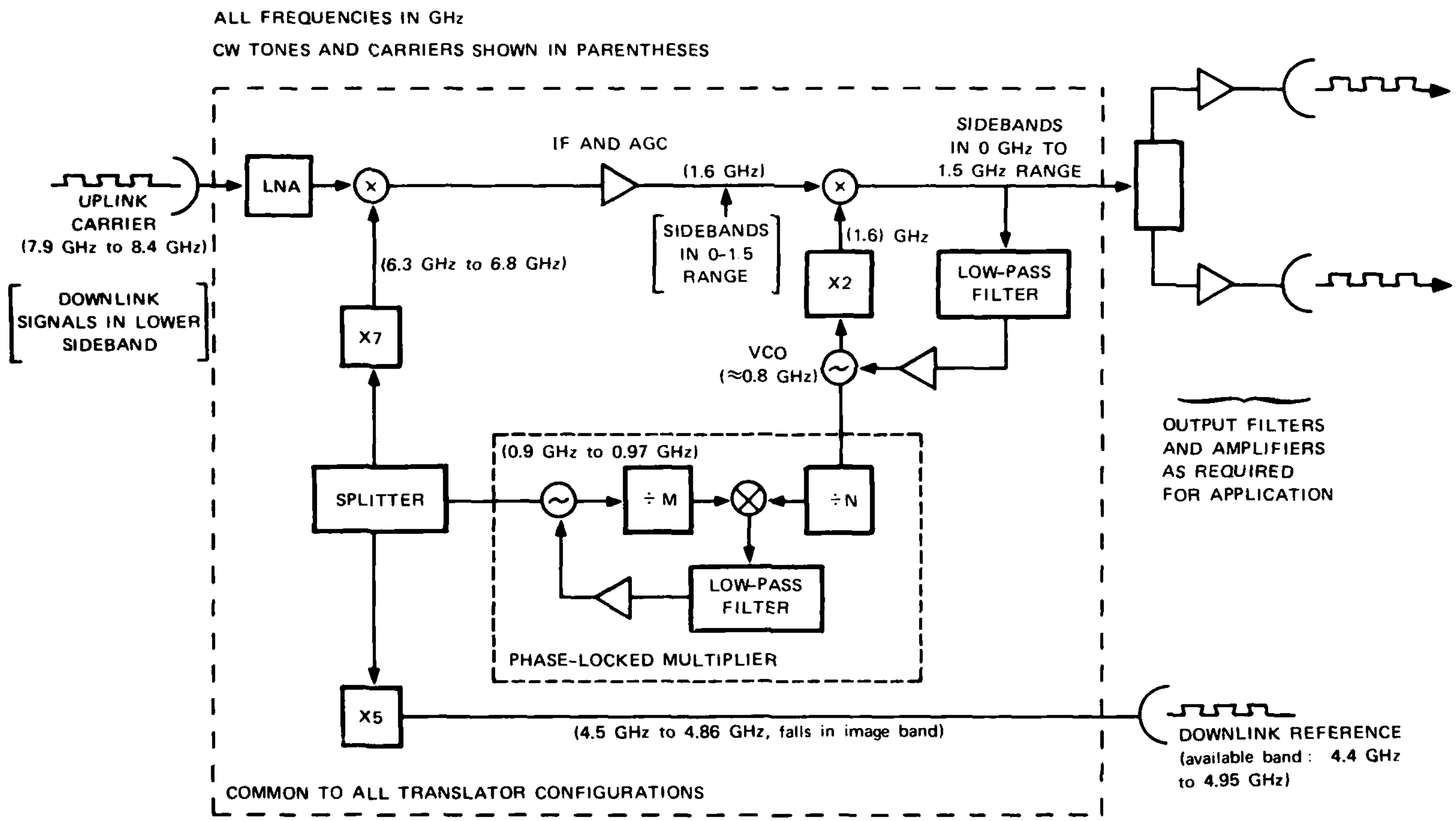


FIGURE 6 BLOCK DIAGRAM OF TRANSPONDER COHERENT TRANSLATOR

example, a 3-band design could be handled by adding another filter, power amplifier, and, possibly, another antenna to the two sets already shown in the block diagram.

Table 4 illustrates the M and N values required for carrier frequencies of 7.9 and 8.4 GHz, respectively. Other values could be obtained by selecting other values for N and M. The principal constraint is that very large values of N and M are undesirable because the equipment may become overly complex. Integers below 100 are certainly feasible. Note that the VCO frequencies have been selected so that digitial divider chips are commercially available and a practical design results.

Table 4

COHERENT TRANSLATOR PARAMETERS

Carrier Frequency (GHz)	Tracking VCO Frequency (GHz)	N	M	Multiplying VCO Frequency (GHz)	IF Carrier Frequency (GHz)	C-Band Reference Frequency (GHz)
7.9	0.8	32	36	0.9	1.6	4.5
8.4	0.7976	32	39	0.9721	1.5952	4.86

3. Simplified Transponder Design

The transponder design presented above satisfies the measurement objectives but needs a reasonably complex multiplier chain to generate the required signals. Consequently, the cost of the transponder is higher than desired for an expendable item. (Note that in some configurations such as a satellite the transponder cost may not be crucial because the transponder may not be expendable.) This subsection describes an alternative design that requires a much simpler harmonic multiplier chain.

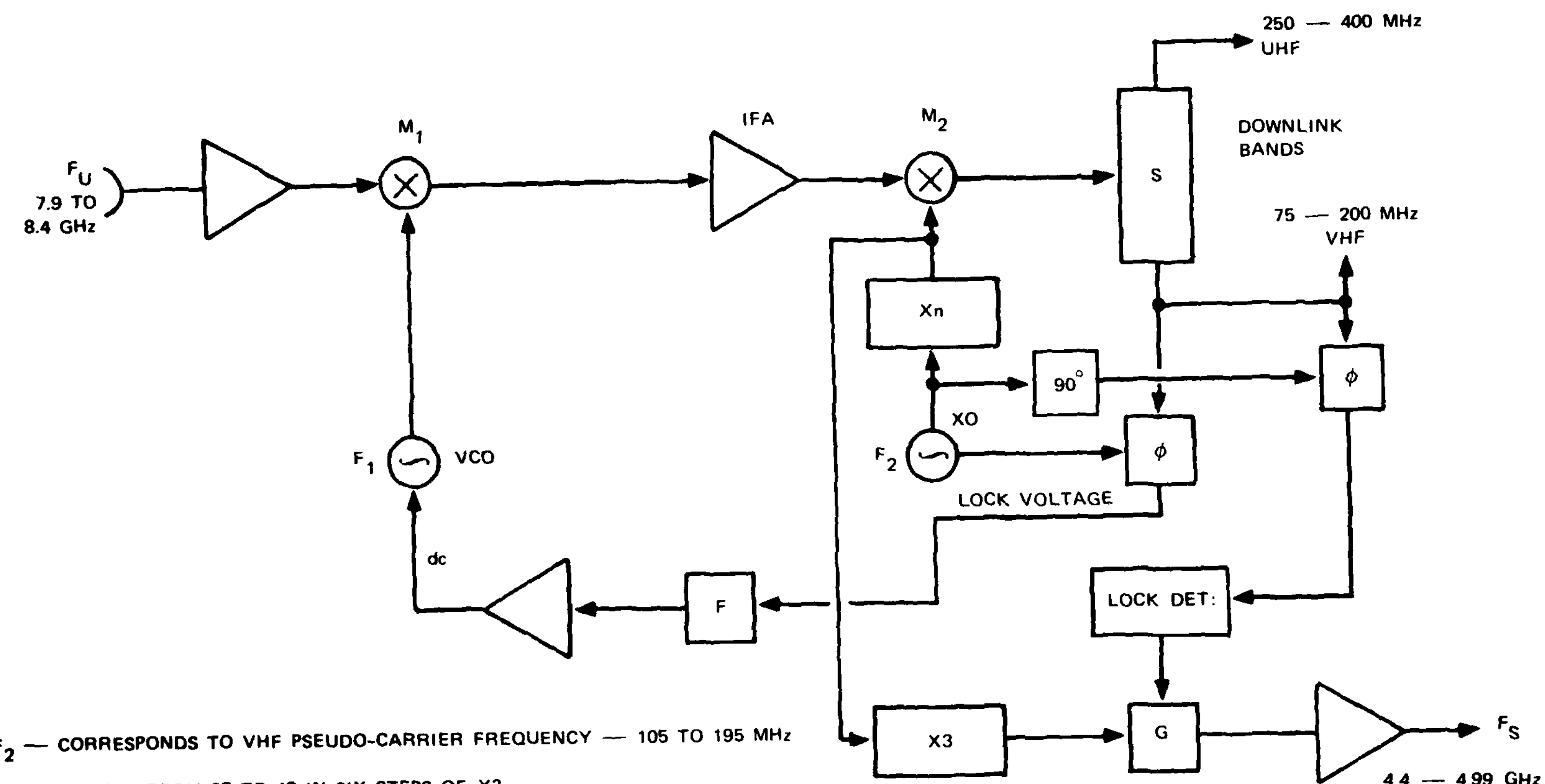
The simplified transponder design has the following system impact because the 8-GHz uplink carrier is no longer used as the reference. First, the system no longer permits the transponder "ranging" on the carrier phase because the downlink reference signal does not contain any uplink Doppler information. Transponder "ranging" can still be performed, but it must be performed on the modulation, and the accuracy will be limited to that defined by the PN code rate, which should be sufficiently good for most purposes. Alternatively, depending on the launch site, transponder tracking may be provided by an auxiliary radar.

Since the downlink reference signal is not shifted by the uplink Doppler frequency, the frequency search problem at the ground receiver is substantially simplified. Furthermore, the frequency assignment problem for the beacon may be eased because the reference is more stable with the simplified transponder.

In addition, with the simplified design, one less uplink signal is required. The carrier component at frequency f_i is not required. The simplified transponder phase locks on the nearest or highest frequency tone. The system impact is that slightly less uplink power is required and the linearity requirement on the ground transmitter is somewhat reduced. Also, the signal generation stage is slightly less complex.

A simplified version of the Beacon Transponder has been devised and is shown in Figure 7. A stable crystal oscillator, F_2, provides an on-board reference frequency locked to the VHF carrier. F_2 is modular, plug-in, and has the same frequency as the VHF carrier used in the experiment. A multiplier (x N) provides the second LO frequency from F_2 and is further multiplied by 8 to provide the C-band downlink. The C-band downlink is thus locked to the VHF carrier frequency. The first LO frequency is provided by a separate VCO and is phase locked to F_2 to provide the necessary down-conversion to offset drift, frequency errors, and Doppler shift in the uplink signals. The IF center frequency for all cases is a third of the C-Band downlink frequency F_S and lies in the range 1.4667 to 1.663 GHz corresponding to NF_2.

By providing six plug-in multipliers in the range X9 to X14 for multiplier N, suitable combinations of C-Band downlink carriers in the range of 4.4 to 4.99 GHz may be obtained using an F_2 frequency between 104 and 185 MHz corresponding to the VHF carrier.



F_2 — CORRESPONDS TO VHF PSEUDO-CARRIER FREQUENCY — 105 TO 195 MHz

n — RANGES FROM 27 TO 42 IN SIX STEPS OF X3

FIGURE 7 BLOCK DIAGRAM OF SIMPLIFIED TRANSPONDER

The uplink carrier F_u can be selected independently of the F_2 frequency of the multiplier by using a VCO for F_1 having a range of 6.3 to 7 GHz.

To prevent an unlocked C-band downlink signal occurring if the system loses lock, a lock detector gates "on" the C-Band transmission; thus, loss of lock is indicated by loss of C-Band downlink reference signal. The frequency relationships for the system are given in Table 5.

Some minor changes are required in the basic receiving system design. A block diagram of the receiver RF and IF system is shown in Figure 8. In this diagram the baseband signal is split after the second mixer, M_2, to provide a narrowband channel to lock the crystal oscillator to the VHF signal (Figure 9). The second channel contains the AGC amplifier to maintain constant signal level drive to the VHF and UHF power amplifiers. An AGC range of 25 dB is thought to be sufficient. The AGC amplifier has been removed from the lock loop in order to prevent phase shift in the amplifier under AGC action from affecting the lock loop.

The multiplier chain is now considerably simplified, with reduction in costs expected.

C. Receiver System

The proposed receiver system is illustrated in Figure 10 for the case of a two-band configuration. An additional set of tone and spread-spectrum receivers would be required if the third band were desired.

The transponder reference in the 4.4-to-4.95-GHz band is received with a parabolic dish that can be pointed automatically in a preprogrammed fashion. A low-noise amplifier is used to amplify the received signal before application to the phase-locked receiver, which includes a down-converter. The phase-locked receiver can be programmed by the microprocessor to be offset to a frequency appropriate to the known trajectory and possibly to oscillator offsets measured prior to transponder launch.

The output of the phase-locked loop is used to drive a coherent synthesizer. This synthesizer generates the six (or possibly nine) reference tones required to measure the relative phase shifts or each of the

Table 5

REVISED SIMPLIFIED SYSTEM FREQUENCY PLAN

C-Band Multiplier	N	LO Frequency, F_2 (MHz)	N x LO Frequency, IF (MHz)	RF Input, F (GHz)	First LO Frequency, F (MHz)	Multipliers	C-Band F_s
27	3	184.8	1663.2	8.4	6.7368	3 x 32	4989.6
				7.9	6.2368		
		162.9	1466.1	8.4	6.9339		4398.3
				7.9	6.4339		
30	10	156.3	1663.0	8.4	6.737	3 x	
						(5 x 2)	4989.0
				7.9	6.237		
		146.7	1467.0	8.4	6.933		4401.0
				7.9	6.433		
33	11	151.2	1663.2	8.4	6.7368	3 x 11	4989.6
				7.9	6.2368		
		133.3	1466.3	8.4	6.9337		4398.9
				7.9	6.4337		
36	12	138.6	1663.2	8.4	6.7368	3 x	
						(3×2^2)	4989.6
				7.9	6.2368		
		122.2	1466.4	8.4	6.9336		4399.2
				7.9	6.4336		
39	13	127.9	1662.7	8.4	6.7373	3 x 13	4988.1
				7.9	6.2373		
		112.8	1466.4	8.4	6.9336		4399.2
				7.9	6.4336		
42	14	118.8	1663.2	8.4	6.7368	13 x	
						(7 x 2)	4989.6
				7.9	6.2368		
		104.7	1465.8	8.4	6.9342		4397.4
				7.9	6.4342		

Note: IF = RF - F_1. C-Band = 3N x F_2.

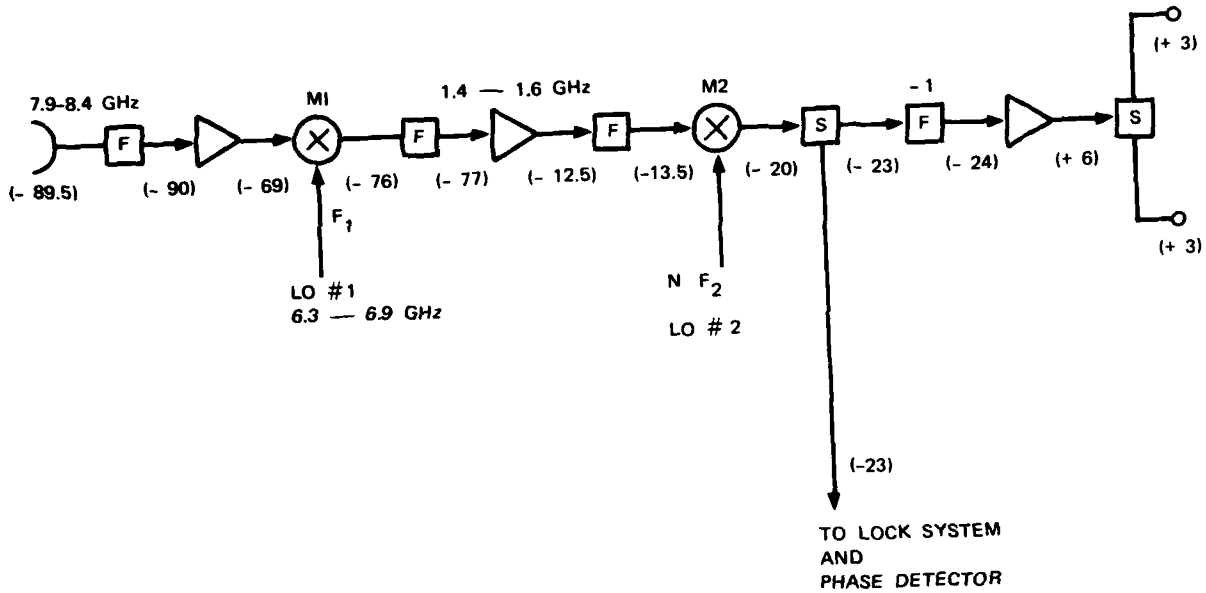


FIGURE 8 RF–IF AMPLIFIER SYSTEM. Numbers in parentheses are signal levels in dBm.

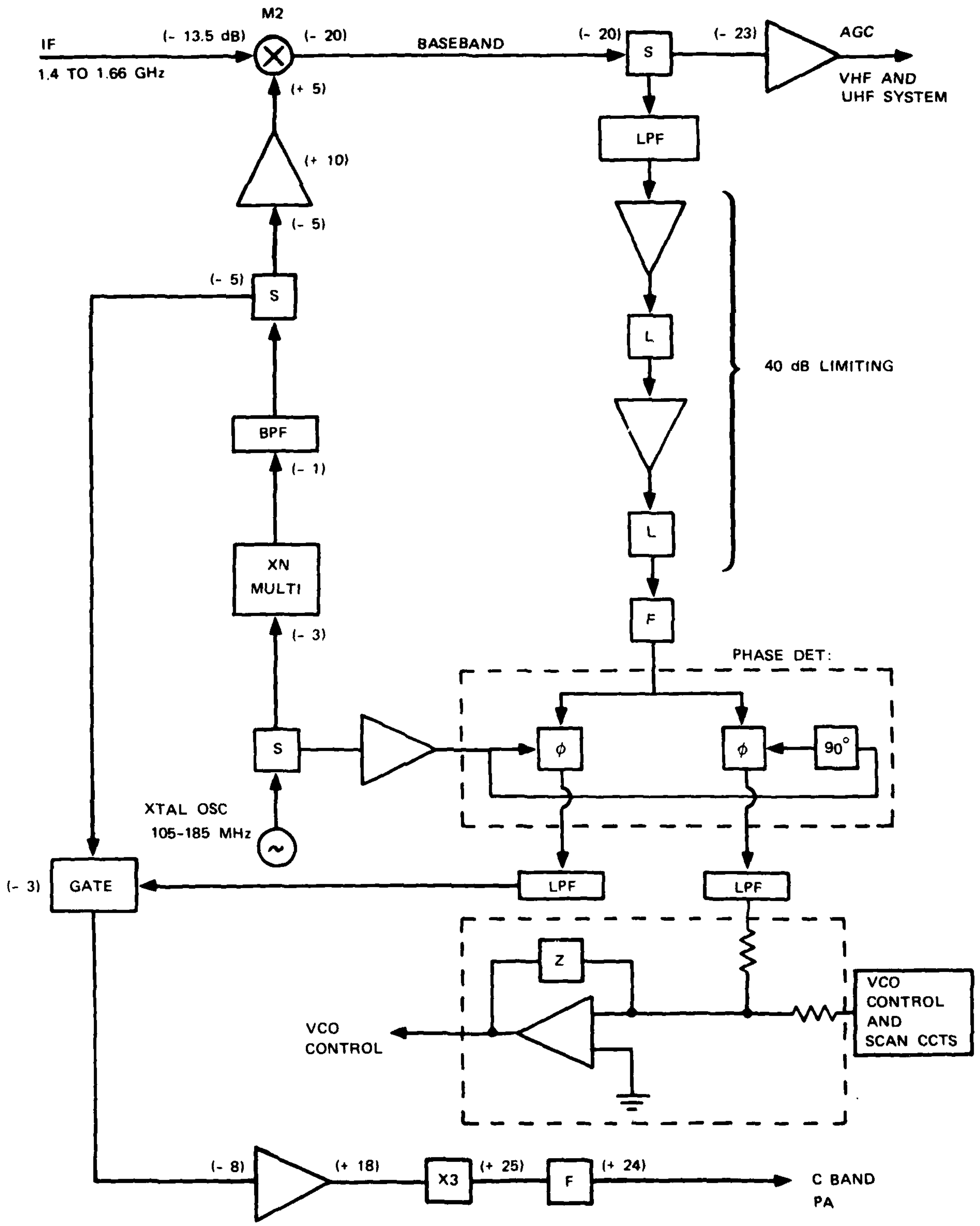


FIGURE 9 VCO AND CRYSTAL OSCILLATOR LOCK LOOP. NUMBERS IN PARENTHESES ARE SIGNAL LEVELS, IN dBm

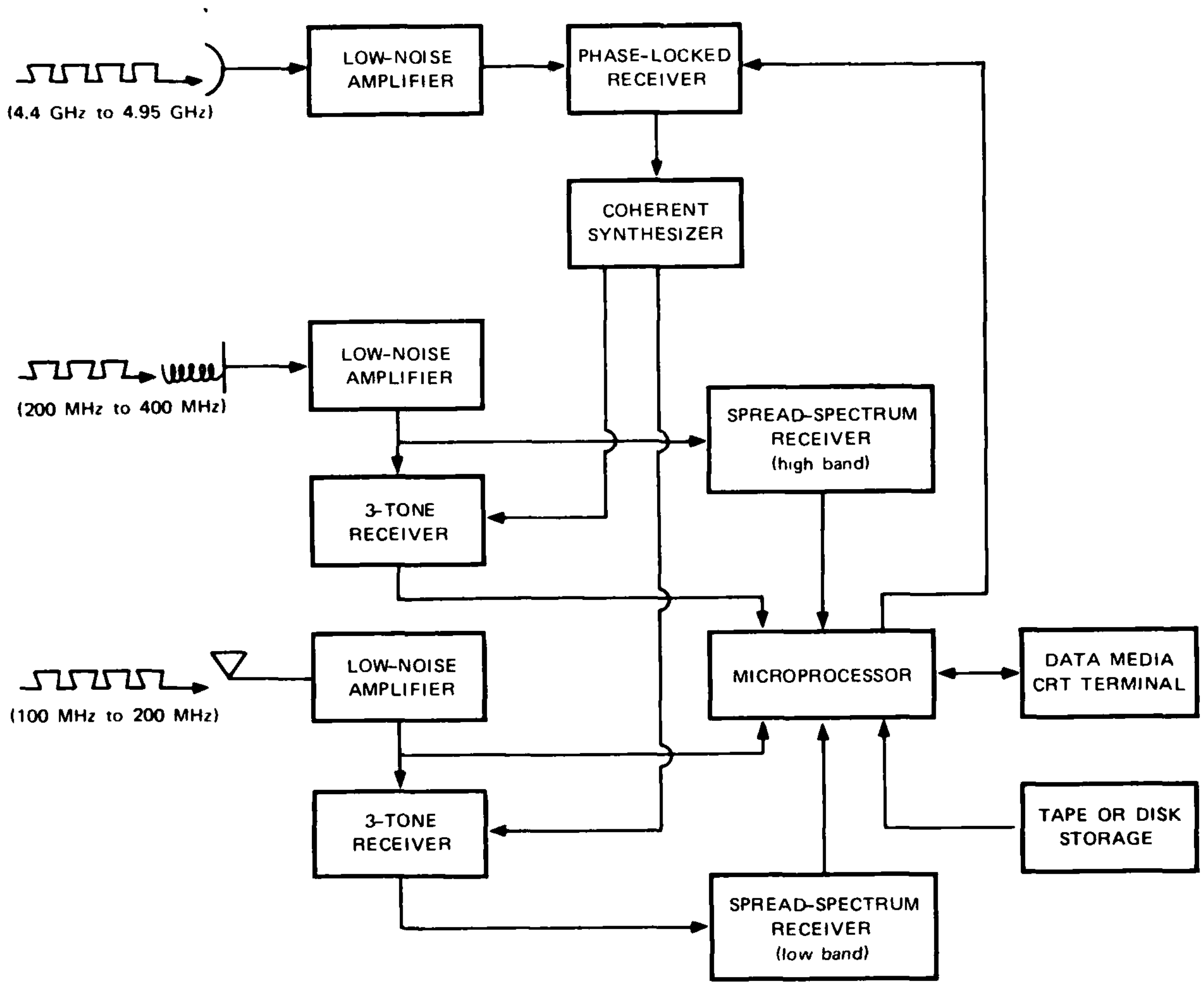


FIGURE 10 OVERALL RECEIVER BLOCK DIAGRAM OF SPREAD-SPECTRUM-TECHNIQUES PROPAGATION EXPERIMENTS

six (or possibly nine) sounding tones. The measured phase shifts are processed by the microprocessor system and stored on a disk system.

Besides supplying the received three tones for each of the two three-tone receivers, the low-noise amplifiers also provide the input signals for the two spread-spectrum receivers. The outputs of the spread-spectrum receivers (which are the channel impulse response) are processed by the microprocessor system and recorded in a disk.

The high band signals, which will be located somewhere in the 200-to-400-MHz region, will be received by a separate helical antenna. The low band signals will be received by an antenna, which may also be helical, depending on the frequency that is mounted on the same drive system. Thus, we propose a 2- or 3-antenna system controlled by a common antenna drive system. The antenna for the reference receiver will be a 5-GHz dish with an approximately 2-ft diameter.

The microprocessor system can be preprogrammed for the particular mission (trajectory, and so forth) through a CRT terminal.

1. Spread-Spectrum Signal Receiver

a. Tracking Section

The tracking portion of the spread-spectrum receiver is illustrated in Figure 10. A tracking system is desired to account for the motion of the transponder. As shown in Figure 10, an envelope (or bandpass) correlator delay-lock loop is used to track the main or strongest path. A search bias can be applied to the PN generator voltage-controlled clock to provide the required acquisition function.

The spread-spectrum receiver (one for each band) provides baseband outputs (I and Q channels) for the baseband cross-correlator or channel impulse response measurement circuit. In order to develop these baseband signals it is necessary to lock to the "carrier" of the spread-spectrum signals. A phase-lock loop operates on the punctual correlator output, which (assuming no data modulation of the spread-spectrum signal) has a carrier component. The phase-lock loop generates both I and Q references signals, which are used to translate the input signal to baseband.

b. Impulse Response Measurement System

Figure 11 is a block diagram of the impulse-response measurement system. It takes its input from the tracking section of the spread-spectrum receiver. It also accepts the code from the PN generator in the tracking section. The outputs are the real and imaginary portions of the channel impulse response. The imaginary portion occurs only when the channel introduces even-order phase distortion.

Note that four correlators are used, because the correlation is computed simultaneously at points separated by one-half chip duration. This spacing guarantees that, as the time base is shifted through the range of delays in one-chip increments, a sample is obtained within one-quarter chip of the maximum correlation. Assuming infinite bandwidth and an ideal M sequence, the maximum correlation loss would then be 2.5 dB. For a practical situation with limited bandwidth, the loss would be considerably less. Thus, half-chip spacing provides adequate resolution.

The selected range for computing the channel cross-correlation has been selected to be 30 chips. As Figure 12 shows, it takes 30 code periods to evaluate the cross correlation over the entire range. Equipment simplicity dictated that we take this approach rather than implement parallel correlators for all 30 time displacements.

The potential disadvantage with the sequential approach is that the cycle time might be too long and the channel impulse response would have changed by the time the cycle completed. Since we propose to use M sequences of several lengths, we have some protection against this problem. However, as the code is shortened the spread-spectrum processing gain against the tones and other interference will be reduced. Also, the longer the code period the greater the movement of the transponder package in the measurement interval--i.e., 30 code periods. Table 3 presents the code period (T_p) measurement period (T_m), and the distance, D,

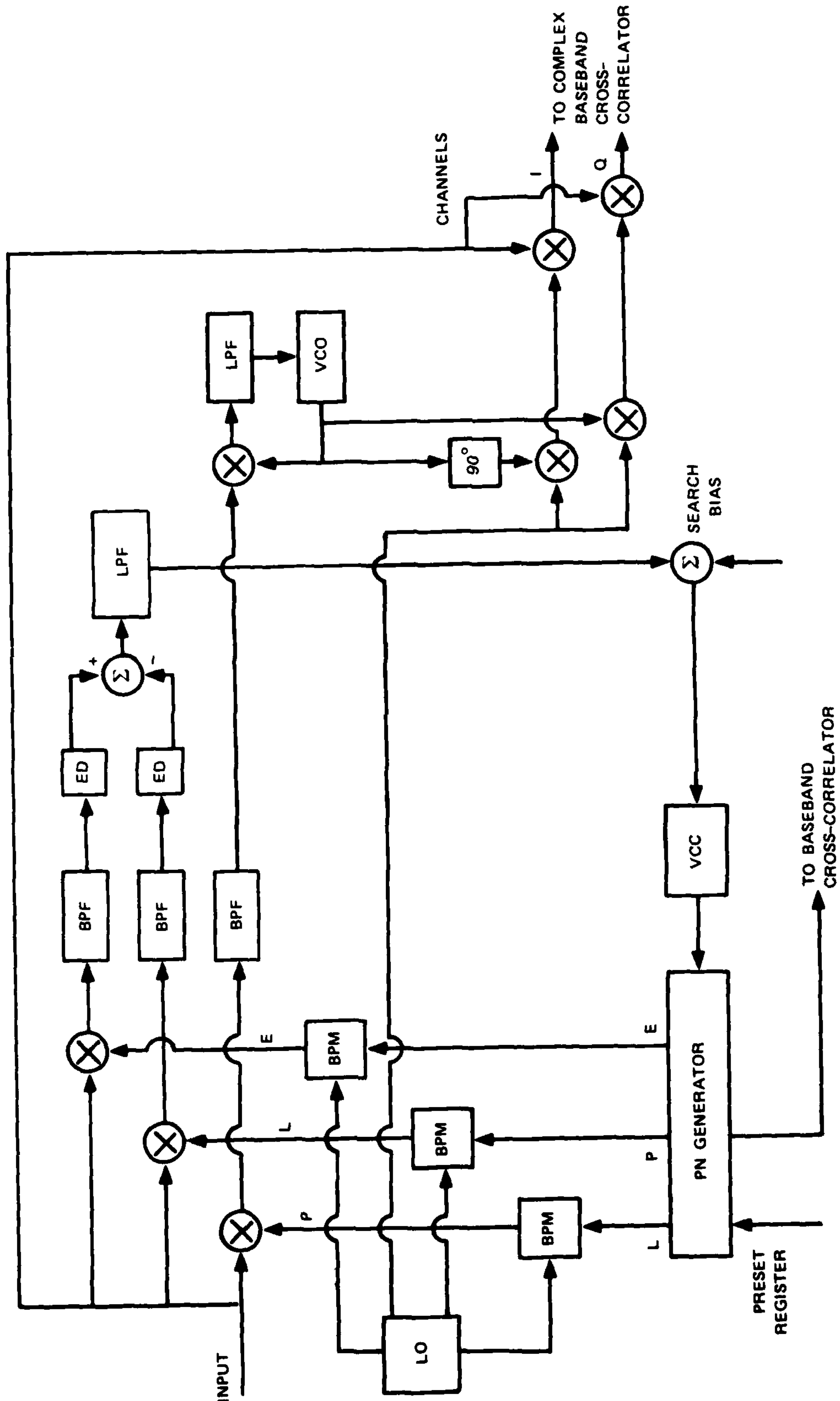


FIGURE 11 SPREAD-SPECTRUM RECEIVER

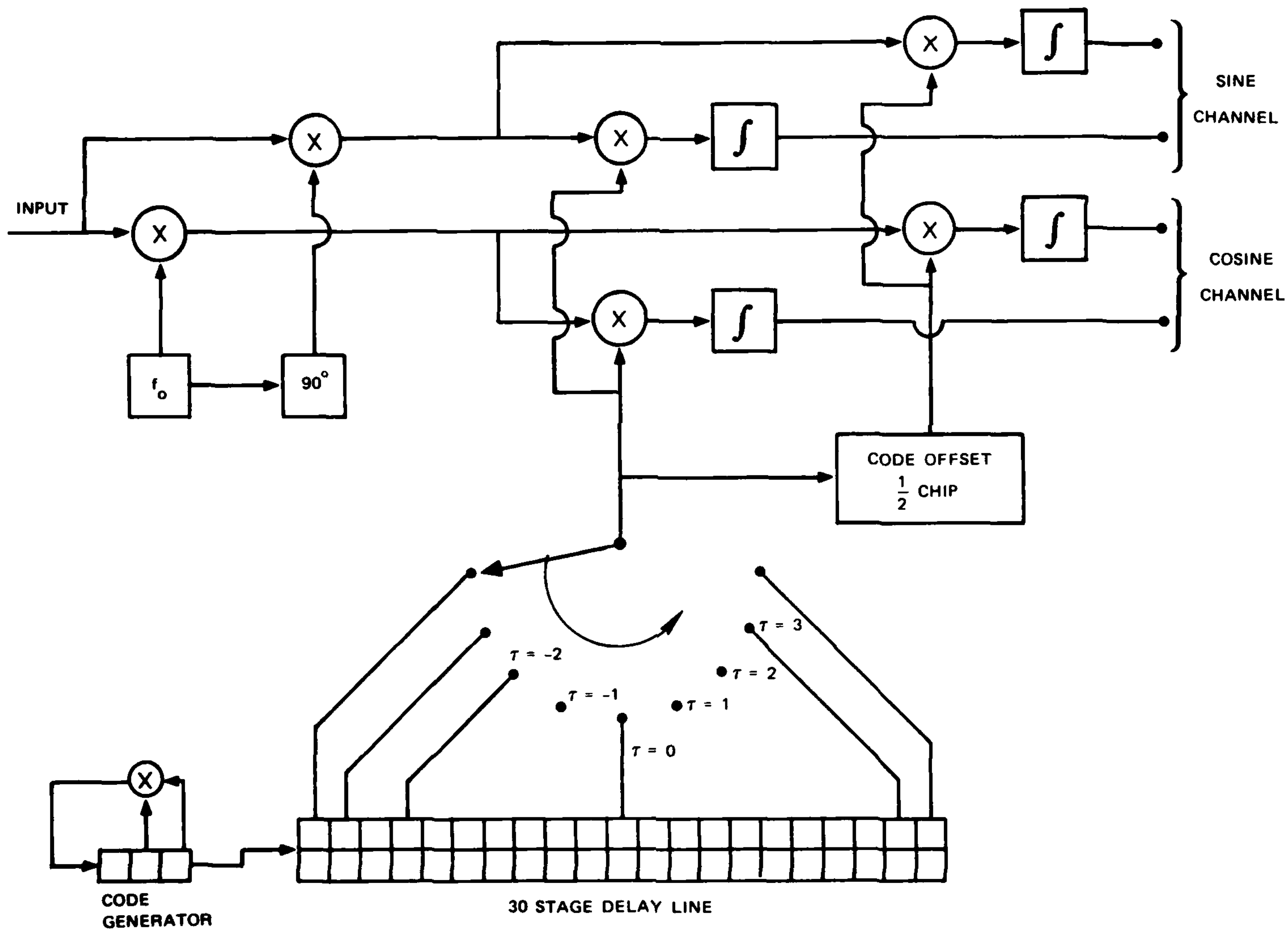


FIGURE 12 BLOCK DIAGRAM OF IMPULSE-RESPONSE MEASUREMENT SYSTEM

the transponder has moved (in feet) during the measurement interval as a function of the PN code chip rate and code length, L. Also presented is the first-null-to-first-null bandwidth of the PN signal. The distance the transponder has moved is based on an assumed velocity of approximately 8,000 km/hr. Given that the transponder is behind the disturbed medium, the radial velocity should not produce a significant effect other than the Doppler shift. Consequently, the most significant velocity is the angular velocity. We expect this to be considerably less than 8,000 km/hr, so the maximum transponder motion in a direction perpendicular to the radial will be no larger than the values cited in Table 3.

If one assumes that the channel changes at a 200-Hz rate, then samples must be taken at a rate of at least 400 Hz. Allowing for practical implementation factors, the sampling rate should be at least 500 Hz. In terms of Table 3, this means that the measurement period must be less than 4 ms. If a code rate of 2.5 megachips/s is selected, the code length must be less than 255. If a code rate of 10 megachips/s is selected, the code length must be less than 1023. If a code rate of 20 megachips/s is selected, the code length must be less than 2047.

The A/D converter system for the PN sounding system has 8-bit accuracy and must accept four new samples each M-sequence code period. Assuming a code period of 1023 chips and 10- and 20-MHz chipping rates for the low and high bands, respectively, the code periods are 102 and 51 μs, respectively. Thus, representative word transfer rates are 40 kHz for the low band and 80 kHz for the high band. We propose separate A/D converters for the two bands because this will give greater flexibility in system operation. For example, different code lengths in the two bands can be more readily accommodated if two A/D converters are used.

2. Tone Receivers

Figure 13 is a block diagram of the tone receiver for one band. There will be two such receivers, and possibly three, if a 3-band design if adopted. The frequencies will be different, of course. Different

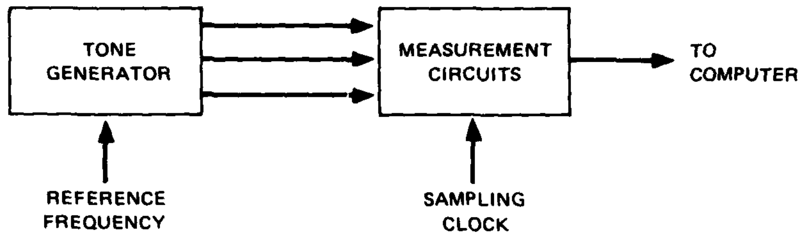


FIGURE 13 BLOCK DIAGRAM OF TONE RECEIVER

multiplication ratios will be required to generate the different frequencies. Other than this difference, the tone receivers will be identical.

The tone receiver operates on the reference or beacon phase-lock loop output to generate the desired signals in exactly the same fashion as the ground transmitter. That is, the received 5-GHz reference tone, which is assumed to be unaffected by the channel distortion, provides the basic frequency source for the receiver. This signal, after appropriate frequency scaling, drives a tone generator identical to that used in the transmitter (see Figure 4), except that the last up-conversion (to 8 GHz) is not required. These tones will then have the same relative phases as the transmitted tones and may be used as reference signals for the phase detectors. These measure the phase shift of the received tones with respect to the transmitted tones. Note that the receiver requires two doubly balanced mixers to implement each phase detector, and that the receiver tone generator must produce two signals, 90° shifted from each other, per tone, while the transmitter tone generator must produce only one signal per tone.

The in-phase and quadrature components are measured in a 200-Hz bandwidth by sampling at a 500-Hz rate as shown in Figure 14. These samples are then A/D converted and the phase shift determined in the computer using the relationship

$$\phi = \tan^{-1} (Q/I) \quad ,$$

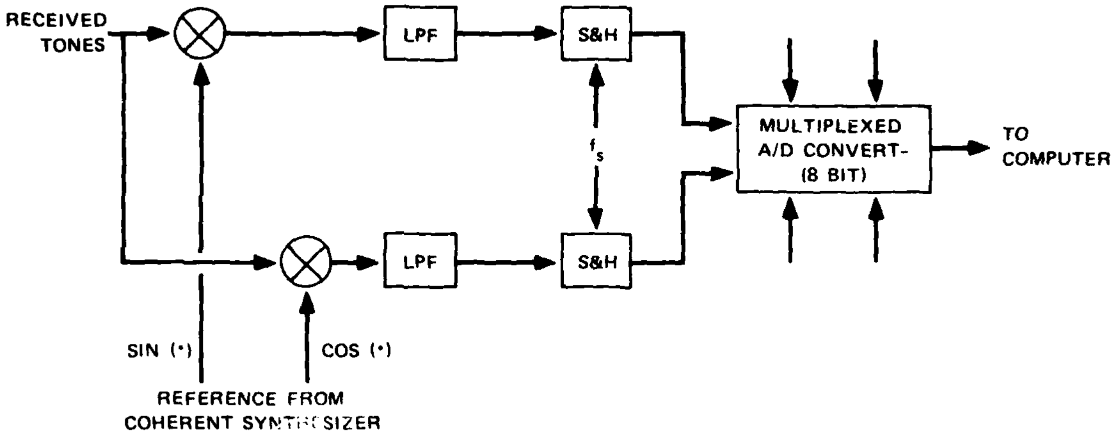


FIGURE 14 BLOCK DIAGRAM OF TONE-RECEIVER QUADRATURE DETECTOR

where Q represents the quadrature-channel measured sample and I the in-phase channel measured sample. By storing the six measured samples per band per 2-ms frame the computer can sequentially evaluate the three phase shifts per band. In fact, it is possible to share one multiplexed A/D converter between the two, or possibly three, bands. Note that an 8-bit accuracy is sufficient, and multiplexed A/D converters are readily available. The output word rate from the tone A/D converter is 6 kilowords/s for a two-band system and 9 kilowords/s for a three-band system.

The amplitude, A, of each tone is determined in the computer system according to

$$A = \sqrt{I^2 + Q^2} \quad .$$

Thus, the tone receivers measure both the phase and amplitudes of the tones.

3. Computer System

Two computer systems are required. The transmitter requires one for the miscellaneous functions described below. The receiver requires a larger computer system that performs similar miscellaneous functions plus data logging. If a monostatic configuration is selected, one computer system is sufficient. However, bistatic configurations require two computers.

a. Data Acquisition

The computer system must acquire and log--i.e., direct to data storage, such as disk--the measurements described below.

First, the computer must measure the relative phase shifts and amplitudes on each of the three tones per band. There will be at least two bands and possibly three. Note that each of these measurements involves an active computation to determine the phase angle and a square root of the sum of two squares to determine the amplitude. These computations must be performed at a rate fast enough to be compatible with a channel sampling rate of 500 Hz.

Second, the cross-correlation of the PN sounding channel must be recorded in both bands. This requires both I and Q channel measurements over a 30-chip range with 1/2-chip resolution. Thus, 120 different values must be recorded per measurement interval. The measurement interval is a function of code rate and code length but should be on the order of 2 ms or less if the bandwidth of the channel impulse response measurements is to be the same as that of the tone measurements.

Third, one may need to record the bit error pattern for a military modem operating in one or more of the bands. If one assumes that this modem is restricted to a data rate of 10 kilobits/s or less, then a 10-kilobits/s rate is adequate to store the bit error pattern. This pattern can be related at a later time to the channel impulse response.

Fourth, and finally, all the above data must be time tagged so that performance may be related to the transponder position and other physical phenomenon. It is the responsibility of the computer system to provide this information.

b. Miscellaneous Computer Functions

The computer system must translate trajectory information into antenna elevation and azimuth angles as a function of time. Thus, prior to launch, a code describing the intended trajectory must be entered. At the time of launch the operation(s) must enter the launch time. This information must be processed both at the transmitter and the receiver site, which potentially will have different algorithms because they have different locations.

In addition, it may be desirable to preprogram transmitter and receiver VCOs to account for the expected Doppler shift. By doing this the acquisition time at the transponder receiver phase-lock loop and the ground phase-lock loop may be reduced. Also, the trajectory information can be used to shift the ground receiver PN code generator time base to roughly the correct position so that tracking can commence without an extensive acquisition period.

V PERFORMANCE ANALYSIS

In this section we analyze the performance of the proposed experimental sounding system. We present results for the acquisition times for the phase-lock loops and the delay-lock loops. We also present results for the amplitude and phase errors associated with the tone measurement system.

A. Phase-Lock-Loop Acquisition Time

Two very important phase-lock-loop acquisition processes are required. First, an acquisition at the transponder is necessary. Second, an acquisition of the reference signal at the ground receiver is required. Table 6 summarizes the acquisition parameters and resulting times.

At the transponder the maximum Doppler shift is 96 kHz, but trajectory information may be used to reduce this uncertainty. Assuming a frequency stability of 10^{-5} (perhaps 10^{-5} from the transponder and 10^{-6} from the ground transmitter) the uncertainty from this source is approximately ±80 kHz. This value could be reduced by pre-launch measurements where the oscillator frequencies of both the ground transmitter and the transponder could be measured. However, it should be noted that the transponder oscillators might significantly change frequency after launch due to the mechanical shock associated with launch.

If one assumes the worst case of a total frequency uncertainty of 176 kHz and a closed-loop bandwidth of 50 kHz, the acquisition time for a conventional phase-lock loop is only 1 ms. This time is sufficiently short that there does not appear to be any very good reason to consider more sophisticated phase-lock loops.

On the downlink the carrier frequency is approximately one-half the uplink frequency. Consequently, the maximum Doppler shift is reduced to ±48 kHz and the frequency stability uncertainty is reduced to ±40 Hz.

Table 6

PHASE-LOCK-LOOP ACQUISITION TIMES

Transponder PLL		
Maximum Doppler shift	=	±96 kHz
Frequency instability	=	±80 kHz
Total Uncertainty	=	±176 kHz
PLL Bandwidth	=	50 kHz
Acquisition Time	=	1 ms
Ground receiver PLL		
Maximum Doppler shift	=	±48 kHz
Frequency instability	=	±40 kHz
Total uncertainty	=	±88 kHz
PLL bandwidth	=	10 kHz
Acquisition Time	=	9 ms

At 4 GHz the total frequency uncertainty is approximately ±88 kHz. Of course, this frequency uncertainty can be reduced by the same methods described above. If one assumes an acquisition bandwidth of 10 kHz (this value is feasible at a range of 60 km for the power levels described in Section III-C), the acquisition time is 9 ms. Since this value is very low, it is possible to narrow the bandwidth and achieve an acquisition time that is still acceptable.

For a phase-lock loop the frequency acquisition time is given by

$$T_f = 4(\Delta f)^2/B^3 \quad , \tag{8}$$

where Δf is the frequency offset and B is the closed-loop equivalent noise bandwidth.[4]

The total acquisition time consists of the frequency plus the phase acquisition time. The latter value is given by

$$T_{\phi} = 4/B \quad . \tag{9}$$

In our case the frequency uncertainty is sufficiently large that T is negligible in comparison to T_F. thus, our acquisition time is effectively given by Eq. (1).

The acquisition time can be reduced either by decreasing Δf or increasing B (or both). The former can be accomplished to a limited extent by using trajectory information and also measured frequencies just prior to launch. Note that since the launch site will be separated from the transmitter site, a communication link is required for tranferring this informatio... Even then the pre-launch frequency information is of questionable value due to the mechanical and thermal shock associated with the launch. Experimentation with the crystal VCO in launch conditions would be required to determine the accuracy of this technique and whether the improvement was worth the effort or not.

At present it does not appear wise to attempt to utilize anything but known trajectory information. We estimate that it should be possible to at least estimate the Doppler offset from the trajectory information to within 10% (certainly, the polarity of the Doppler is known). Consequently, we expect that the maximum frequency uncertainty can be reduced by a factor of 2 to ±45 kHz. As a result of using trajectory information, the acquisition time could be reduced by a factor of 4.

If the PLL bandwidth is increased (the second alternative), the reference-signal phase-lock-loop SNR will be lowered. This degradation will result in a poorer measurement of the channel characteristics as described in Section V. Alternatively, one may think of all increased loop noise bandwidth as decreasing the maximum range or requiring increased transponder power.

If the phase-lock-loop acquisition times present a significant problem, a more complex phase-lock-loop design could be employed. A hybrid automatic frequency control (AFC)/automatic phase control (APC) loop could be employed. For the presently planned applications, this sophistication is not required. However, a simple modification to both the transponder and ground receiver phase-lock loops would allow the proposed design to work with much greater frequency uncertainty.

B. Delay-Lock-Loop Acquisition Time

We assume that we have constructed a delay-lock-loop tracking system with a discriminator that is two chips broad, because this provides the better search rate for the commonly employed discriminator types. In this case it may be shown that the maximum search rate, R, in a noise-free environment must be limited by the inequality

$$R \leq 2B \quad , \tag{10}$$

where B is the closed-loop equivalent-noise bandwidth of the delay-lock loop, and R is measured in chips/s. If we conservatively limit the search rate to one-half the maximum value to reduce the probability of missed lock due to noise, then the acquisition times are those specified in Table 7. Note that they are presented at two different loop bandwidths.

The nominal 200-Hz tracking bandwidth is presented because this is the result that would be obtained if no special acquisition procedures were employed. For longer code lengths the acquisition time is rather lengthy--perhaps 10 s in the worst case. Of course, if acquisition does not occur on the first but on the second or third pass, the times of Table 7 need to be multiplied by two or three. In these cases the acquisition tones definitely may be longer than desired.

One simple approach to reducing the acquisition time is to increase the loop bandwidth. Increasing the loop bandwidth by a factor of 5 to 1000 Hz results in a 7-dB SNR decrease, but the acquisition time is reduced by a factor of five provided an adequate SNR is available. Table 7 also presents the acquisition time for the case of a 1000-Hz bandwidth.

Note that the closed-loop bandwidth is a function of the RF SNR because the loop gain is a function of the signal strength of the coded signal. The latter varies with transponder range and the RF SNR in the AGC bandwidth. For low SNR the closed-loop bandwidth will be less than

Table 7

DELAY-LOCK LOOP ACQUISITION TIME

Loop Bandwidth (Hz)	Code Length (chips)	Maximum Acquisition Time (s)
200	127	0.64
	255	1.28
	511	2.55
	1023	5.1
	2047	10.2
1000	127	0.13
	255	0.25
	511	0.5
	1023	1.02
	2047	2.04

for high SNRs. Consequently, the system should be designed for the worst case or maximum range situation. At shorter ranges the SNR will be much higher and the loop bandwidth wider. As a result a higher search rate is possible, but we do not plan to take advantage of this fact other than to permit the operator to select the wider closed-loop bandwidth.

If one assumes that the maximum relative velocity of the transponder is 3.5 km/s, the channel itself provides a search velocity that is described in Table 8. This velocity-induced search rate may either slow or speed the selected search rate. Under the most likely circumstance, acquisition will occur during a closing situation and the search rate will be increased from that planned by the amount given in Table 8.

Based on Table 8 and our limit on search rate as specified by

$$R = B \quad , \tag{11}$$

Table 8

TRANSPONDER MOTION INDUCED
CHIP SEARCH RATE

Chip Rate	Effective Search Rate (chips/s)
10 MHz (low band)	116
20 MHz (high band)	232

either the code tracking loop must have a programmed bias offsetting the transponder motion-induced search velocity, or else the closed-loop tracking bandwidth must be increased from 200 to some number in excess of 232 Hz. In this case the closed-loop bandwidth of 1000 Hz seems a reasonable choice.

C. Phase Lock Loop Noise Analysis

In this section we analyze the steady-state noise performance of the phase-lock loops associated with the channel sounding tones and the reference beacon. We commence by analyzing the phase jitter propagation (multiplication, division, cancellation, and addition) through links using both the original and the simplified transponders. Next we determine the phase jitter as a function of loop SNR. Finally, for a representative example based on the link calculations of Section III-C, we evaluate the standard deviation of the tone phase jitter.

1. Phase Jitter Analysis of the Coherent Beacon Chain

In this section we analyze the noise performance (phase jitter) for the two transponder designs. For both designs we assume that the observed signal at the transponder input is given by

$$z(t) = s(t) + m(t) + n(t) \quad , \tag{12}$$

where s(t) are the desired tones, m(t) is other desired signals, and n(t) is the receiver noise. The received tones are given by

$$s(t) = k \sum_{i=1}^{L} \cos[p_i \omega_o t + p_i \theta] \quad , \tag{13}$$

where the parameters p_i are integers used in generating the coherent tones. For a two-band design L = 7 and 6 for the original and simplified transponder designs, respectively.

a. Original Transponder Phase Jitter Analysis

Figure 15 is a block diagram of the original transponder design. The block diagram is conceptual in the sense that we show the VCO operating at the basic frequency f_o, which is the fundamental frequency for the frequency synthesizer. In our proposed implementations, the transponder VCO would operate at a high harmonic of f_o. Thus, the phase jitter $\phi(t)$ will not be directly measurable in the proposed implementations. We show only the tone components of the received signal s(t) as an input. The channel sounding tone outputs are denoted y(t), while the reference beacon is denoted x(t). The phase jitter induced by the receiver noise n(t) is $\phi(t)$ referenced to the basic oscillator frequency f_o of the ground transmitter. The effective phase jitter in the transponder phase-lock loop is $p\phi(t)$, and its standard deviation must be considerably less than 30° if we are to be assured that the probability of loss of lock is sufficiently low. Thus, $\phi(t)$ will be very small.

The channel sounding tone output is given by

$$y(t) = \sum_{i=2}^{L} \cos[(p_i - p_1)\omega_o t + (p_i p_1)\theta - p_1 \phi(t)] \quad , \tag{14}$$

and the reference beacon by

$$x(t) = \cos[p_r \omega_o t + p_r \theta + p_r \phi(t)] \quad . \tag{15}$$

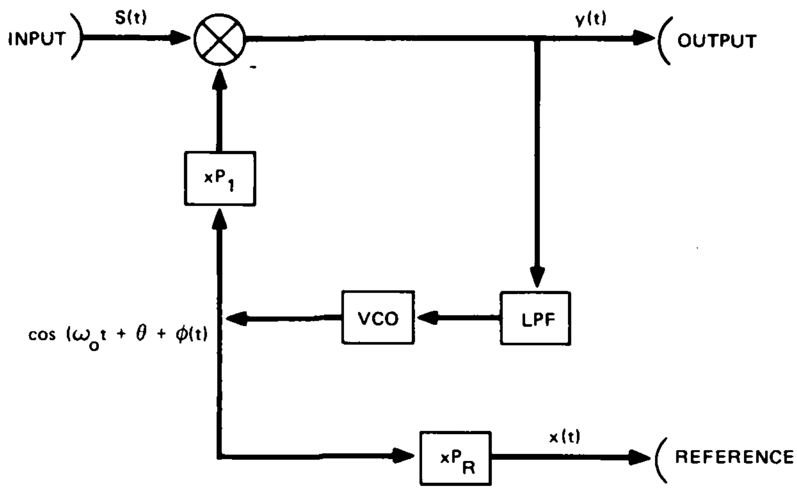


FIGURE 15 BLOCK DIAGRAM OF ORIGINAL TRANSPONDER DESIGN

The ground receiver phase-locks to the beacon and divides by p_r to produce the reference signal

$$r(t) = \cos[\omega_o t + \theta + \phi(t) + \Psi(t)] \quad . \tag{16}$$

where $\Psi(t)$ represents the phase jitter induced by the ground receiver noise referenced to the basic oscillator frequency f_o. The reference signal is used as the source signal for the receiver synthesizer.

The i^{th} tone output from the receiver synthesizer is given by

$$f_i(t) = \cos\{(p_i - p_1)\omega_o t + (p_i - p_1)[\phi(t) + \Psi(t)]\} \tag{17}$$

and the i^{th} component of the received channel sounding tones is given by

$$y_i(t) = \cos[(p_i - p_1)\omega_o t + (p_i - p_1)\theta - p_1\phi(t)] \quad . \tag{18}$$

When $f_i(t)$ and $y_i(t)$ are compared in a phase detector the difference $\gamma(t)$ is found to be

$$\gamma(t) = p_i\phi(t) + (p_i - p_1)\Psi(t) \quad . \tag{19}$$

Note that the ground receiver phase jitter has much less impact than the transponder phase jitter because $(p_i - p_1)$ is normally much smaller than p_i.

As an example, consider the case when the nominal carrier frequency is 8 GHz and the nominal oscillator frequency is 5 MHz. In this case $p_1 = 1600$, $p_i = 1580$, and $p_i - p_1 = -20$. However, it should be noted that $\Psi(t)$ is the received beacon phase noise referenced to f_o. Since the beacon is approximately 4 GHz, the integer dividing ratio is about 800. Thus, the phase jitter in both the transponder and the receiver phase-lock loops is reduced very substantially (by approximately 1600 and 800, respectively) in the process of creating $\phi(t)$ and $\Psi(t)$.

Nevertheless, $\gamma(t)$ is approximately 80 times more sensitive to $\phi(t)$ than $\Psi(t)$.

b. Simplified Transponder Phase Jitter Analysis

Figure 16 is a conceptual block diagram of the simplified transponder design. The reference beacon output is given by

$$x(t) = k_x \cos\,(p_n\,\omega_i t + \;p_n\theta) \quad . \tag{20}$$

The IF signal is given by

$$k(t) = \frac{1}{2}\sum_{i=1}^{L} \cos[p_i\omega_o\, t + p_i\theta - \omega_2\, t - \beta] \tag{21}$$

and the desired phase-locking component $\cos[p_L(\omega_o - \omega_2)t + p_L\theta - \beta]$ is phase-compared with $\sin[p_2\,\omega_1\, t + p_2\alpha]$ to produce the locking voltage

$$\sin[p_2\,\omega_1\, t + p_2\alpha \;-\; (p_L\omega_o - \omega_2)t + p_L\theta - \beta] \quad . \tag{22}$$

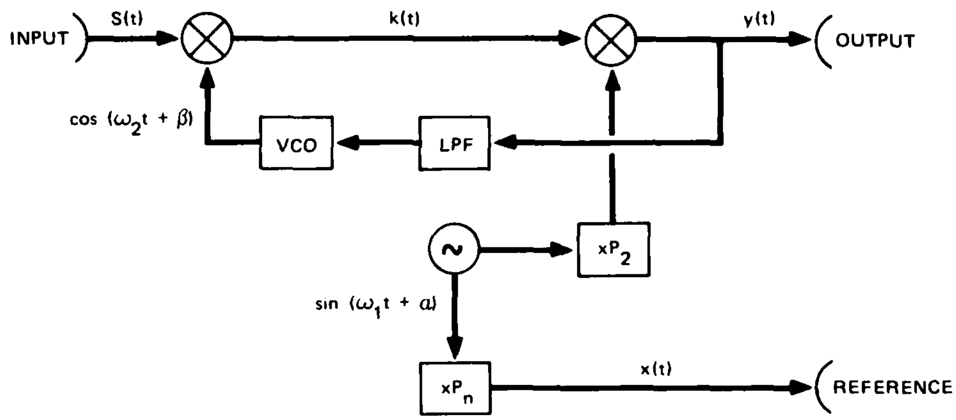


FIGURE 16 BLOCK DIAGRAM OF SIMPLIFIED TRANSPONDER

Lock is obtained only when the VCO adjusts itself so that

$$\omega_2 = p_L \omega_o - p_2 \omega_1 \tag{23}$$

and

$$\beta = p_L \theta - p_2 \alpha \quad . \tag{24}$$

In this case k(t) is given by

$$k(t) = 1/2 \sum_{i=1}^{L} \cos\{[(p_i - p_L)\omega_o + p_2\,\omega_1]\,t + (p_i - p_L)\theta + p_2\alpha\} \tag{25}$$

and the channel sounding tones by

$$y(t) = -1/4 \sum_{i=1}^{L} \sin\{[(p_i - p_L)\omega_o\;t + (p_i - p_L)\theta]\} \quad . \tag{26}$$

The phase-locking occurs at approximately 8 GHz and the phase noise $\phi_s(t)^*$ in the PLL shows up directly in the IF as

$$k(t) = 1/2 \sum_{i=1}^{L} \cos\{[(p_i - p_L)\omega_o + p_2\omega_1]t + (p_i - p_L)\theta + p_2\alpha - \phi_s(t)\} \quad (27)$$

and in the output as

$$y(t) = 1/4 \sum_{i=1}^{L} \sin\{[(p_i - p_L)\omega_o]t + (p_i - p_L)\theta - \phi(t)\} \quad . \quad (28)$$

Note that the reference beacon is not corrupted by the PLL phase jitter. Consequently, there is no phase cancellation that takes place in the ground receiver.

Consider now the i^{th} tone component

$$y_i(t) = -1/4 \sin[(p_i - p_L)\omega_o]t + (p_i - p_L)\theta - \phi(t)\} \quad . \quad (29)$$

The effect of the ground receiver PLL phase noise is the same as it was with the original transponder as analyzed in the previous section. That is, the measured phase error will be given by

$$\gamma_s(t) = -\phi_s(t) + (p_i - p_1)\Psi(t) \quad . \quad (30)$$

Note that the performance specified by Eq. (30), (simplified transponder) differs only slightly from the performance specified by Eq. (19), original transponder), because the standard deviation of $\phi_s(t)$ and $p_i\phi(t)$ will be essentially the same. In fact, the variances of the first terms are related by:

---

*This phase noise is measured at the radio frequency of approximately 8 GHz rather than at the basic frequency f_o used in the previous section.

$$\mathrm{Var}\{\phi_s(t)\} = (p_1/p_i)^2 \ \mathrm{Var}\{\phi(t)\} \quad . \tag{31}$$

Since p_L and p_i are very close to each other,

$$\mathrm{Var}\{\gamma_s(t)\} \approx \mathrm{Var}\{\gamma(t)\} \quad . \tag{32}$$

2. Phase Jitter as a Function of PLL SNR

For PLL substantially above threshold and therefore operating in a linear mode it may be shown that the variance of the noise-induced phase jitter measured in radians is given by

$$\sigma_\phi^2 = 1/\mathrm{SNR\ loop} \qquad (\mathrm{rad}^2)$$

where SNR loop is the SNR measured in the PLL closed-loop equivalent noise bandwidth.

3. Nominal Phase Jitter at Maximum Range

Using the nominal link parameters of Section III-C we find that the transponder PLL and the ground beacon tracking PLL have loop SNRs of 30 dB and 20 dB, respectively. Thus, we find that

$$\mathrm{Var}\{p_i\phi(t)\} \approx \mathrm{Var}\{\phi_s(t)\} = 10^{-3} \qquad (\mathrm{rad}^2)$$

and that

$\mathrm{Var}\{(t)\} = 10^{-2}$ (rad^2) referenced to 4 GHz, but referenced to 5 MHz (f_o) we have $\mathrm{Var}\{\Psi(t)\} = 1.56.10^{-6}$.

Consequently,

$$\mathrm{Var}\{(p_i - p_1)\Psi(t)\} \approx 6.25.10^{-4} \qquad (\mathrm{rad}^2)$$

where we assume that the sounding tones lie at approximately 100 MHz and the variance of the overall phase jitter is given by

$$\mathrm{Var}\{\gamma(t)\} = 1.625 \times 10^{-3} \quad (\mathrm{rad}^2) \quad .$$

Therefore, the standard deviation is 4.10^{-3} rad or 2.3°. Clearly, any larger value of phase jitter will tend to degrade the system performance. However, at the maximum range a phase jitter of this magnitude is acceptable.

VI CONCLUSION

This report covers the design of the transparent-transponder system down to the level immediately before actual circuit design and mechanical layout. That is, signal levels and frequencies and component specifications such as gains and losses, bandwidths, and so on, have been established in detail. Careful consideration has been given to noise levels, dynamic range, intermodulation levels and products, and avoidance of harmonic regeneration and self-locking. Particular attention was paid to the problem of mutual interference between the various diagnostic and channel-sounding signals and whatever systems-test signals may be present for any particular application. Because only limited time spans may be available in both rocket-borne and possible satellite applications of this system, the times required for signal acquisition by phase-locked oscillators at both ends of the system and by PN tracking loops can be critically important, so those aspects were also given careful study. The chosen SNRs provide more than adequate acquisition times for all presently conceivable applications for the system. A transponder that maintains linearity through automatic gain control is superior to one that hard-limits the incoming signals.

Although many of the systems calculations presented here necessarily made use of nominal parameters that would apply for a generic mission because a specific application for the system has not been defined yet, the parameters are reasonable and the appropriate tradeoffs are discussed. Mission-dependent parameter variations are not expected to be large, and thus would not have a significant impact.

Two basic transparent-transponder configurations evolved during the course of this study and design effort. They differ in the way the uplink signals are acquired and translated for retransmission through the disturbed ionosphere back to a ground station, and thus in their complexity. The first, or "basic transponder" operates on an 8-GHz uplink signal carrier directly and is the more complex of the two because it requires a more involved frequency-multiplier chain. The more evolved, simpler system, which is called the "simplified transponder," locks on to a VHF pseudocarrier, which is actually a CW signal in the lower sideband of the true 8-GHz uplink signal. Its cost is the loss of round-trip Doppler information, which is not serious. Both systems use the same fundamental principle of translating the lower sidebands of the uplink signal to the VHF-UHF range for retransmission to the ground. Furthermore, the receiving systems for both approaches are nearly identical and could easily be compatible with both types of transponders. The transmitter systems would be essentially the same as well.

Two types of missions have been contemplated for this transparent-transponder system: rocket-borne and aboard earth-satellites. The fundamental premise for both is the provision of a system capable of handling a wide variety of signals that could be deleteriously affected by regions of disturbed plasmas. Although the ground station(s) would be essentially the same for both types of mission, and could be used interchangeably for both, the detailed design of the transponder could be substantially different because of cost and lifetime, and mission duration requirements. A rocket mission could tolerate shorter-lifetime (perhaps nearly "off-the-shelf") components and a larger instantaneous power consumption than possible in a satellite application.

In spite of a higher initial hardware cost, however, the satellite application of this system is probably the most cost-effective overall, and more fully exploits the unique capabilities of the concept. The possibility of making many passes over disturbed parts of the ionosphere (e.g., at the equator, or at auroral latitudes) is intrinsically complementary to the ability to easily change the types of signals passed through the transponder and their parameters. Besides being available for specific programs involving specialized types of communications signals, an orbiting transponder of this capability can also readily deal with the types of signals used for more routine or synoptic studies, such as the DNA Wideband Satellite set. We note that the Wideband Satellite signal set would be easily handled by the transparent transponder, with the exception that the reference signal would be about 5 GHz rather than at 2.8 GHz, which is a desirable feature. Thus, existing ground-station equipment could be used effectively.

An important aspect of this design effort was to provide the ability to pass a variety of signals at once through the transponder. The signals are not restricted to fall in the same bands nor to be separated in frequency. This raises the possibility of simulating the interaction between signals passing through different parts of a nuclear-disturbed plasma, as might be encountered in a jamming scenario.

No particular mission or specific application has been established for the system described here. The objective of this effort was to establish the basis and set the detailed specifications for the system and its components so that construction could proceed quickly and efficiently.

REFERENCES

1. W. Sollfey, "Transionospheric Propagation of Pulsed Signals," Memorandum RM-5558-NASA, The Rand Corporation, Santa Monica, CA (May 1968).

2. Robert F. Daly, "On Modeling the Time Varying Frequency Selective Radio Channel," Technical Report 2--Part II, Contract DA36-039 SC-90859, SRI Project 4172, Standord Research Institute, Menlo Park, CA (July 1964).

3. E. J. Fremouw, D. J. Barnes, R. C. Livingston, M. D. Cousins, B. C. Fair, and G. K. Durfy, "Wideband Satellite Experiment--Preparation for Launch," Final Report--Phase III, SRI Project 1972, Stanford Research Institute, Menlo Park, CA (November 1974).

4. D. Richman, "Color-Carrier Reference Phase Synchronization Accuracy in NTSC Color Television," Proc. IRE, Vol. 42, pp. 106-133 (January 1954).

Appendix A

RFI AND DOWNLINK BAND SELECTION

Interference can be a very serious problem because of the wide bandwidths occupied during time-domain channel soundings (and for frequency-domain work too, if there are many tones) and spread-spectrum signal testing, plus the need to operate in the VHF range in order to obtain relatively narrow coherence bandwidths. This problem is particularly severe in the CONUS, where there is a high density of often high-powered transmitters. This was recognized as a constraint on the PLACES program, and ESL, Inc. made an RFI survey in the Florida panhandle area to try to quantify interference levels.* This appendix contains the results from a similar, but less extensive, measurement of the "background" signal level between 50 and 350 MHz. Our measurements were done on the roof of a building at SRI International on 2 February 1979, in the San Francisco Bay Area, which is probably very cluttered with signals.

Our spectrum analyzer was set up to scan over 50-MHz intervals, and the center frequency was shifted in 25-MHz steps. Figure A-1 is a block diagram of the measurement setup. The signal generator provided for signal-strength calibration was set to the center frequency of each scan. A quarter-wave vertical antenna was used, which means that gain was maximum at the horizon and there was no azimuthal directionality. Table A-1 presents the spectrum analyzer settings and Table A-2 is a list of the antenna lengths. Measured cable losses were less than 0.5 dB for the RG-214/U section and less than 2 dB for the RG-58/U piece.

Figures A-2 and A-3 summarize the measurements. By far the most dominant feature is the FM broadcast band from 88 to 108 MHz. The central

*Letter dated 5 January 1979 from C. W. Prettie to Captain L. A. Wittwer.

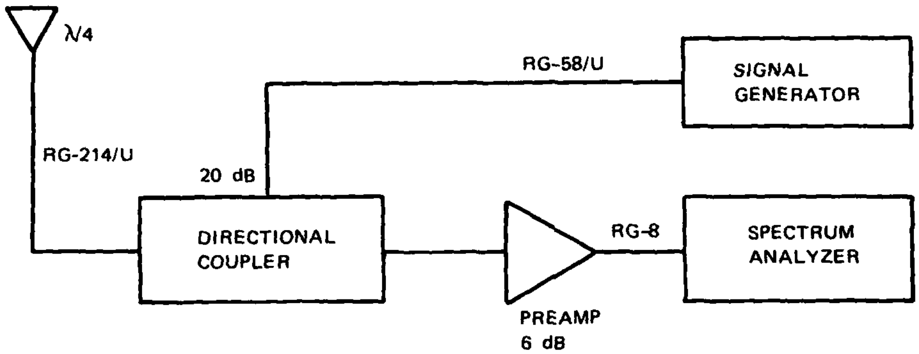


FIGURE A-1 TEST SETUP

Table A-1

SPECTRUM ANALYZER PARAMETERS

Scan width	50 MHz (5 MHz/division)
Bandwidth	100 kHz
Scan time	1 ms/division
Log reference	-30 dB
Signal generator	-40 dBm

part of the aircraft band (108 to 136 MHz) has much less energy. However, the signals are very sporadic, as are signals at slightly higher frequencies (Figure A-4). The so-called VOR navigational aids, however, which occupy the 103-to-118-MHz band, produce continuous signals. Note that the TV broadcasts produce 4 to 5 orders of magnitude less average energy density than appears in the FM band. At least in this area, it would be far more preferable to make measurements below 88 MHz and avoid the FM band.

Since the aircraft band is probably very sensitive to even the slightest amount of interference, 132 MHz seems to be the lower limit

Table A-2

ANTENNA LENGTHS

Center Frequency (MHz)	Antenna Length (cm)
75	99.1
100	74.3
125	59.4
150	49.5
175	42.4
200	37.1
225	33.0
250	29.7
275	26.9
300	24.8
325	22.9

for the next-lowest possible band for measurements. There appear to be a number of narrowband signals within the 140-to-165-MHz range, however, whose effects would have to be considered. Except for the TV carriers, things seem to be fairly quiet above that. (The 300-to-350-MHz range, which is not shown, was similar to the 250-to-300-MHz range.) Our measured signal levels are comparable to those reported by ESL, Inc.

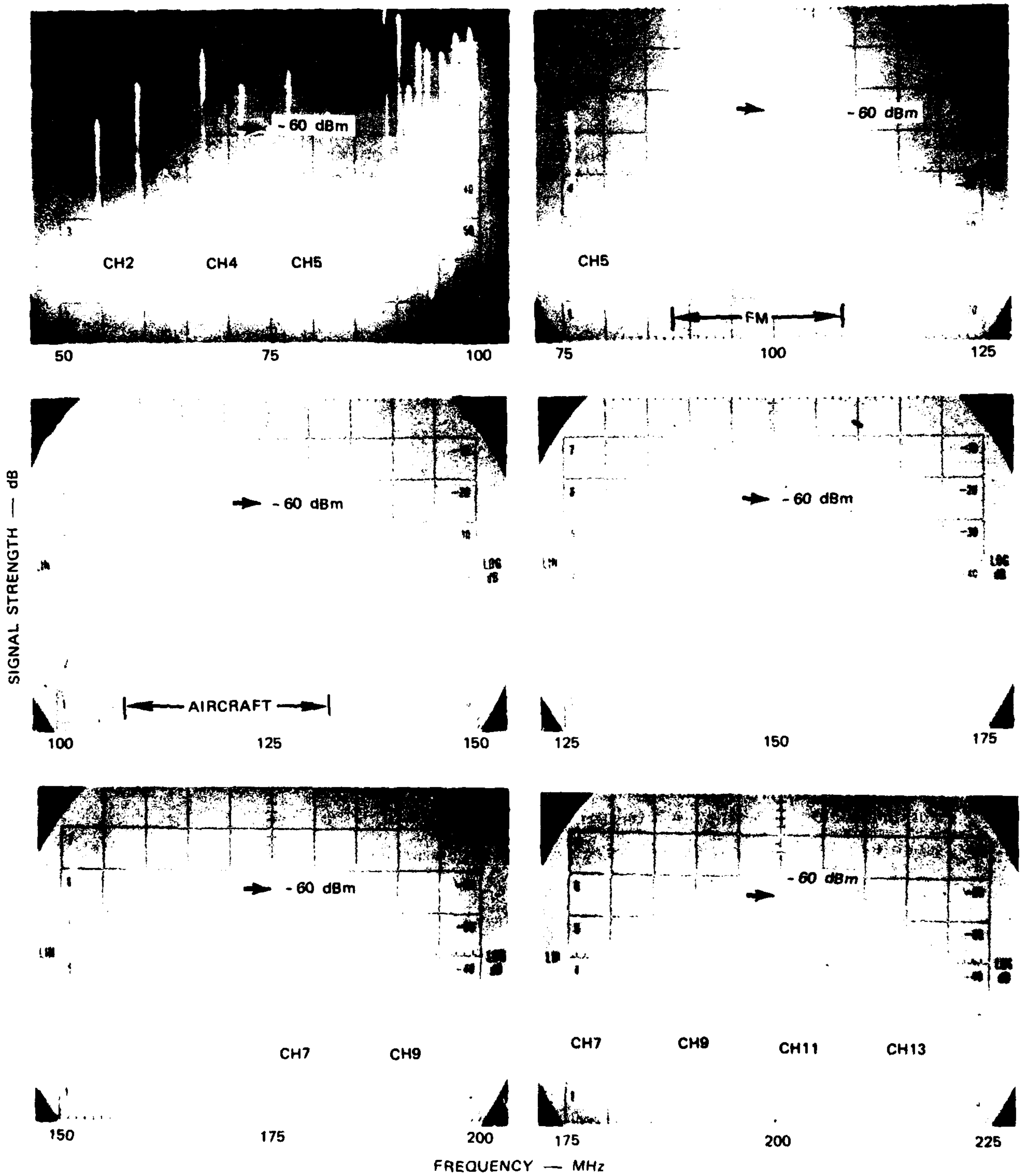


FIGURE A-2 SPECTRUM ANALYSIS FROM 50 TO 225 MHz

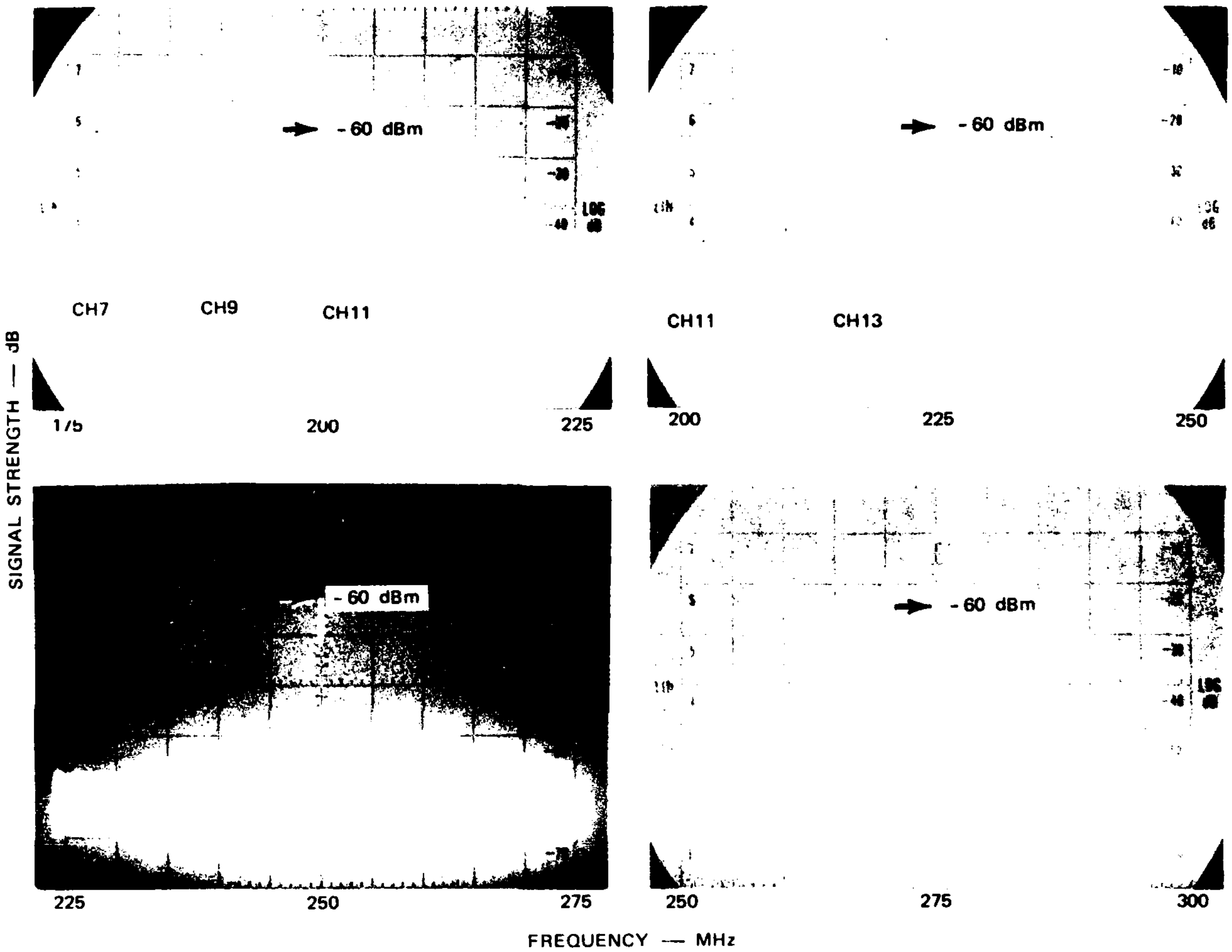


FIGURE A-3 SPECTRUM ANALYSIS FROM 175 TO 300 MHz

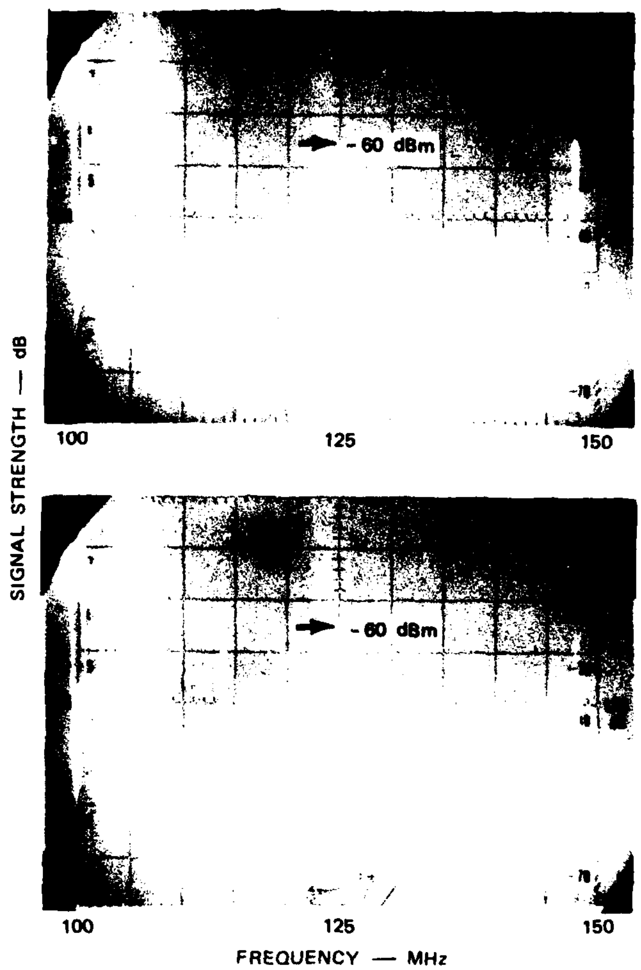


FIGURE A-4 100-TO-150-MHz SPECTRA AT DIFFERENT TIMES

DISTRIBUTION LIST

DEPARTMENT OF DEFENSE

Command & Control Technical Center
ATTN: C-312, R. Mason
ATTN: C-650, G. Jones
3 cy ATTN: C-650, W. Heidig

Defense Communications Engineer Center
ATTN: Code R123
ATTN: Code R410, N. Jones

Defense Nuclear Agency
ATTN: NAFD
ATTN: STNA
ATTN: RAEE
ATTN: NATD
3 cy ATTN: RAAE
4 cy ATTN: TITL

Defense Technical Information Center
12 cy ATTN: DD

Field Command
Defense Nuclear Agency
ATTN: FCP

WWMCCS System Engineering Org
ATTN: J. Hoff

DEPARTMENT OF THE ARMY

U.S. Army Chemical School
ATTN: ATZN-CM-CS

U.S. Army Communications Command
ATTN: CC-OPS-W
ATTN: CC-OPS-WR, H. Wilson

U.S. Army Nuclear & Chemical Agency
ATTN: Library

U.S. Army Satellite Comm Agency
ATTN: Document Control

DEPARTMENT OF THE NAVY

COMSPTEVFOR
Department of the Navy
ATTN: Code 605, R. Berg

Naval Research Laboratory
ATTN: Code 4780, S. Ossakow

DEPARTMENT OF THE AIR FORCE

Air Force Weapons Laboratory
Air Force Systems Command
ATTN: SUL
ATTN: NTYC
ATTN: NTN

Air Force Wright Aeronautical Lab
ATTN: W. Hunt
ATTN: A. Johnson

DEPARTMENT OF THE AIR FORCE (Continued)

Air Weather Service, MAC
Department of the Air Force
ATTN: DNXP, R. Babcock

DEPARTMENT OF ENERGY CONTRACTORS

Los Alamos National Laboratory
ATTN: D. Simons

Sandia National Lab
ATTN: Org 1250, W. Brown

DEPARTMENT OF DEFENSE CONTRACTORS

E-Systems, Inc
ATTN: R. Berezdivin

ESL, Inc
ATTN: J. Marshall

General Research Corp
ATTN: J. Ise, Jr
ATTN: J. Garbarino

University of Illinois
ATTN: Security Supervisor for K. Yeh

JAYCOR
ATTN: J. Sperling

Kaman Tempo
ATTN: DASIAC
ATTN: W. Knapp

M.I.T. Lincoln Lab
ATTN: D. Towle

Mission Research Corp
ATTN: Tech Library
ATTN: R. Bogusch

Mitre Corp
ATTN: G. Harding
ATTN: A. Kymmel
ATTN: B. Adams
ATTN: C. Callahan

Pacific-Sierra Research Corp
ATTN: H. Brode

Physical Dynamics, Inc
ATTN: E. Fremouw

R & D Associates
ATTN: R. Lelevier
ATTN: M. Gantsweg
ATTN: P. Haas
ATTN: B. Gabbard

R & D Associates
ATTN: B. Yoon

DEPARTMENT OF DEFENSE CONTRACTORS (Continued)

SRI International
ATTN: W. Jaye
ATTN: R. Leadabrand
ATTN: D. Neilson
ATTN: C. Rino
ATTN: J. Petrickes
ATTN: W. Chesnut
ATTN: R. Livingston
ATTN: R. Tsunoda
ATTN: G. Smith
ATTN: G. Price
ATTN: M. Baron
4 cy ATTN: A. Burns
4 cy ATTN: D. Magill
4 cy ATTN: D. Barnes

DEPARTMENT OF DEFENSE CONTRACTORS (Continued)

Rand Corp
ATTN: C. Crain
ATTN: E. Bedrozian

Science Applications, Inc
ATTN: L. Linson

DAT

ILM