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ANALYSIS OF THE PROPOSED UTILIZATION OF THE TDRS SYSTEMS BY THE HEAO-C SATELLITE

FINAL REPORT
NGR-01-001-021

MAY 29, 1974

ALABAMA AGRICULTURAL AND MECHANICAL UNIVERSITY
SCHOOL OF TECHNOLOGY
HUNTSVILLE, ALABAMA

158

ANALYSIS OF THE PROPOSED UTILIZATION OF THE TDRS SYSTEM BY THE HEAO-C SATELLITE

FINAL REPORT
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Submitted by

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CONTENTS

1. Introduction

This document reports the result of a study of possible utilization of the Tracking and Data Relay Satellite System (TDRS) by the High Energy Astronomical Observatory Satellite (HEAO-C). The primary function of the study was to assess the impact upon the HEAO telecommunications system of the proposed relay satellite-to-ground-link configuration.

The TDRS-system design is being managed by the Goddard Space Flight Center. The system was originally planned for the late seventies, and it is designed to perform the function of most of the NASA ground-tracking and communications network at a net cost savings for NASA. At this time, NASA hopes to have the TDRS system in operation by 1979, with private industry, using the NASA technology, deploying the system and NASA leasing most of the TDRS capacity.

The HEAO program is being managed by the Marshall Space Flight Center. NASA plans to launch HEAO spacecraft in 1977, 1978, and 1979. The HEAO-C satellite (1979) is a potential user of the TDRS system.

If the HEAO-C does indeed become a TDRS system user, the cost of the satellite telecommunications system will definitely be increased. Only the decreased costs of maintaining the ground stations can justify the increased cost and complexity of the HEAO-C telecommunications system.

Several ground-rules were accepted in defining the scope of the study. They included accepting the present definition of the TDRS system (antenna gains, transmitter power, etc.), and determining the parametric requirements of the HEAO-C in order to maintain the required communications standards. This report, then, is primarily a description of the TDRS system as it relates to the proposed TDRS-HEAO-C communication links, a description of the HEAO-C data handling requirements, and a discussion of the resulting HEAO-C telecommunication requirements, and a design of a proposed HEAO-C telecommunications system.

Detailed effort in the study has been directed toward several specific system components. This includes the pseudonoise spread-spectrum modem transponder for the HEAO-C and ground receiver. More general consideration was given to components such as the spacecraft antenna system and error control coding and decoding schemes. Also general power budget data were calculated to confirm the feasibility of the proposed communication link.

2. Tracking and Data Relay Satellite System, Current Specifications

This section describes the TDRS telecommunications system as it relates to the proposed HEAO-C TDRS communications link. Table 2-1 shows the TDRS frequency plan. Table 2-2 shows the Telecommunication service specifications for low and medium data rate users. The HEAO-C will be a MDR user,therefore in this section most of the material will relate to the MDR section of the TDRS system.

The TDRS is designed to accomodate two medium data rate users. This is accomplished with the telecommunications substem block diagram as shown in figure 2-1. The TDRS-user forward link has S-band and Ku-band capability. The MDR TDRS antenna system is assured to be an 8 (eight) foot parabolic dish antenna with 60% efficiency. (This figure is in variance with some initial TDRS design, such as the 6.5 foot dish suggested by the RI TDRS study.) The system has the capability of supporting two MDR users with any combination of Ku or S-band, and one of the MDR space-to-space forward link transponders and antenna can serve as a back-up for the TDRS-Ground Ku-band link.

One initial design of the TDRS specifies that the MDR-S band capability is for users with data rate less than 10^5 bps. Exception to the specifications is taken in this study.

The TDRS transponder functions as a linear translator in both the forward and return link. For this reason, and because power flux density impinging on the earth's surface from the TDRS satellite is limited by IRAC standards, the data must be spread in bandwidth, this spread spectrum is achieved by modulating the information signal by a pseudonoise signal. The pseudonoise signal selected for this application is a shift register generated pseudo random digital sequence.

The IRAC guidelines for maximum power flux density at the earth's surface for various frequency bands are given in table 2-3.

	Links	Frequency	Channel Bandwidth
Forward Link	• LDR	400.5 to 401.5 MHz	1 MHz 4 - 250 Khz channels
	• MDR S-band Ku-band	 2025 to 2120 MHz 14.6 to 15.2 GHz	 95 MHz channel 4 - 100 MHz channels
	• TDRS/GS Ku-band VHF S-band	 13.4 to 13.64 GHz 148.26 MHz 2200 to 2290 MHz	 240 MHz 90 MHz
	• Tracking/Order Wire	2066 MHz	
	• Ku Beacon	15.0 GHz	
Return Link	• LDR	136 to 138 MHz	2 MHz (20 users multiple accessed/TDRS)
	• MDR S-band	 2200 to 2300 MHz	 20-10 MHz slots in 5 MHz steps or 100 MHz wide open
	• Ku-band	13.6 to 14.0 GHz	4-100 MHz channels
	• TDRS/GS Ku-band VHF S-band	 14.6 to 15.2 GHz 136.11 MHz 2025 to 2110 MHz	 200 or 600 MHz channel 85 MHz
	• Tracking/Order Wire	2249 MHz	

Table 2-1 TDRS Frequency Plan

Description	LDR User	MDR User
Number of users	Forward: Minimum of 1 Return: 20	Minimum of 1
Frequency	Forward: VHF, UHF, S-band Return: VHF	S- or X- or Ku-band
Communications requirement	Forward: 100 to 1000 bps Return: 1 to 10 kbps	Forward: 100 to 1000 bps Return: 10 to 1000 kbps
Constraints	·Linear transponder in return link ·High RFI ·Flux density (IRAC): VHF $\leq$ -144 dBw/m^2/4 khz UHF $\leq$ -150 dBw/m^2/4 khz S-band $\leq$ -154 dBw/m^2/4 khz ·EIRP = +30 dBw/channel (VHF, UHF) = +41 dBw/channel(S) ·BER = 10^{-5}	·Linear TDRS transponder return link ·Variable frequency ·Flux density (IRAC): S-band $\leq$ -154 dBw/m^2/4 khz X-band $\leq$ -150 dBw/m^2/4 khz Ku-band $\leq$ -152 dBw/m^2/4 khz ·BER = 10^{-5}

Table 2-2 TDRS Telecommunications Service Specifications

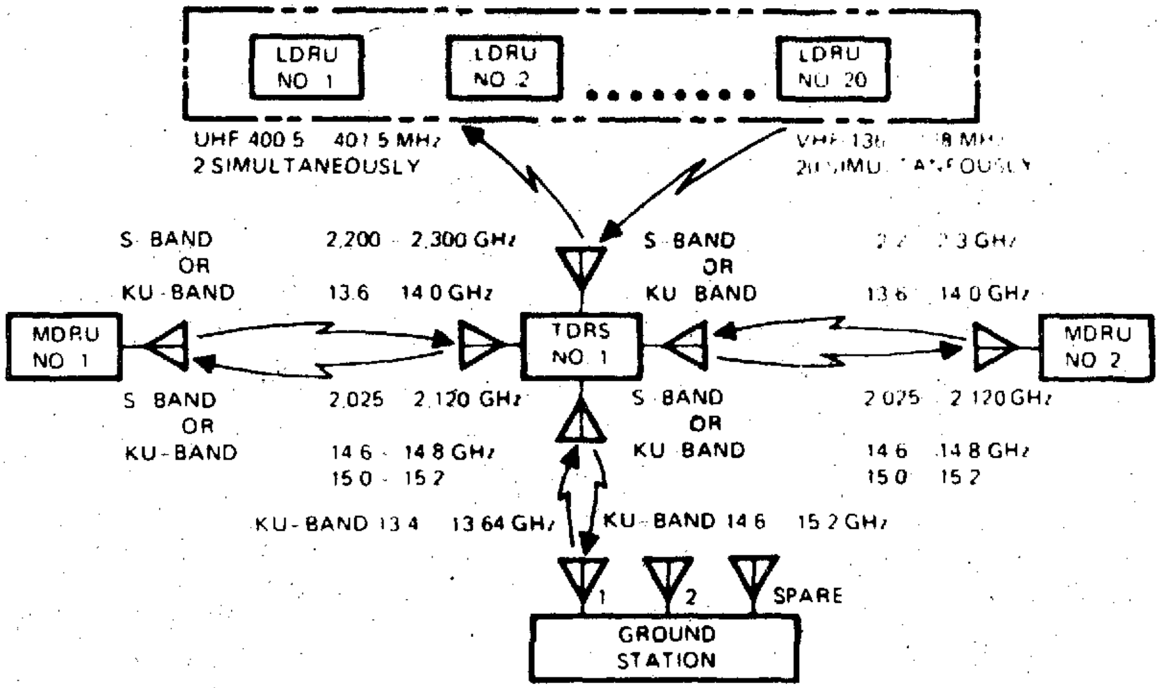


Figure 2-1 Telecommunications Service Subsystem Block Diagram

Frequency Band	Flux Density in 4 kHz
VHF	-144 dBw/m^2
UHF	-150 dBw/m^2
S-band	-154 dBw/m^2
X-band	-150 dBw/m^2
Ku-band	-152 dBw/m^2

Table 2-3 IRAC Guidelines

3. HEAO-C, Telecommunications Requirements

The Telecommunication requirements of the HEAO-C satellite are assumed as follows:

	LOW RATE MODE	1 TDRS
Forward Link:	1 kbps	Command Channel
Return Link:	6.4 kbps	Real time telemetry
	12.8 kbps	Recorded data
	19.2 kbps	TOTAL DATA RATE

	LOW RATE MODE	2 TDRS
Forward Link:	1 kbps	Command Channel
Return Link:	6.4 kbps	Real time telemetry
	3.2 kbps	Recorded data
	9.6 kbps	TOTAL DATA RATE

	HIGH RATE MODE	
Forward Link:	1 kbps	Command Channel
Return Link:	128 kbps	Real time experimental data

Power link margin is specified as 10db for a telemetry BER of 1 part in 10^6.

Uplink commands at 1 kbps are PCM/PSK/FM modulated on a 70kHz subcarrier which is phase modulated on the carrier.

The HEAO-C command and data handling subsystem block diagram is shown in figure 3-1. The nomenclature used in this diagram is as follows:

CDHS: Command and Data Handling System
PCD: Primary Command Decoder

SCD : Secondary Command Decoder

SCP : Stored Command Programmer

CCU : Central Clock Unit

RMU : Remote Multiplexing Unit

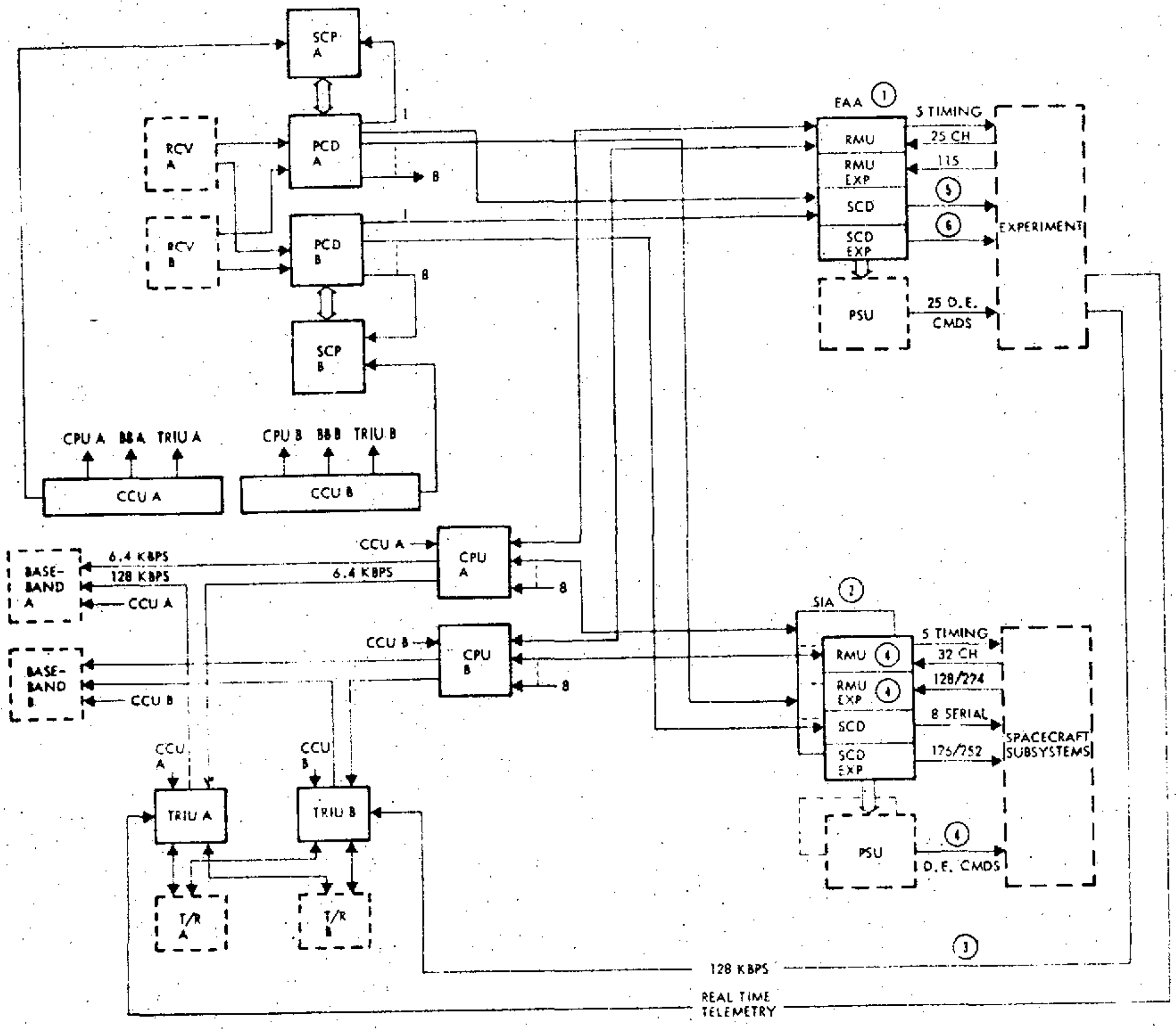
TRIU: Tape Recorder Interface Unit

USB : Unified S-Band

TA : Transfer Assembly

The overall communication and tape recorder system block diagram for HEAO-C - TDRS capability is shown in Figure 3-2.

The impact on the HEAO-C instrumentation of the TDRS relay link would be the reduction or elimination of the HEAO-C tape recorder requirement.



NOTES:

1. 4 NON-REDUNDANT EAA FOR HEAO A
   1 REDUNDANT EAA FOR HEAO B
   3 NON-REDUNDANT EAA FOR HEAO C
2. 1 REDUNDANT SIA FOR HEAO A, B, C
3. REDUNDANT OUTPUTS FROM A SINGLE EXPERIMENT
4. REQUIREMENT UNDER STUDY
5. 8 SERIAL HEAO A AND C
   16 SERIAL HEAO B
6. 90 DISCRETE COMMANDS HEAO A AND C
   NO DISCRETE COMMANDS HEAO B

FIGURE 3-1

COMMAND AND DATA HANDLING SUBSYSTEM BLOCK DIAGRAM

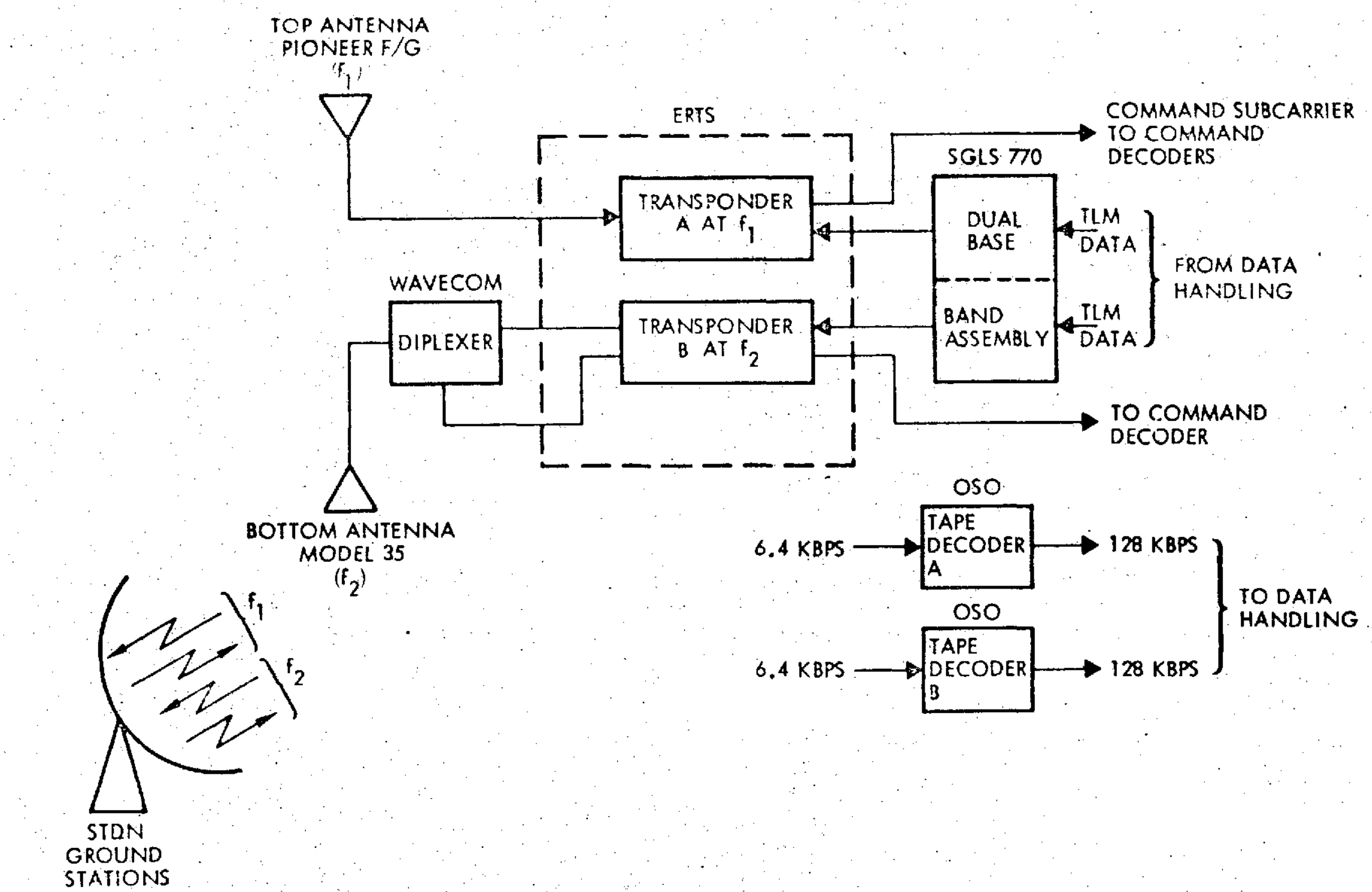


Figure 3-2 Communication and Tape Recorder System Block Diagram

4. Communication Link Analysis

This section is divided into two parts, a procedure for factoring the effects of spread spectrum techniques into power budget calculations and TDRS-HEAO-C power budget calculations.

(a) Procedure for Factoring the Effects of Spread Spectrum Techniques into Power Budget Calculations

Normal power budget calculations for single-user, single-channel applications consist of the following steps:

(1) Determining the system parameters such as transmitter power transmitter loss, transmitter antenna gain, atmospheric loss, polarization loss, space loss, receiver antenna gain, receiver noise figure, preamplifier noise temperature, receiver losses, and bandwidth.

(2) Using appropriate parametric relations, calculate the contrast-ratio at the receiver output.

(3) Alternately, given the required contrast ratio for a given error rate limit at the demodulator, optimize the parameters in (1) to yield the required contrast ratio. The optimization procedure might require minimization of system cost, minimization of transmitter power, minimization of receiver antenna gain. Normally, the procedure requires some subjective parametric tradeoffs based on engineering judgement, along with objective parametric tradeoffs based on a mathematical optimization of the system model.

The purpose of this section is to give a straightforward procedure for including the effects of a spread spectrum implementation upon the final contrast ratio for a communications links under conditions that include multipath and RFI conditions.

An analytical consideration of the problem is complicated by the fact that under realistic conditions both RFI and multipath interference are characteristically nonstationary.

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MULTIPATH GEOMETRY

A single model for the purpose of contrast ratio estimation under conditions of multipath is based on the configuration of figure 1,

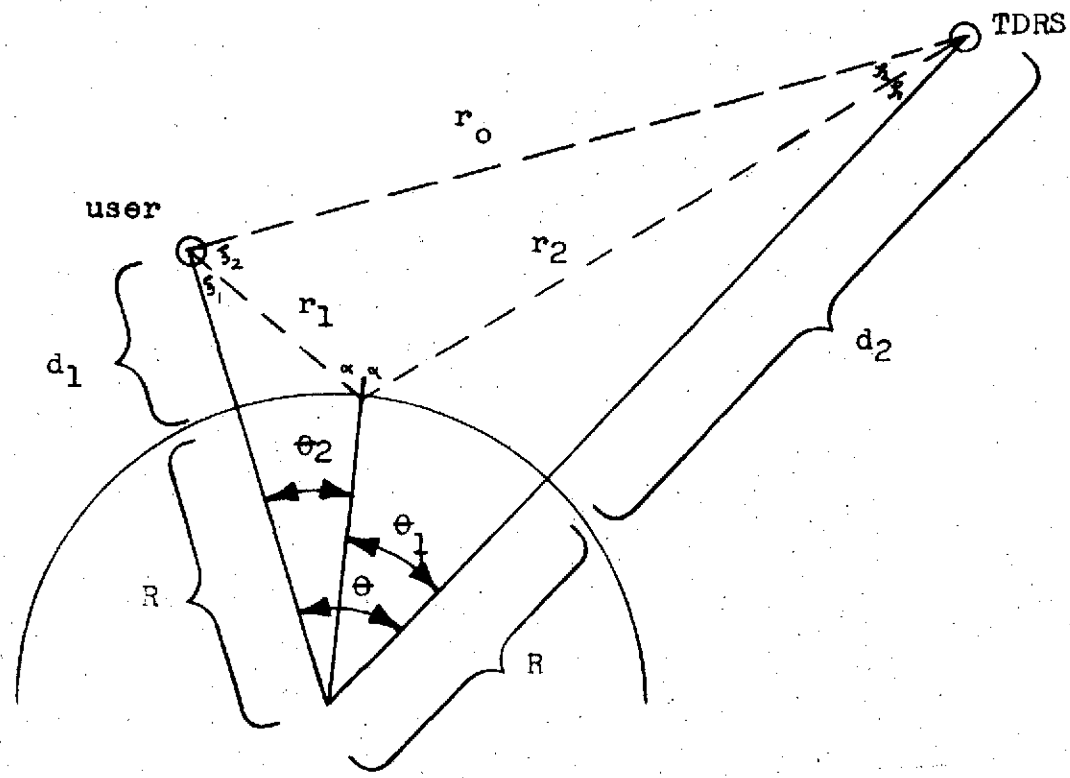

Figure 4-1 TDRS-USER DIRECT AND MULTIPATH SIGNAL ROUTES

$$r_o^2 = (d_1 + R)^2 + (d_2 + R)^2 - 2\,(d_1 + R)\,(d_2 + R)\,\cos(\theta)\;. \tag{4-1}$$

Using the simplifying assumption that θ is small the law of sines as applied to figure 1 yields

$$\frac{R + d_1}{\alpha} = \frac{R}{\xi_1} \tag{4-2}$$

$$\frac{R + d_2}{\alpha} = \frac{R}{\xi_1}\;. \tag{4-3}$$

Also

$$\xi_1 + \theta_1 - \alpha = 0 \tag{4-4}$$

$$\xi_1 + \theta_2 - \alpha = 0\;. \tag{4-5}$$

And

$$\theta_1 + \theta_2 = \theta\;. \tag{4-6}$$

Solution of these equations yields

$$\xi_1 = \frac{R\,\theta(R + d_2)}{R(d_1+d_2)+2d_1d_2} \tag{4-7}$$

$$\mathcal{S}_1 = \frac{R\theta(R+d_1)}{R(d_1+d_2)+2d_1d_2} \tag{4-8}$$

$$\alpha = \frac{\theta(R+d_1)(R+d_2)}{R(d_1+d_2)+2d_1d_2} \tag{4-9}$$

$$\theta_1 = \frac{\theta(R+d_2)\,d_1}{R(d_1+d_2)+2d_1d_2} \tag{4-10}$$

and

$$\theta_2 = \frac{\theta(R+d_1)\,d_2}{R(d_1+d_2)+2d_1d_2} \quad . \tag{4-11}$$

The approximate time delay for the reflected signal path is

$$T_d \approx \frac{1}{3X10^8}\left[d_2 + d_1 - (d_2 - d_1)\right]$$
$$= \frac{2d_1}{3X10^8} \quad , \tag{4-12}$$

and the slant-range to the user or TDRS from the reflection point is approximately equal to the orbit altitude.

The spherical surface of the earth causes divergence of the reflected signal in figure 1. This results in an attenuation

$$A = \frac{rs^2}{(d_1+rs)^2} \quad , \tag{4-13}$$

where rs is the distance from the point of reflection to the virtual signal source. For small α,

$$A = \frac{R^2}{(2d_1 + R)^2} \quad . \tag{4-14}$$

Figure 4-2 illustrates the reflecting area for a particular TDRS-USER position.

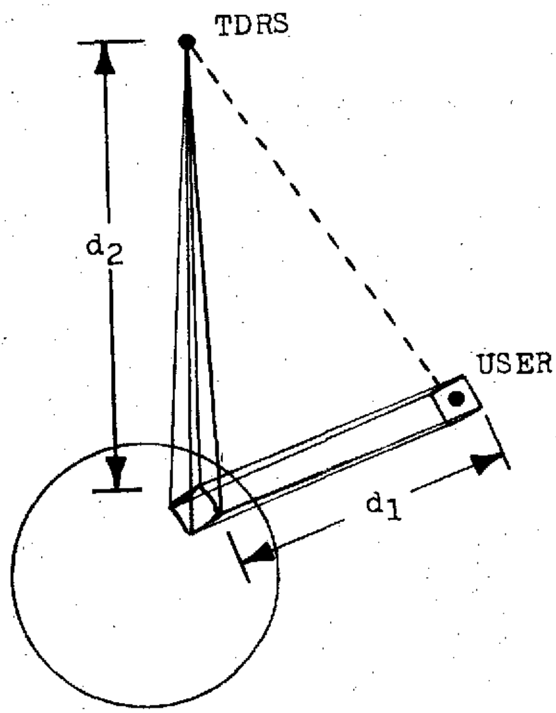

Figure 4-2 TDRS-USER Geometry

In general, because of the complex nature of the earth's surface, the multipath signal to the user arrives from a relatively large area of the earth's surface. This results in a Doppler and delay distortion of the returned signal. This spread in multipath delay causes a distortion in the autocorrelation function of the PN code modulated on the multipath component of the TDRS-User signal. This distortion reduces the likelihood that the user PN transponder will "lock-on" the multipath component of the signal. If the bandwidth of the autocorrelation filter is B, the digital PN rate is fc, and

$$\frac{2d_1}{3 \times 10^8} > \frac{1}{B} , \qquad (4\text{-}15)$$

then the multipath component of the user input signal is spread with minimum effect on the transponder PN-loop tracking.

To facilitate the power budget analysis, the PN spread spectrum system will now be described.

General Description of Spread Spectrum Systems

In general, spread spectrum systems are those communications schemes which utilize a wider spectrum bandwidth than the channel information signal bandwidth. A simple FM communications system can be interpreted as a spread spectrum system under this definition. The communications processing is performed in such a way that the extra spectrum utilization results in improved signal-to-noise ratio at the receiver output. In other words, spectrum has been traded for increased signal-to-noise ratio.

Spread spectrum systems are very effective in situations where interference is a significant factor. In space communications, the spread spectrum concept has been proposed to combat multipath fading and provide interfference rejection in such applications as communication-relay-satellite to user links.

Pseudo-Noise Systems

A pseudo-noise system is a particular form of spread spectrum system that combats multipath fading and interference, and in addition can provide ranging, telemetry synchronization, and addressing. The term "pseudo-noise" refers to noise-like sub-carriers used in such systems, usually generated from a maximum length digital sequence generator. Maximum length digital sequences have many properties that make them statistically similar to true random digital sequences. One of these is the similarity of the autocorrelation functions of maximum length and true random digital sequences as illustrated in figure 4-3. In the figure, T_B is the digital clock period and $L = 2^n-1$.

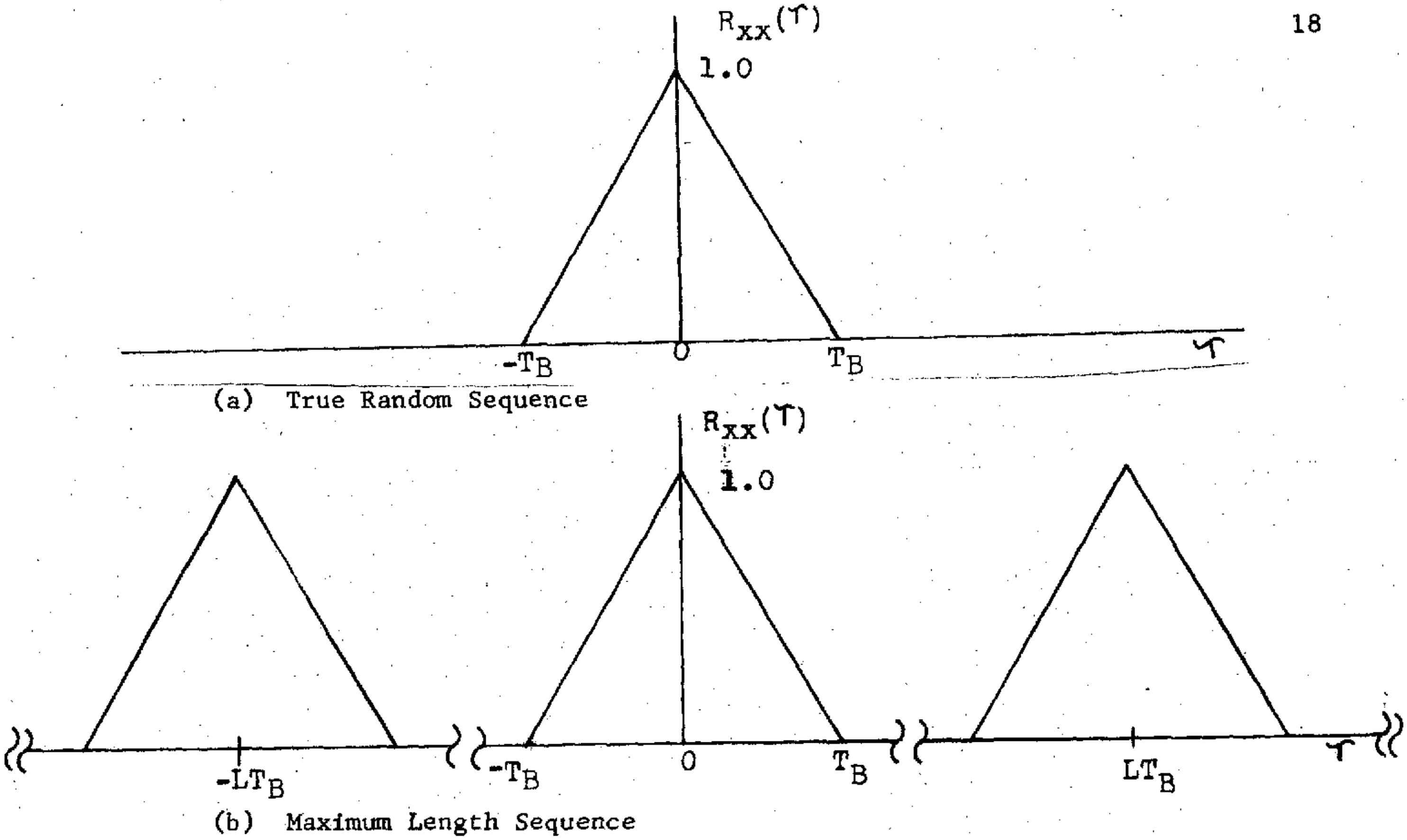

Figure 4-3 Autocorrelation Function of True Random Sequence and Maximum Length Sequence.

The power spectrum of the true-noise sequence is the Fourer transform of the autocorrelation function,

$$S(\omega) = F\ (R_{xx}(\tau)) = \int_{-\infty}^{\infty} R_{xx}(\tau)\varepsilon^{-i\omega\tau}\, d\tau = T_B(\mathrm{Sin}(\omega T_B/2))^2/(\omega T_B/2)^2 \quad . \tag{4-16}$$

Figure 4-4illustrates the power spectrum.

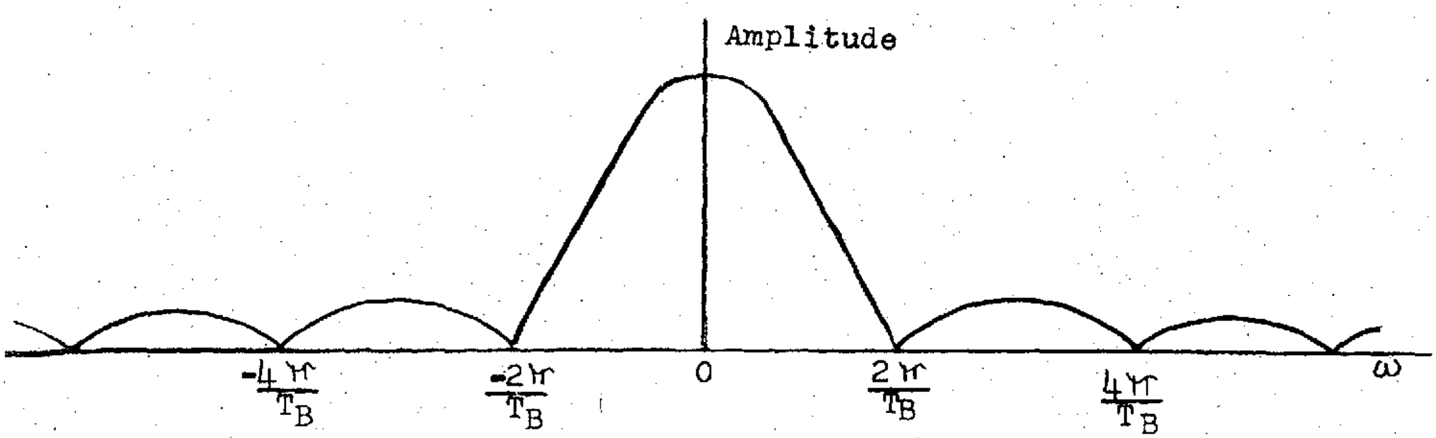

Figure 4-4 Power Spectrum of a Random Digital Sequence

The spectrum of the pseudo-random or pseudo-noise sequence has the envelope of figure 4-4 but consists of impulse functions spaced at intervals of $2\pi/(LT_B)$ radians.

If L is large, the unit impulse functions are closely packed, with L spectral components between $\omega=0$, $\omega=2\pi/T_B$.

Transponder Design

To illustrate the capability of the PN spread spectrum scheme, a typical space PN-transponder design will be reviewed. Figure 4-5 is the transponder design.

A simplified diagram of the communication capability of the transponder is given in figure 4-6.

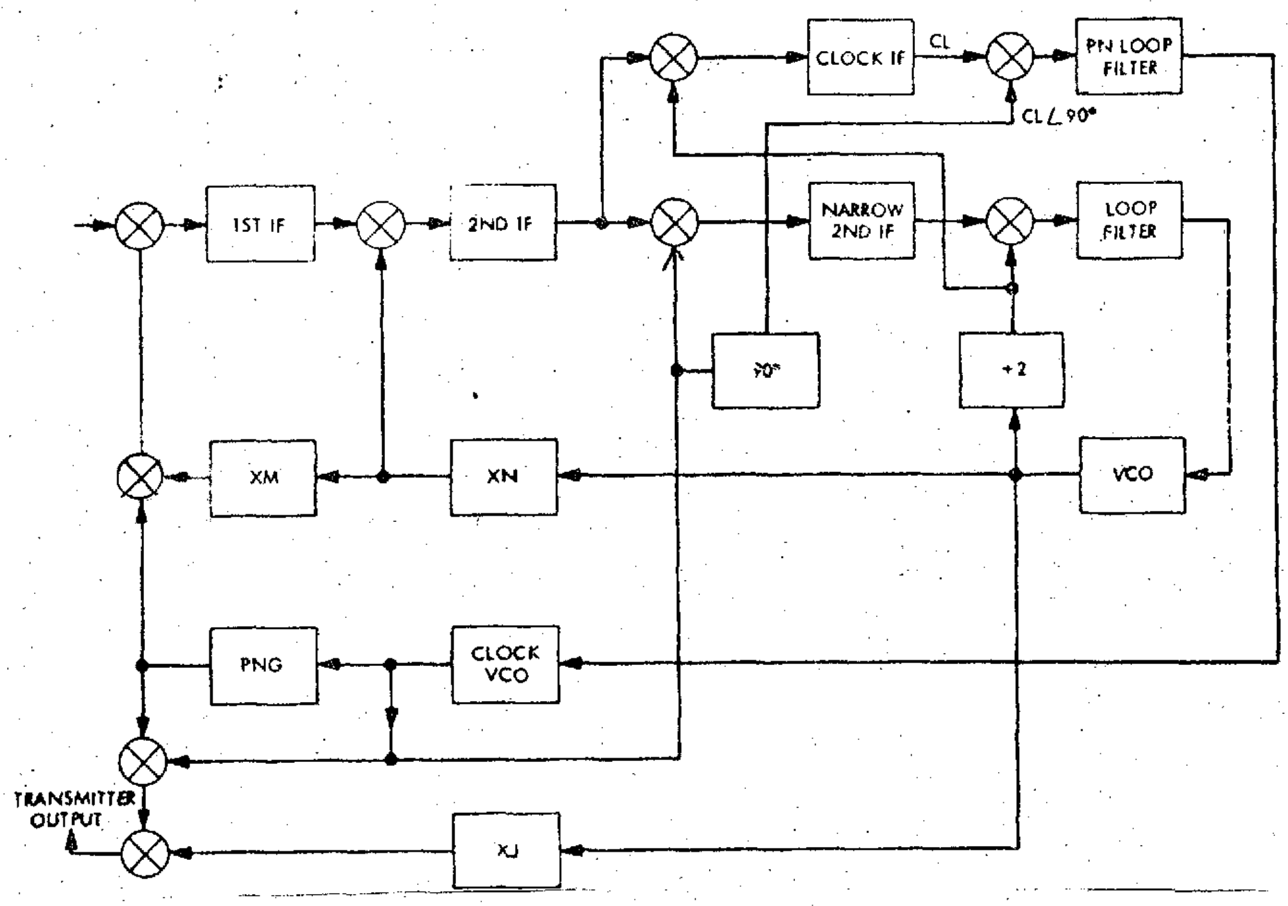

Figure 4-5 Complete PN Transponder

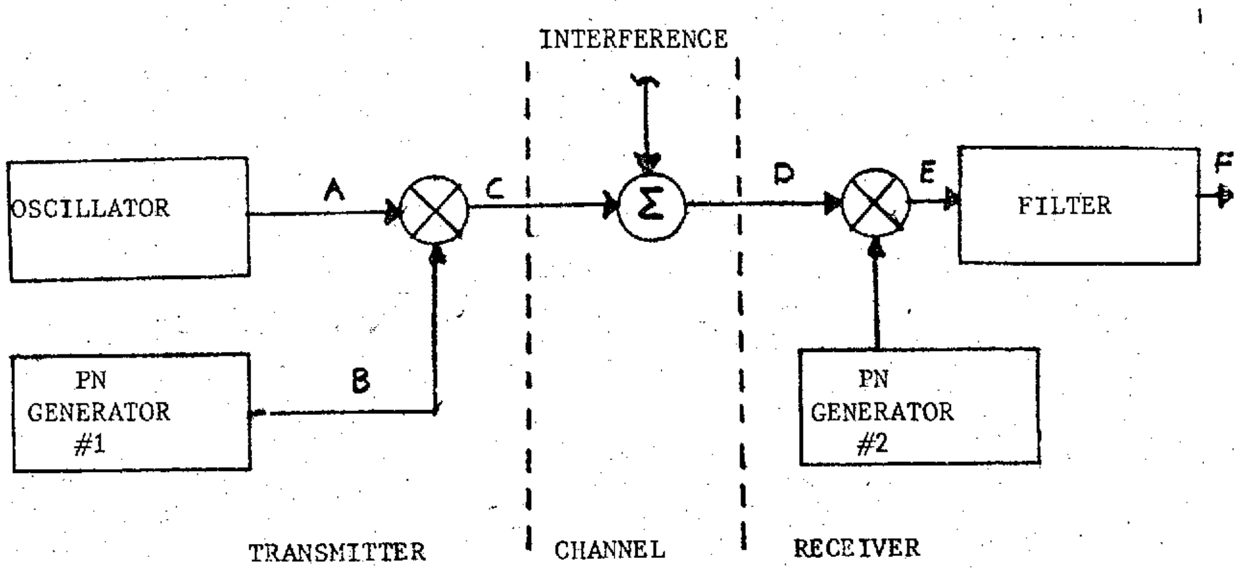

Figure 4-6 Simplified PN System Diagram

The spectrum representation of the signals at point A and B are given below, the spectrum of the signal at point C is the convolution of the signals at A and B.

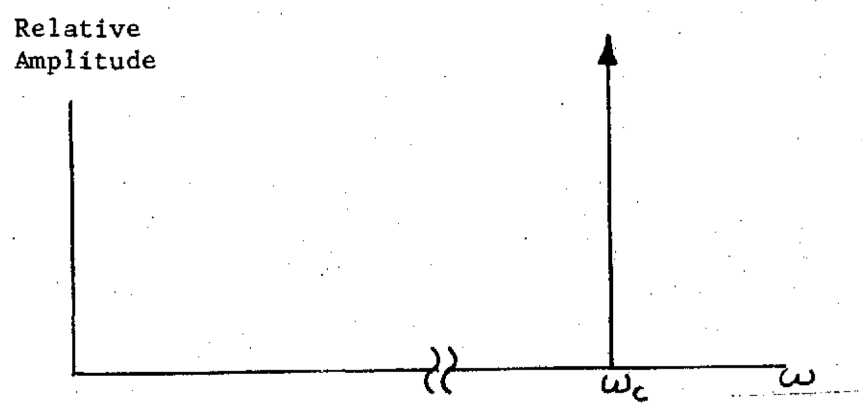

Figure 4-7 Spectrum at Point A

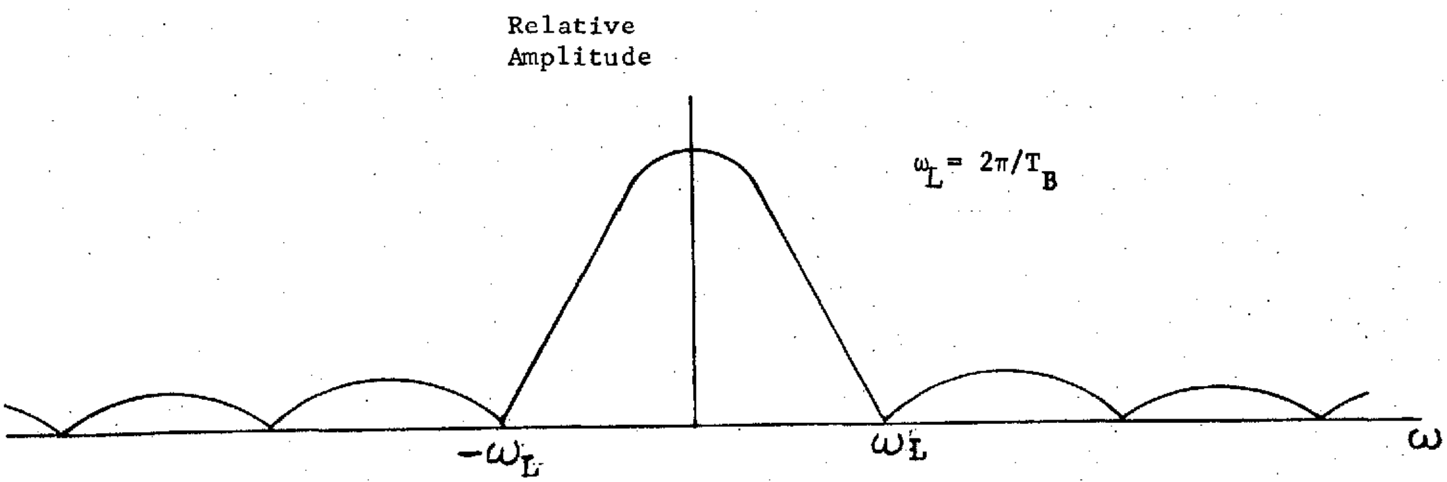

Figure 4-8 Spectrum at Point B

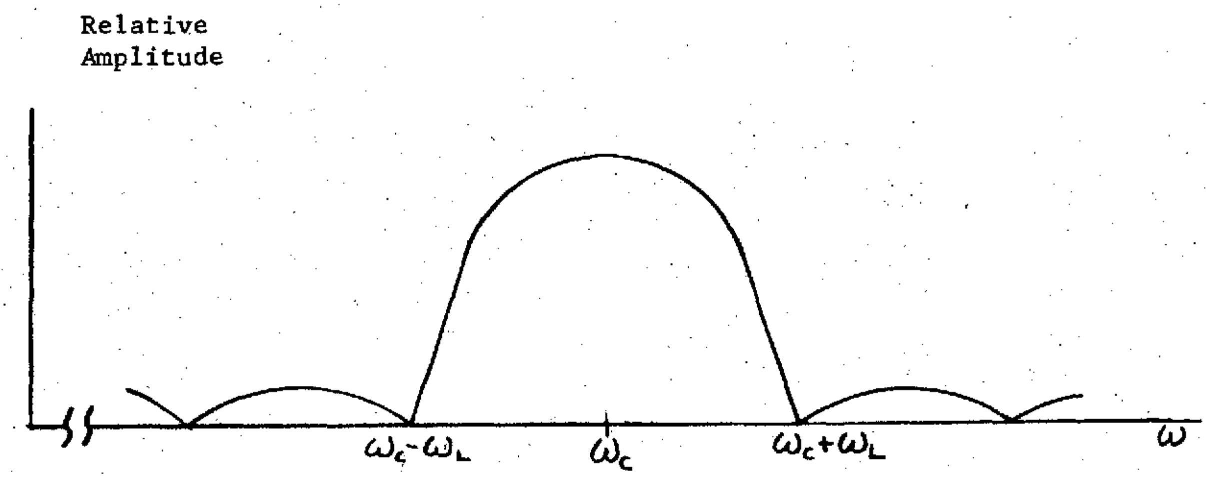

Figure 4-9 Spectrum at Point C

The spectrum point D, after additive interference in the channel is shown in figure 4-10

The interference is assured to be an additive sinusoidal at frequency ω_I radians/sec.

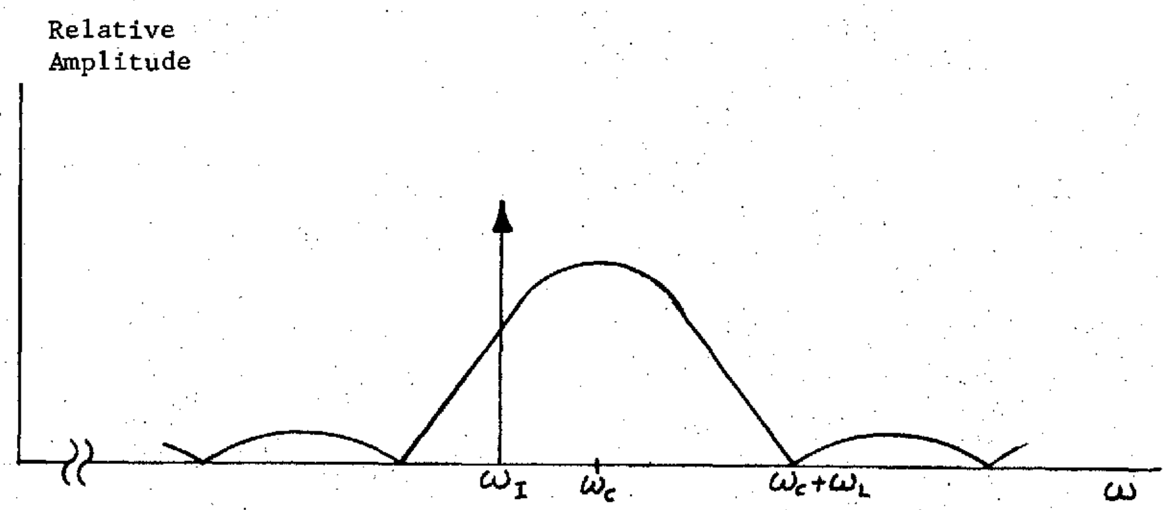

Figure 4-10 Spectrum at Point D

Assuming the PN generator in the receiver is synchronized with the PN generator in the transmitter, the spectrum representation of the signal at point E is the convolution of the PN sinc envelope with the composite signal at point D, and the results are shown in figure 4-11.

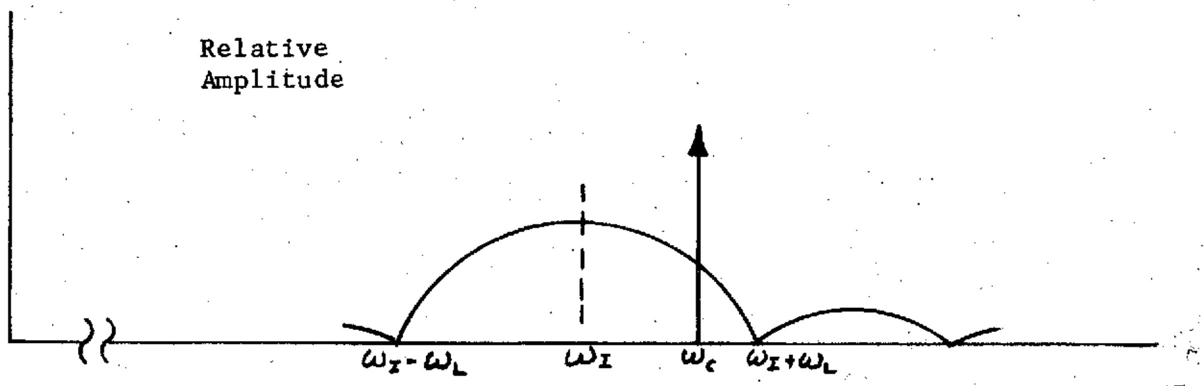

Figure 4-11 Spectrum at Point E

The filter in figure 4-6 is a band pass unit centered around ω_c and the spectrum at point F is shown in figure 4-12.

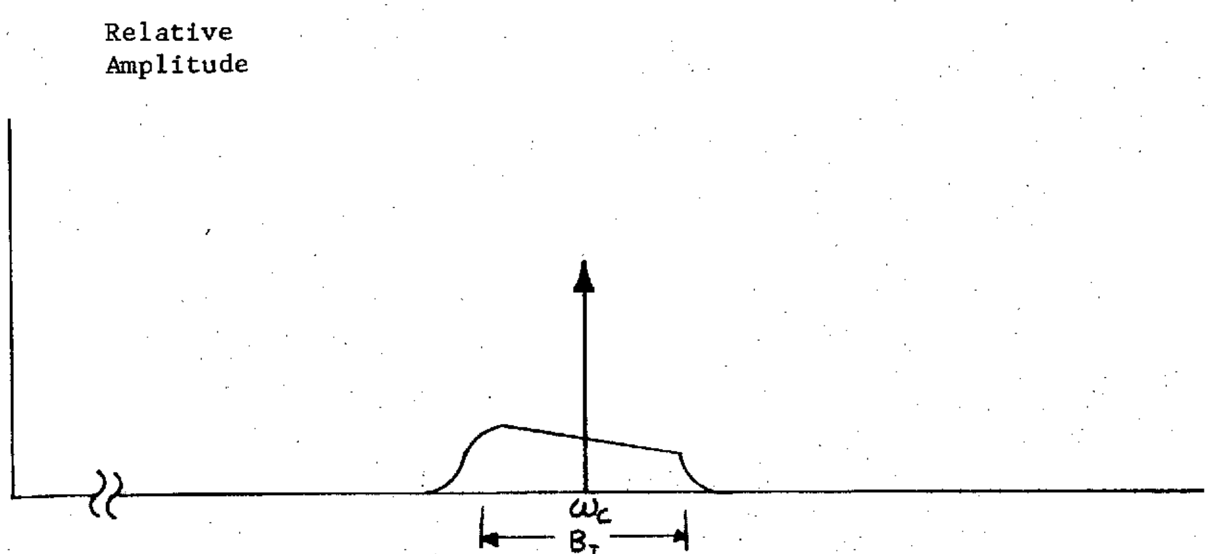

Figure 4-12 Spectrum at Point F

The interfering signal added in the channel is spread by the PN demodulation process, and only a portion of the energy associated with this interfering signal is passed to the output by the filter.

The operation of the PN modulation-demodulation scheme in Figure 4-6 requires that the PN generator and digital clock driving this generator in the receiver be synchronized with the PN generator and the clock in the transmitter. The PN transponder shown in Figure 4-5 includes the necessary circuitry to achieve this sychronization. A simplified block diagram of the transponder is shown in Figure 4-13.

The operation $\otimes$ is convolution in the frequency domain or multiplication (Corresponding to digital modulo-two addition) in the time domain. The clock phase-lock loop, the carrier phase-lock loop, and the PN generator are shown in Figure 4-14, 4-15, and 4-16 respectively.

The transponder "locked" condition means the carrier phase, the clock phase, and the PN generator must all be synchronized with the signal comprising the input signal, PN $\otimes$ CL $\otimes$ CA1. The time required for synchronization depends upon PN code lengths and loop bandwidths. For a particular TDRS-user transponder design, studies have shown a lock-up period of several minutes is required. This long period required for achieving synchronization could be excessive for some communications users, and this question deserves further investigation.

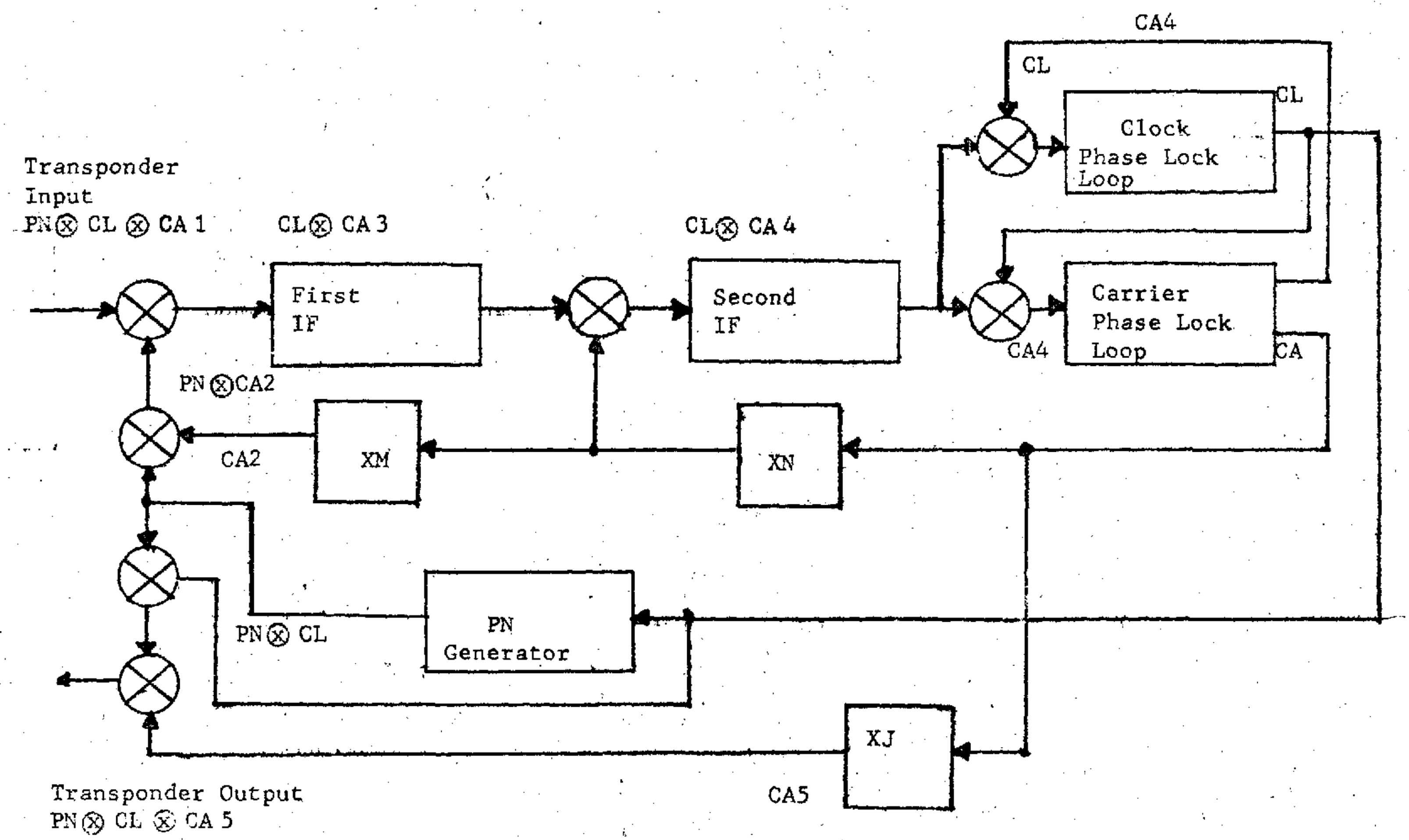

Figure 4-13 Simplified Transponder Block Diagram

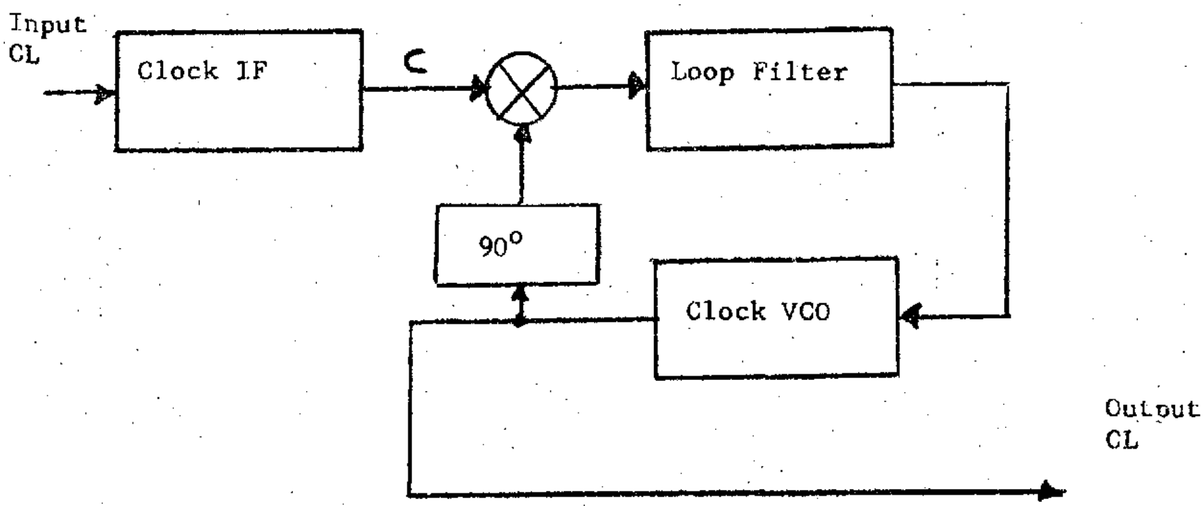

Figure 4-14 Clock Phase Lock Loop

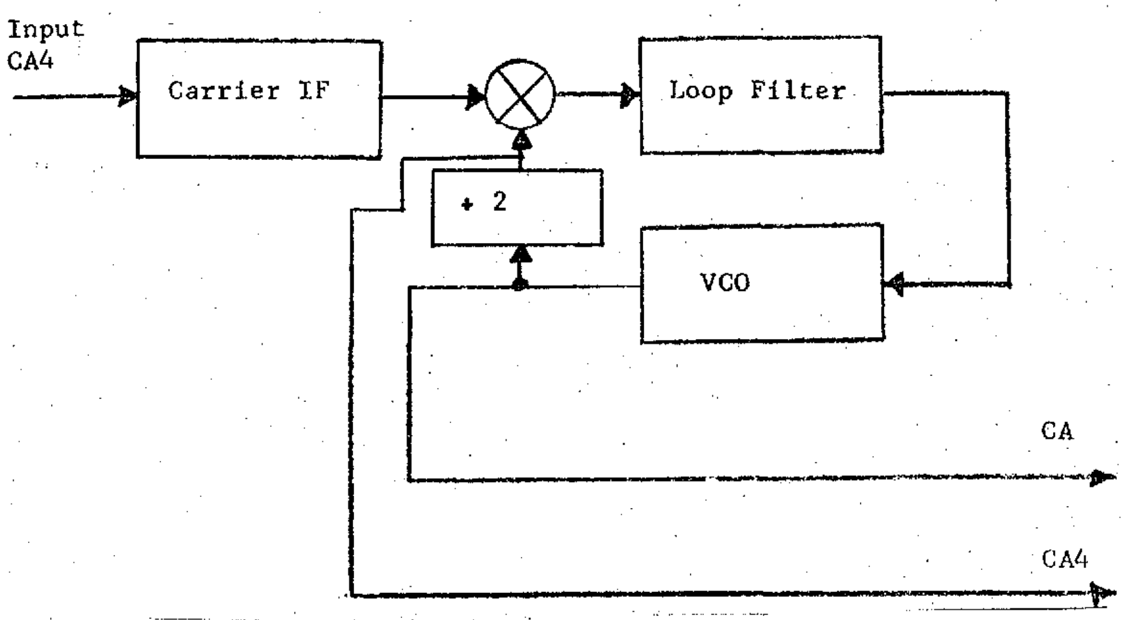

Figure 4-15 Carrier Phase Lock Loop

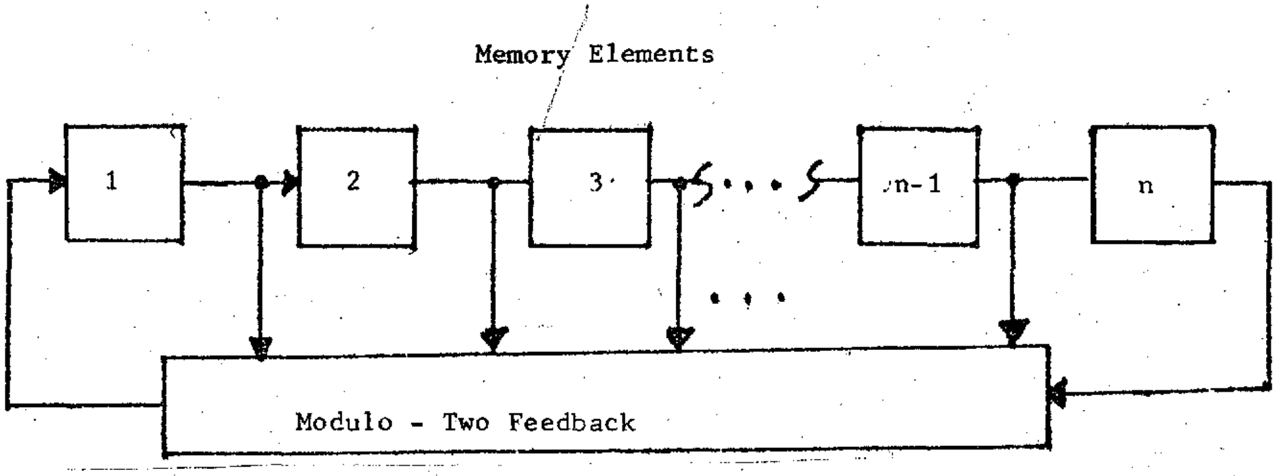

Figure 4-16 Maximum Length PN Sequence Generator

PN Parameters

In a multiple user system such as the conceptual TDRS design, the number of expected users affects the choice of PN parameters. In the TDRS design approximately 40 simultaneous users must share the assigned spectrum. It was concluded from early studies that the number of suitable PN codes was not sufficient to allow user identification by assigning a unique PN code to each user. A possible means of avoiding this problem is the use of hybrid-sum sequences rather than maximum-length sequences as the PN code. Figure 4-17 is a block diagram of a hybrid-sum sequence formed from the modulo-two sum of several maximum-length digital sequences. A recent study has shown that the statistical properties of hybrid-sum sequence are potentially better than those of maximum length sequences, and are much easier to evaluate analytically.

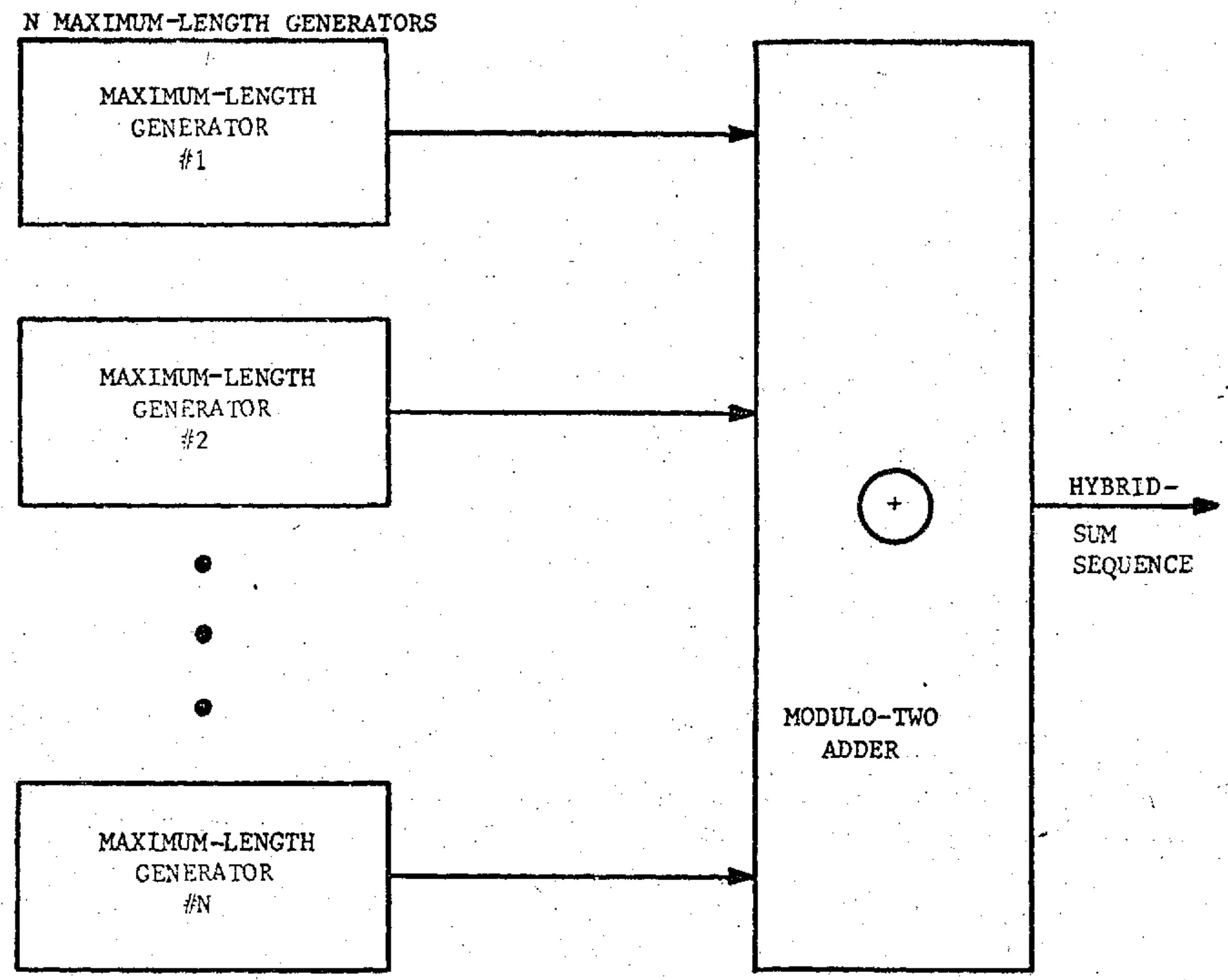

Figure 4-17 Hyrid-sum sequence generator

Currently a study is being performed to evaluate the capability of these codes to improve the lock-up and tracking performance of PN transponders for use in the TDRS-user system.

The autocorrelation functions of hybrid sum sequences for k-even and k-odd, where k is the number of ML sequences forming the HS sequence, is shown in figure 4-18. The autocorrelation function for the case k=1 (single ML sequence) is a two level function in the general case. With this assumption the clock phase-lock-loop error signal (point C, fig. 4-14) is shown in figure 4-18. As in the case of ML sequences there is a stable lock point and tracking can be achieved. The time parameter L is given as

$$L = \prod_{i=1}^{k} (2^{ni} - 1) \quad ,$$

and there are many possible values of L depending on the different selections of ni, the component generator sizes.

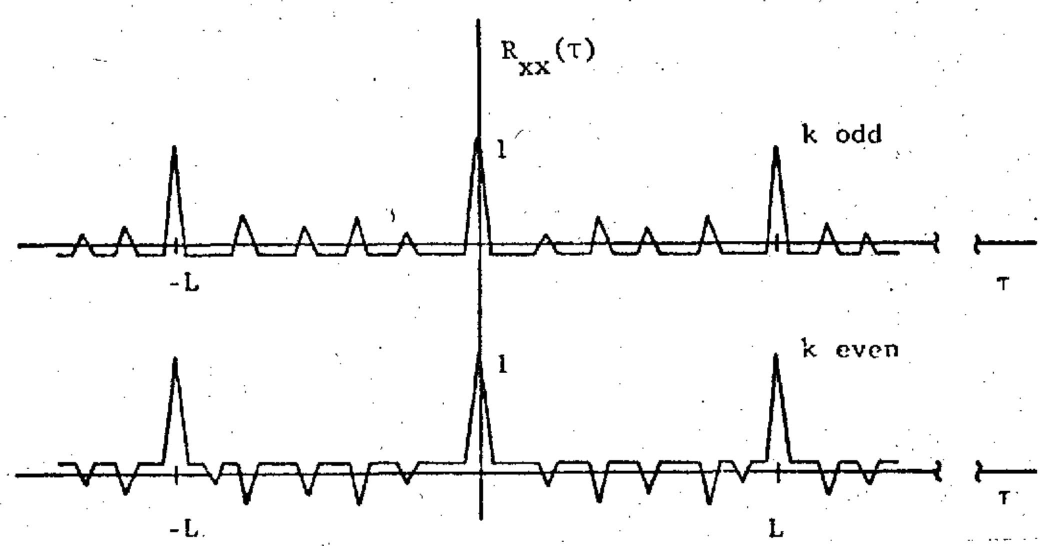

Figure 4-18. Examples of the autocorrelation function of sum sequences from k generators

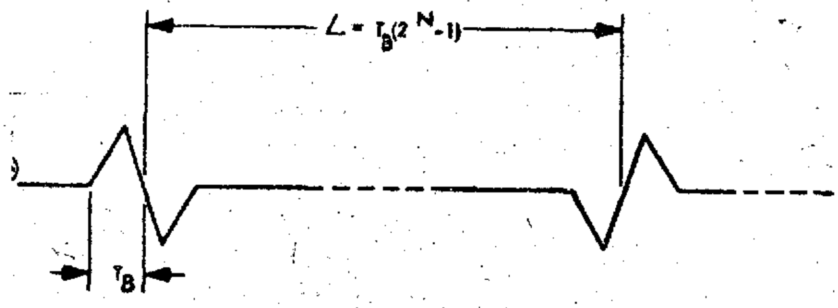

Figure 4-19 Error Signal of Clock Phase Lock Loop

The expected improved performance of the HS system is based on a preliminary design of a lock-up mode where each component ML sequence locks independently. Since the component ML sequence can be much shorter than the final HS sequence, lock up time can be reduced. Also, because of the improved statistical properties of the HS sequence, lock communication performance should be improved to some extent.

These considerations are important to TDRS users, for example, improved lock-up performance of the transponder might eliminate the need for significant user data storage and assure real-time data from user to ground over a longer period near TDRS handover. An in-depth study of the parameters of the spread-spectrum hybrid-sum transponder design as applied to the TDRS system is continuing.

Power Budget Calculations

Normal power budget calculations can be modified to calculate the expected performance of a spread-spectrum communications system in the presence of multipath.

The final contrast ratio is

$$CR = \frac{P}{kT}$$

Where P is the signal power density, k is Boltzman's Constant, and T is the effective system temperature. T can be expressed

$$T = T_r + Tm$$

where Tr is the effective receiver temperature which is calculated from the receiver noise-figure and the antenna temperature, and Tm is the effective temperature of the spread multipath component. If the information bandwidth is B_I and the PN clock rate is fc, then

$$G = \frac{fc}{B_I}$$

is the system processing gain.

If Pm is the total multipath power component of the returned signal, then the spectrum at the multipath component is spread, and

$$Pe = \frac{Pm}{B_I G}$$

is the effective power density of the spread multipath signal. The effective temperature due to the multipath component is then

$$Tm = \frac{Pm}{kGB_I}$$

The following procedure can be utilized to modify the standard contrast ratio calculations when there is a multipath signal component and a spread spectrum system is being used.

(1) Calculate the system noise temperature, TR, as the sum of the receiver noise temperature and the antenna temperature.

(2) Calculate the signal power, P_s, at the user with standard techniques.

(3) Calculate the multipath signal power at the user using

$$Pm = \alpha^2 \frac{P_s R^2}{(2d_1 + R)^2}$$

where

Ps is the signal power
R is the radius of the earth
d1 is the altitude of the user
(see fig. 1)
α is the earth reflectivity

(4) Calculate the effective multipath temperature as

$$T_m = \frac{P_m}{kGB_I}$$

where k is Boltzmans' Constant, and G is the processing gain,

$$G = \frac{fc}{BI} .$$

The calculation above must be based on the condition

$$\frac{2d_1}{3 \times 10^8} > \frac{1}{B}$$

where B is the bandwidth of the tracking filter in the PN phase lock-loop.

(5) Calculate

$$T = T_r + T_m .$$

(6) Calculate the contrast-ratio

$$CR = \frac{P_s}{kTB_I} .$$

As an example, consider a user orbit

d1 = 200 km

with a PN clock rate of 40 MHZ and an information bandwidth of 1.5 MHZ.

The analysis procedure is as follows:

(1) Assume the system noise temperature referred to the receiving antenna is $700^{o}K$.

(2) Assume the signal power at the user is - 80 dbm.

(3) The multipath signal component is

Pm = 83.6 dbm.

where

$\alpha = .707$

(4) The multipath effective temperature is

$T_m = 220K$

(5) The total effective system temperature is

$T = 920^{o}K$.

(6) The contrast ratio is

CR = 16 db

The above analysis is appropriate because

$$\frac{2d_1}{3X10^8} = 1.33 \text{ X } 10^{-3}, \frac{1}{B} = 1.024 \text{ X } 10^{-6}$$

where the bandwidth of the loop tracking filter is

B = 980 KHZ.

(b) Power Budgets and Links Performance

This section describes power budget calculations used to evaluate the proposed HEAO-C-TDRS forward and return links. The TDRS-Ground link was not considered since this properly would be part of the TDRS system design, and in any case it is not the critical link. The HEAO-C-TDRS (return) link is critical because of the required data rate. The 1Kbps TDRS-HEAO-C (Forward) link required performance margin is easily achieved if the return link parameters meet their required margins.

The following tables give the power budget calculations and form a parametric study of the link. The following lists the tables included:

TABLE	DESCRIPTION	PARAMETERS
4-1	S-BAND HEAO-C-TDRS RETURN LINK	2.025 GHZ 3 db TRANSMITTER ANTENNA GAIN
4-2	X-BAND HEAO-C-TDRS RETURN LINK	10.357 GHZ 3 db TRANSMITTER ANTENNA GAIN
4-3	Ku-BAND HEAO-C-TDRS RETURN LINK	15.200 GHZ 3 db TRANSMITTER ANTENNA GAIN
4-4	S-BAND HEAO-C-TDRS RETURN LINK	2.025 GHZ 18.8 db TRANSMITTER ANTENNA GAIN
4-5	S-BAND HEAO-C-TDRS FORWARD LINK	2.025 GHZ 3 db RECEIVER ANTENNA GAIN
4-6	Ku-BAND HEAO-C-TDRS FORWARD LINK	15.200 GHZ 3 db RECEIVER ANTENNA GAIN
4-7	S-BAND HEAO-C-TDRS RETURN LINK	2.025 GHZ 3 db TRANSMITTER GAIN, LOW RATE MODE

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The calculations for the power budgets (1-6) were made with the assumption of the HIGH RATE MODE in the return link and 1 KBPS command signal in the forward link. The power budget for case 7 is for the low rate mode in the return link. In all cases of the return link the HEAO-C transmitter power is specified as 10 watts. In all cases the TDRS-HEAO-C range is assured as 3.58×10^7 METERS. The TDRS-MDR configuration is assured to include an 8 ft. 60% effective aperture reflective antenna.

The antenna pointing loss calculations are based upon a $.3^\circ$ RMS pointing error of the TDRS-MDR (8ft) dish antenna.

Other parametric data, such as the TDRS receiver equivalent noise temperature, are taken from contractor TDRS studies.

S-BAND HEAO-C-TDRS RETURN LINK
2.025 GHZ CONFIGURATION 1

1.	Transmitter Power 10 WATT		10 dbw	
2.	Circuit Loss		-1 db	
3.	Transmitter Antenna Gain		3 db	
4.	Pointing Loss		odb	
5.	+ = 2.025 GHZ Space Loss $d=3.58\times10^{7}$ ng		-189.7	
6.	Polarization Loss		odb	
7.	8'DISH-60% Receiver Antenna Gain		32.3 db	
8.	Pointing Loss		-.3db	
9.	Net Circuit Loss		-157.4	
10.	Received Signal Power		-145.7 dbw	
11.	Thermal Noise Spectral Density TEQ = 900%		-199.0dbw/HZ	
12.	Receiver Signal Power to Thermal Noise Spectral Density		53.3 db	
13.	Channel Rate DATA 128kBPS		51.1 db BPS	
14.	Coding Gain		6 db	
15.	S/N Contrast Ratio		8.2	
16.	Required S/N For PE = 10^{-6}		10.8	
17.	Margin		-2.6	
18.	Desired Margin		10 db	
19.	70.5KBPS Allowed Channel Rate No Margin		48.5 db BPS	

X-BAND HEAO-C-TDRS RETURN LINK

10.357 GHZ CONFIGURATION 2

1.	Transmitter Power		10 dbw	
2.	Circuit Loss		-1 db	
3.	Transmitter Antenna Gain		3 db	
4.	Pointing Loss		odb	
5.	Space Loss $f= 10.357x10^9$ HZ, $d=3.58x10^7$ m		-203.7 db	
6.	Polarization Loss		odb	
7.	Receiver Antenna Gain 8' DISH -60%		46.5 db	
8.	Pointing Loss .3°		- 2.3 db	
9.	Net Circuit Loss		-157.2 db	
10.	Received Signal Power		-148.5 dbw	
11.	Thermal Noise Spectral Density TEQ= 900°		-199.0 dbw/HZ	
12.	Receiver Signal Power to Thermal Noise Spectral Density		50.5 db	
13.	Channel Rate Data 128K BPS		51.1 db BPS	
14.	Coding Gain		6 db	
15.	S/N Contrast Ratio		5.4 db	
16.	Required S/N For $PE=10^{-6}$		10.8 db	
17.	Margin		- 5.4 db	
18.	Desired Margin		10 db	
19.	Allowed Channel Rate No Margin 37K BPS		45.7 db BPS	

KU-BAND HEAO-C-TDRS RETURN LINK 15.2 GHZ CONFIGURATION 3				
1.	Transmitter Power low		10 dbw	
2.	Circuit Loss		-1 db	
3.	Transmitter Antenna Gain		3 db	
4.	Pointing Loss		odb	
5.	Space Loss $f=15.2x10^9$, $d=3.58x10^7 m$		-207.1	
6.	Polarization Loss		odb	
7.	Receiver Antenna Gain 8'DISH -60%		49.8 db	
8.	Pointing Loss $.3^o$		-6.18 db	
9.	Net Circuit Loss		-161.5 db	
10.	Received Signal Power		-151.5 dbw	
11.	Thermal Noise Spectral Density $TEQ=900^o$		-199.0 dbw/HZ	
12.	Receiver Signal Power to Thermal Noise Spectral Density		47.5 db	
13.	Channel Rate 128K BPS		51.1 db BPS	
14.	Coding Gain		6 db	
15.	S/N Contrast Ratio		2.4	
16.	Required S/N for $PE=10^{-6}$		10.8 db	
17.	Margin		-8.4	
18.	Desired Margin		10 db	
19.	Allowed Channel Rate No Margin No MARGIN 18.5K BPS		42.7 db BPS	

S-BAND HEAO-C-TDRS RETURN LINK

2.025 GHZ CONFIGURATION 4

1.	Transmitter Power 10 WATT		10 dbw	
2.	Circuit Loss		-1 db	
3.	30% 24-5.2 Transmitter Antenna Gain		18.8 db	
4.	Pointing Loss 1.5°		- .2 db	
5.	Space Loss		-189.7 db	
6.	Polarization Loss		odb	
7.	8' DISH - 60% Receiver Antenna Gain		32.3 db	
8.	Pointing Loss		-.3 db	
9.	Net Circuit Loss		-140.1 db	
10.	Received Signal Power		-130.1 dbw	
11.	Thermal Noise Spectral Density		-199.0 dbw/HZ	
12.	Receiver Signal Power to Thermal Noise Spectral Density		68.9 db	
13.	Channel Rate DATA 128K BPS		51.1 dbBPS	
14.	Coding Gain		6 db	
15.	S/N Contrast Ratio		23.8 db	
16.	Required S/N for $PE=10^{-6}$		10.8 db	
17.	Margin		13 db	
18.	Desired Margin		10 db	
19.	Allowed Channel Rate No Margin 2.56 MBPS		64.1 dbBPS	

S-BAND HEAO-C-TDRS FORWARD LINK
2.025 GHZ CONFIGURATION 5

1.	Transmitter Power 17.7 watts		12.5 dbw	
2.	Circuit Loss		- 1 db	
3.	8' DISH -60% Transmitter Antenna Gain		32.3 db	
4.	Pointing Loss		-.3 db	
5.	Space Loss		-189.7 db	
6.	Polarization Loss		odb	
7.	Receiver Antenna Gain		3 db	
8.	Pointing Loss		-.1 db	
9.	Net Circuit Loss		-155.8 db	
10.	Received Signal Power		-143.3 dbw	
11.	Thermal Noise Spectral Density		-199.5 dbw/HZ	
12.	Receiver Signal Power to Thermal Noise Spectral Density		56.2 db	
13.	Channel Rate DATA 1K BPS		30 dbBPS	
14.	Coding Gain		odb	
15.	S/N Contrast Ratio		26.2 db	
16.	Required S/N for $PE=10^{-9}$		12.8 db	
17.	Margin		13.4 db	
18.	Desired Margin		10 db	
19.	Allowed Channel Rate No Margin		-	

KU-BAND HEAO-C-TDRS FORWARD LINK CONFIGURATION 6				
1.	Transmitter Power		-2 dbw	
2.	Circuit Loss		-1 db	
3.	8'DISH - 60% Transmitter Antenna Gain		49.8 db	
4.	Pointing Loss .3°		-6.2 db	
5.	Space Loss		-207.1 db	
6.	Polarization Loss		odb	
7.	Receiver Antenna Gain		3 db	
8.	Pointing Loss		-.1 db	
9.	Net Circuit Loss		-161.6 db	
10.	Received Signal Power		-163.6 db	
11.	Thermal Noise Spectral Density		-195.1 dbw/HZ	
12.	Receiver Signal Power to Thermal Noise Spectral Density		31.5 db	
13.	Channel Rate Data 1KBPS		30 db BPS	
14.	Coding Gain		6 db	
15.	S/N Contrast Ratio		7.5 db	
16.	Required S/N for $P_E = 10^{-9}$		12.8 db	
17.	Margin		-5.3 db	
18.	Desired Margin		10 db	
19.	Allowed Channel Rate No Margin		——	

S-BAND HEAO-C-TDRS RETURN LINK
2.025 GHZ CONFIGURATION 7

1.	Transmitter Power		10 dbw	
2.	Circuit Loss		-1 db	
3.	Transmitter Antenna Gain		3 db	
4.	Pointing Loss		odb	
5.	Space Loss		-189.7 db	
6.	Polarization Loss		odb	
7.	Receiver Antenna Gain		32.2 db	
8.	Pointing Loss		-.3 db	
9.	Net Circuit Loss		-155.7 db	
10.	Received Signal Power		-145.7 dbw	
11.	Thermal Noise Spectral Density		-199.0 dbw/HZ	
12.	Receiver Signal Power to Thermal Noise Spectral Density		53.3 db	
13.	Channel Rate 19.2 KBPS		42.8 dbBPS	
14.	Coding Gain		odb	
15.	S/N Contrast Ratio		10.5 db	
16.	Required S/N		- .3 db	
17.	Margin		10 db	
18.	Desired Margin		—	
19.	Allowed Channel Rate No Margin		—	

Several observations can be made from these power budget calculations. These include:

1. For the HEAO-C configuration with low-gain transmitting antenna, the high data rate mode cannot be supported at S-band, X-band, or Ku-band.

2. For the HEAO-C configuration with high-gain transmitting antenna (18.8db), the high data rate mode can be supported with a 13db margin at S-band.

3. For the HEAO-C configuration with low-gain transmitting antenna, the low data rate mode cannot be supported at S-band without coding. Even with coding the margin would be only 5.7 db (assuming a 6 db coding gain).

4. For the forward or command TDRS-HEAO-C link, the S-band configuration will provide the 1KBPS link with a 13.4 db margin. The Ku-band configuration cannot support the link with the required data rate, link margin, and error probability, even with error control coding. The specifications of TDRS transmitter power for the S-band and Ku-band links are from a contractor's (NAR) TDRS configuration.

5. Proposed HEAO-C Modifications

The proposed modifications in the design of the HEAO-C telecommunications system includes error control coding, a spread spectrum modem for TDRS relay, and a command pointing controlled S-band phased array antenna. Specific problems involved in the spread spectrum modem design are discussed in sections 4,6, and 7. This section discusses the error control coding scheme and the HEAO-C antenna system.

(a) Error Control Coding

A previous study of possible coding schemes for digital data in a satellite relay communications link has concluded that convolutional encoding in conjunction with soft-decisionViterbi decoding gives favorable performance gain with minimum increased hardware complexity.

Figure 5-1 is a result of a computer simulation of a rate 1/3, constraint length 7 convolutional coding scheme with a 3-bit soft decision Viterbi decoder. As can be seen from the figure, at a bit error rate of 10^{-5}, a 6 db coding gain results with coded ideal coherent PSK as compared with uncoded ideal coherent PSK. Figure 5-2 is a diagram of a rate 1/3, constraint length 7 convolutional encoder.

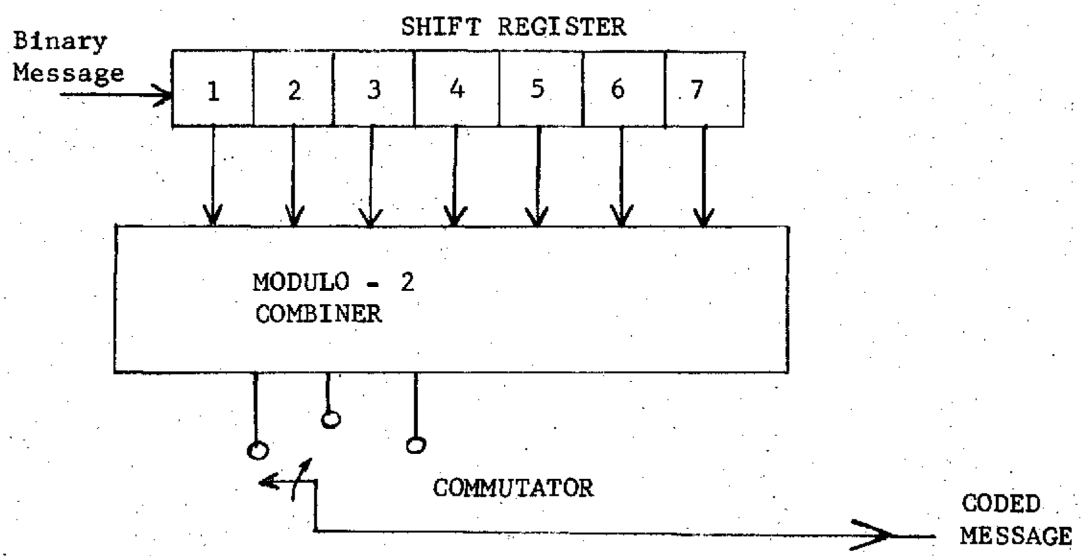


Figure 5-2 K = 7, V = 3 Convolutional Encoder

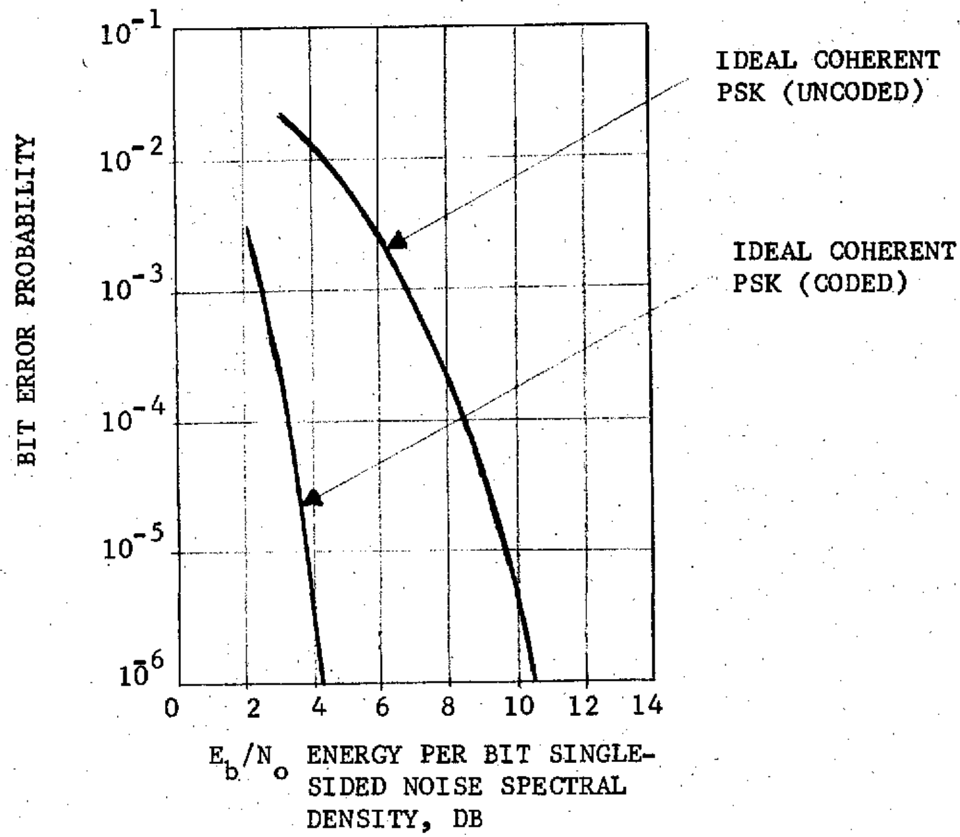


Figure 5-1 - Simulation results for $K = 7$, $V = 3$ convolutional encoding/soft-decision Viterbi decoding

The modulo-2 combiner forms a modulo-2 combination of selected register stages to form each of the three commutator nodes. This can be expressed as

$$G_i = (g_{1i}, g_{2i}, \ldots g_{7i}), \tag{5-1}$$

and

$$C_i = \sum_{j=1}^{7} g_{ji} X_j \qquad \text{MOD-2} \tag{5-2}$$

where X_j is the contents of the jth shift register stage and g_{ji} is 0 or 1 depending upon whether the jth stage contributes, modulo-2, to the ith commutator pole.

The operation of the encoder is as follows: The binary message may be much larger than the constraint length. The first bit of the message is switched into the shift register, whose other stages are logical zero, and a complete cycle of the commutator is made. The next bit of the sequence is switched into the register, the initial bit shifted to register stage two and another synchronous cycle of the commutator is made. Using the synchronous shift and cycle procedure the message sequence is encoded. At the end of the binary message seven zeros are attached, and when they are shifted into the register and accompanying code generated by the commutator, the shift register is in the all zero state once more. For an L - bit message,

$$L_c = 3(L+6) \tag{5-3}$$

bits from the coded message.

Decoding may be accomplished by sequential or Viterbi algorithms. The sequential decoding method may be described as a tree searching procedure, the exact details depending upon which particular algorithm is being used. The decoding procedure is best described by example, K = 7 is large for the purpose of an example, so a K = 4, V = 3 example is given.

The tree structure for a K = 4, V = 3 truncated code is shown in figure 5-4. The encoder for the code is shown in figure 5-3.

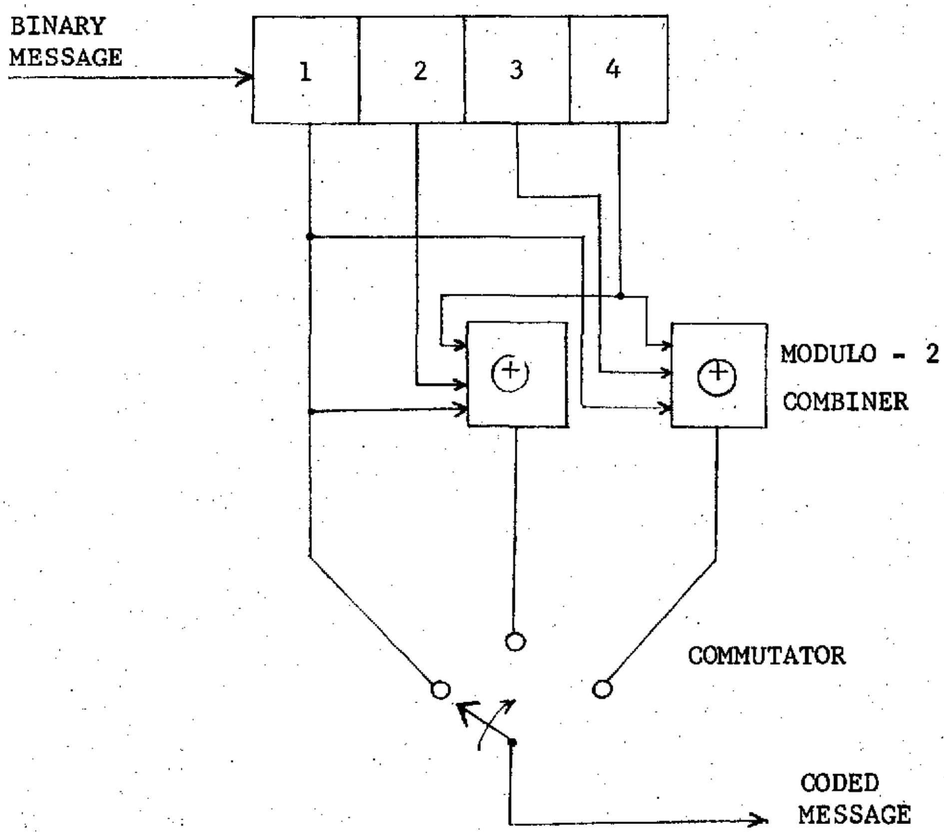


Figure 5-3 Encoder for tree structure of figure 5-4, K=4, V=3.

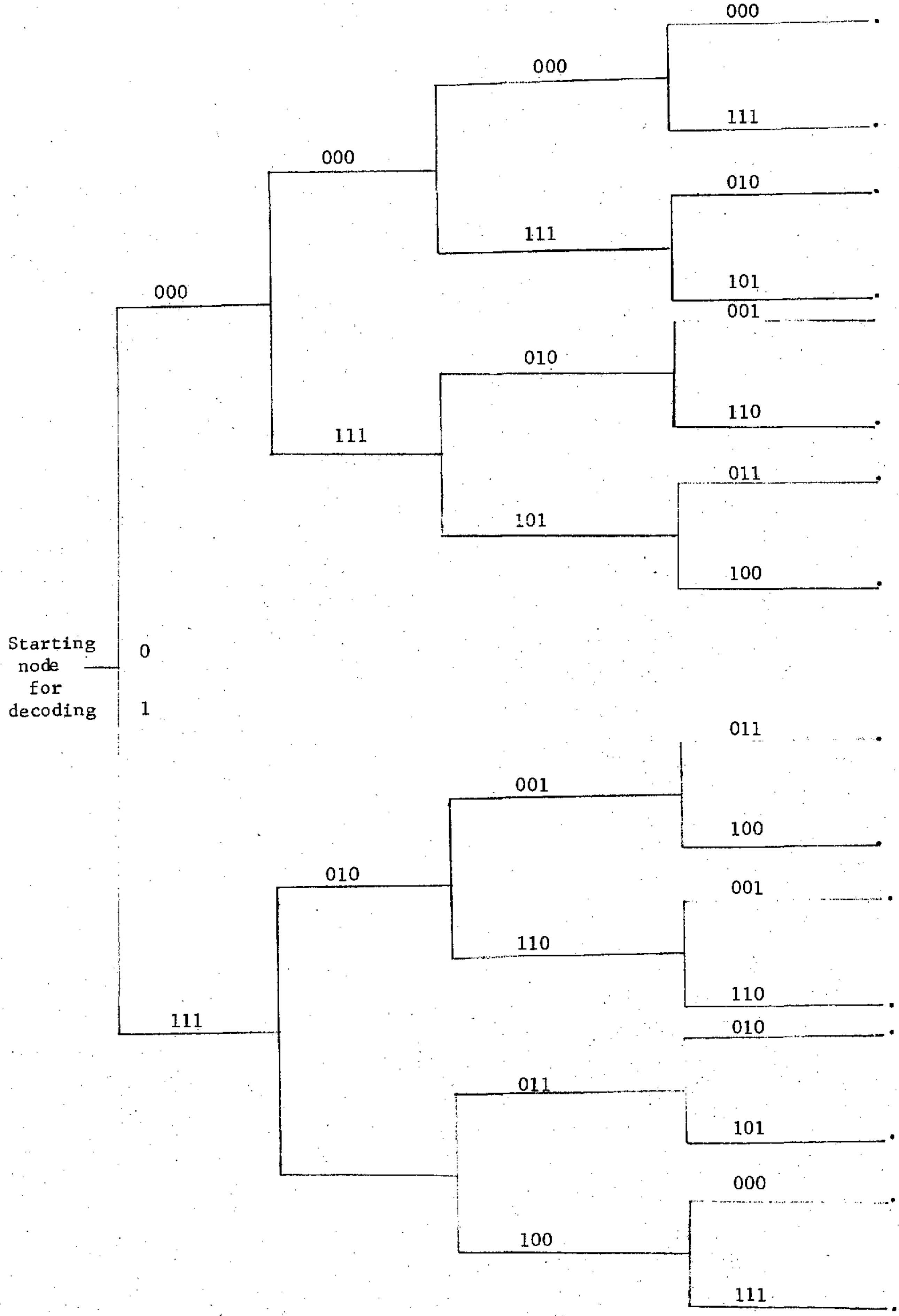


Figure 5-4 Tree Structure for K=4 V=3 truncated code

As an example assume the message

$$X=(1011) \tag{5-4}$$

is to be transmitted. The encoder of figure 5-3 provides the coded message,

$$Y=(111,010,110,110). \tag{5-5}$$

Assuming the channel introduces the noise

$$N=(100,101,000,010), \tag{5-6}$$

the received code is

$$R=(011,111,110,100). \tag{5-7}$$

The sequential decoder will form the quantity

$$d_i = w\left[R_i \oplus Y_i\right], \tag{5-8}$$

where i represents the ith three bit sequence, w represents the weight function, and d_i is called the Hamming distance. The decoder makes each decision at each mode of the code tree based on minimizing the Hamming distance. However the decisions are tentative, and if the decoder finds in successive steps that it has probably made a wrong bit decision it is able to backtrack and try another branch of the code tree.

In the example, the decoded message would begin

$$C = 11.. \tag{5-9}$$

the initial decision for the second bit being made in error. Proceeding down the error branch however significantly large values of d_i are encountered. Backtracking and trying the

$$C=10.. \tag{5-10}$$

branch gives significantly smaller values of d_i on successive steps.

The decoder algorithm is based on monitoring the statistical properties of the sum of d_i as the decoder proceeds into the code tree. If the sum of the d_i terms approaches a buildup rate of $\frac{1}{2}V$ then the decoder declares an error and backtracks to a new branch.

Expected buildup of the d_i sum for the correct branch is PV where P is the channel transision probability for the binary symmetric channel. The branch decision criteron is buildup somewhere between V/2 and PV. The decoder keeps track of the branches it has explored and avoids needless retracing of any branch.

(b) Command Pointing Controlled Phased Array Antenna

The proposed phased array antenna for the HEAO-C-TDRS return link will be a command pointing type with pointing commands formulated and communicated from the ground. An example of a phased array airborne steerable antenna system that would be applicable for this application was developed by Texas Instruments Incorporated for NASA under contract NAS8-25847. The antenna is an 128-element spiral array and achieved the performance parameters listed below.

Subsystem Performance	Value
Antenna (2282 MHz)	
Boresight gain (dB)	23.9
60-degree scan gain (dB)	20.3
Boresight axial ratio (dB)	0.3
60-degree scan axial ratio (dB)	2.0
Weight (pounds)	6.48
Boresight sidelobe level (dB)	19.5
60-degree scan sidelobe level (dB)	9.0
Module	
Noise figure (dB)	6.0
Receive gain (dB)	24
Diplexer isolation (dB)	35
Peak phase shifter phase error (degrees)	10
RMS phase shifter phase error (degrees)	5.2
Phase shifter amplitude error (dB)	0.5
Phase linearity (degrees)	5
Power output (dBW)	0.5
Transmit gain (dB)	19.0
Transmit efficiency (percent)	25
Weight (pounds)	0.275
Transmit Manifold (128-Element)	
Peak phase error (degrees)	± 5.25
Peak amplitude variation (dB)	± 0.6
Peak output VSWR (Ratio:1)	1.65
Loss (dB)	3.2
Weight (pounds)	6.48

These performance parameters were used in the TDRS-HEAO-C power budget calculations.

The block diagram of the HEAO-C communications system with error control coding and phased array antenna implementations is shown in figure 5-5.

The TDRS-HEAO-C forward link is established on the low gain (near isotropic) antenna. Pointing control commands are coded and the return forward links are established with the phased array.

The current HEAO-C, NON-TDRS communication system block diagram is shown in figure 5-6.

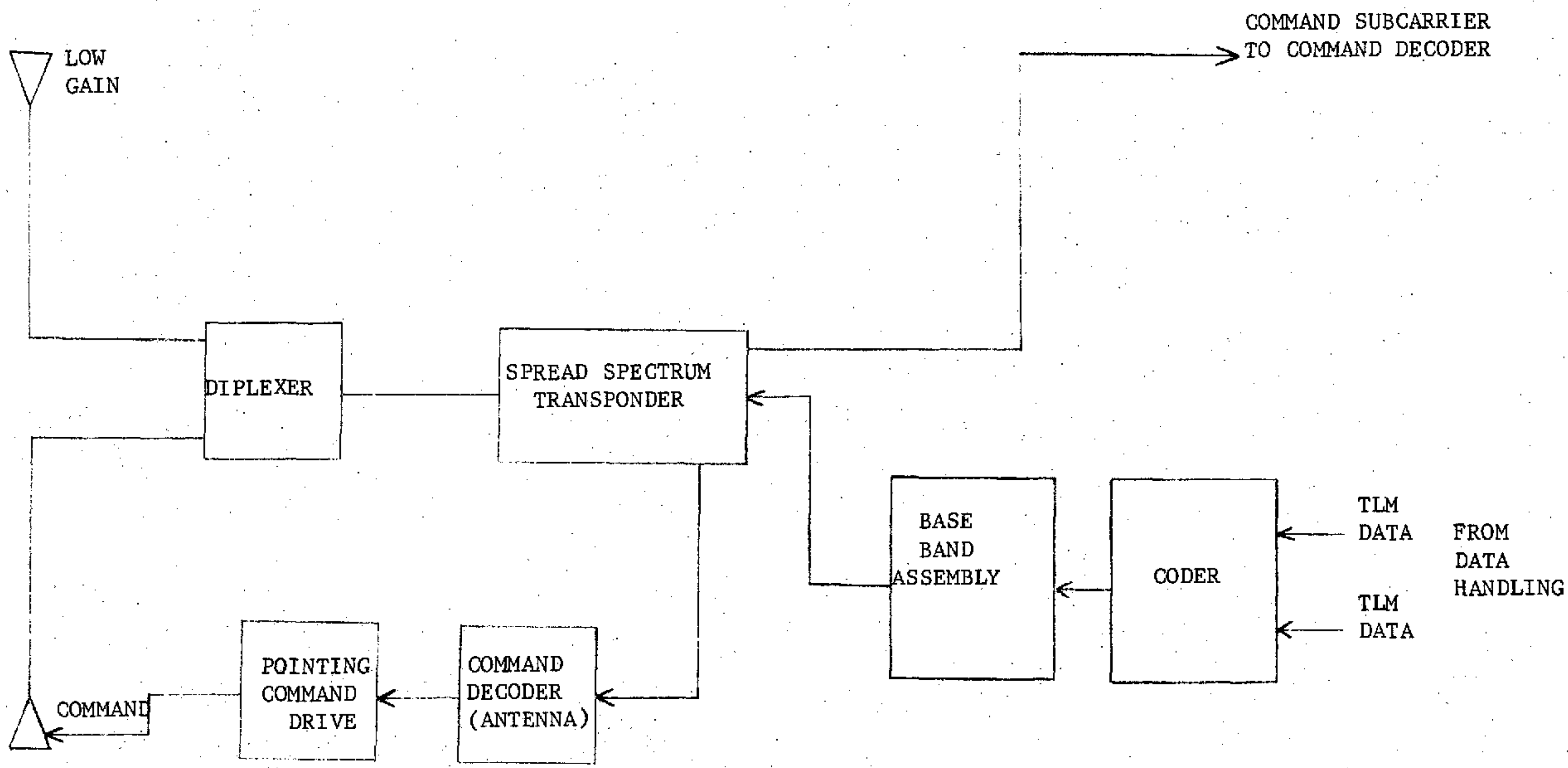


Figure 5-5 HEAO-C Communication System Block Diagram

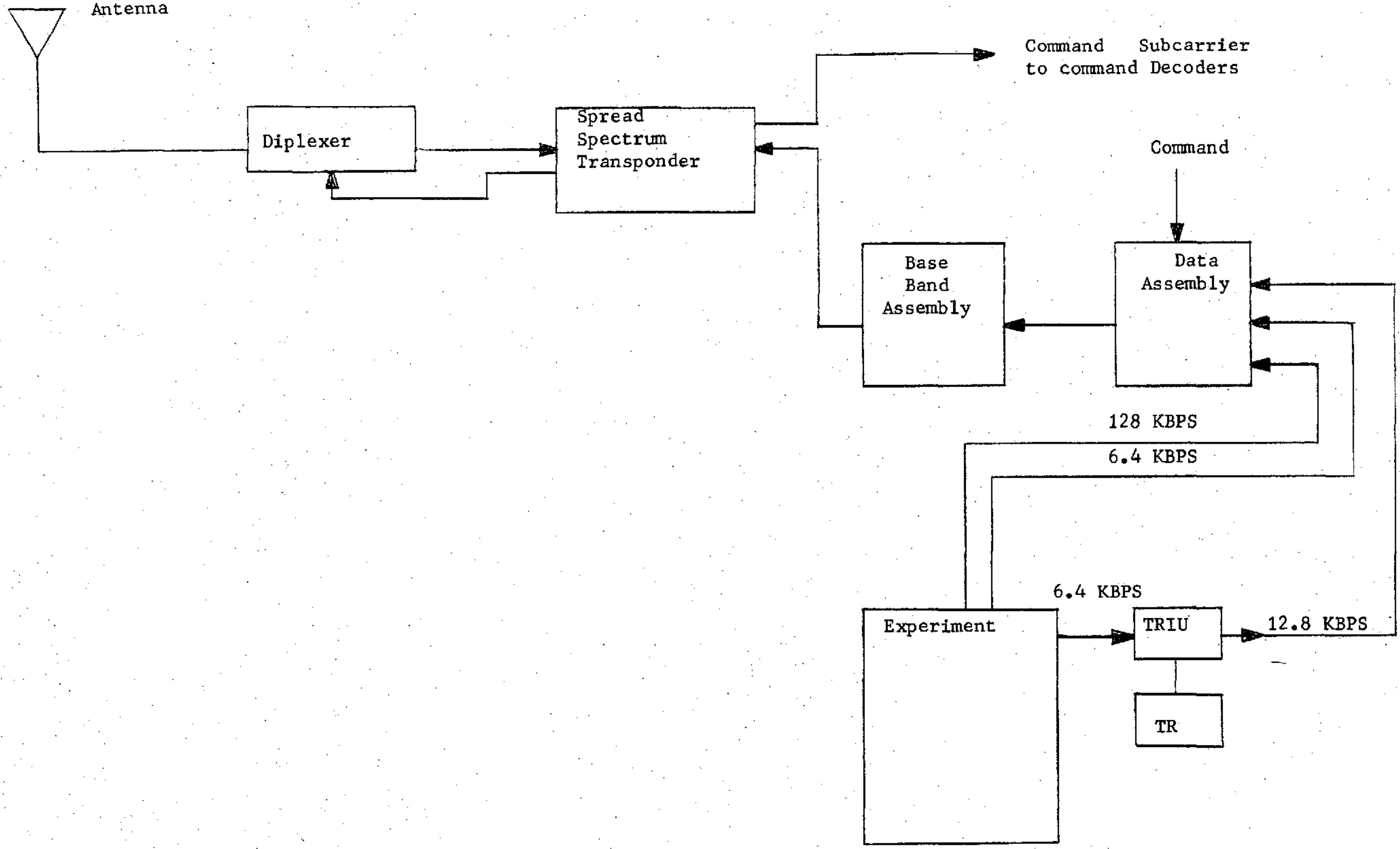


Figure 5-6 HEAO-C,NON - TDRS, COMMUNICATION
SYSTEM BLOCK DIAGRAM

6. CODE INTERFERENCE

This section describes the problem of code interference in a spread spectrum receiver. The spread spectrum system uses a wide-band modulating signal to provide the spectrum spreading, and the receiver performs a despreading operation. The spread-despread operation results in a signal processing gain relative to correlated or coherent interference such as multipath or jammer interference. Code interference results when several spread spectrum systems share the same channel. This condition exists in the proposed TDRS-multiple user communications configuration.

The code interference problem is of concern during the two modes of spread-spectrum modem operation.

1. CODE AND CARRIER ACQUISITION
2. CODE TRACKING AND DATA DESPREADING

The result of code interference during the two modes is manifested in different forms of systems performance degradation. During acquisition, code interference results in increased false lock probability, and increased expected acquisition time. During tracking and data despreading, code interference results in data contrast-ratio degradation and increased probability of lost code lock.

For spread spectrum systems operating with small margins, such as the proposed TDRS-USER configurations, the effect of code interference is an important consideration.

Theapproach in the analysis of the code interference problem was to model the process in such a way that a statistical moment analysis could be made. Also, with a slightly different model, a computer simul -

ation of the process was developed and evaluated.

Figure 6-1 is an overall diagram of a spread-spectrum receiver. The local PN sequence generator provides the signal that despreads PN⊗CL⊗CA. Figure 6-2 is a PN sequence acquisition scheme. During both the tracking and acquisition modes, an interfering code input can be modeled as shown in figure 6-3. The filtering operation essentially forms the weighted sum of the last M bits of the local PN_A sequence convolved with the interfering code PN_B. For the analysis it will be assumed that

$$CL_B \neq CL_A \tag{6-1}$$

so that the two sequences slide past each other in the filter convolution operation. In the analysis codes A and B could be the same, in which case the code self noise would be evaluated. The first step in the analysis is to synthesizethe code sliding operation. Figure 6-4 shows possible models for three bit codes A and B.

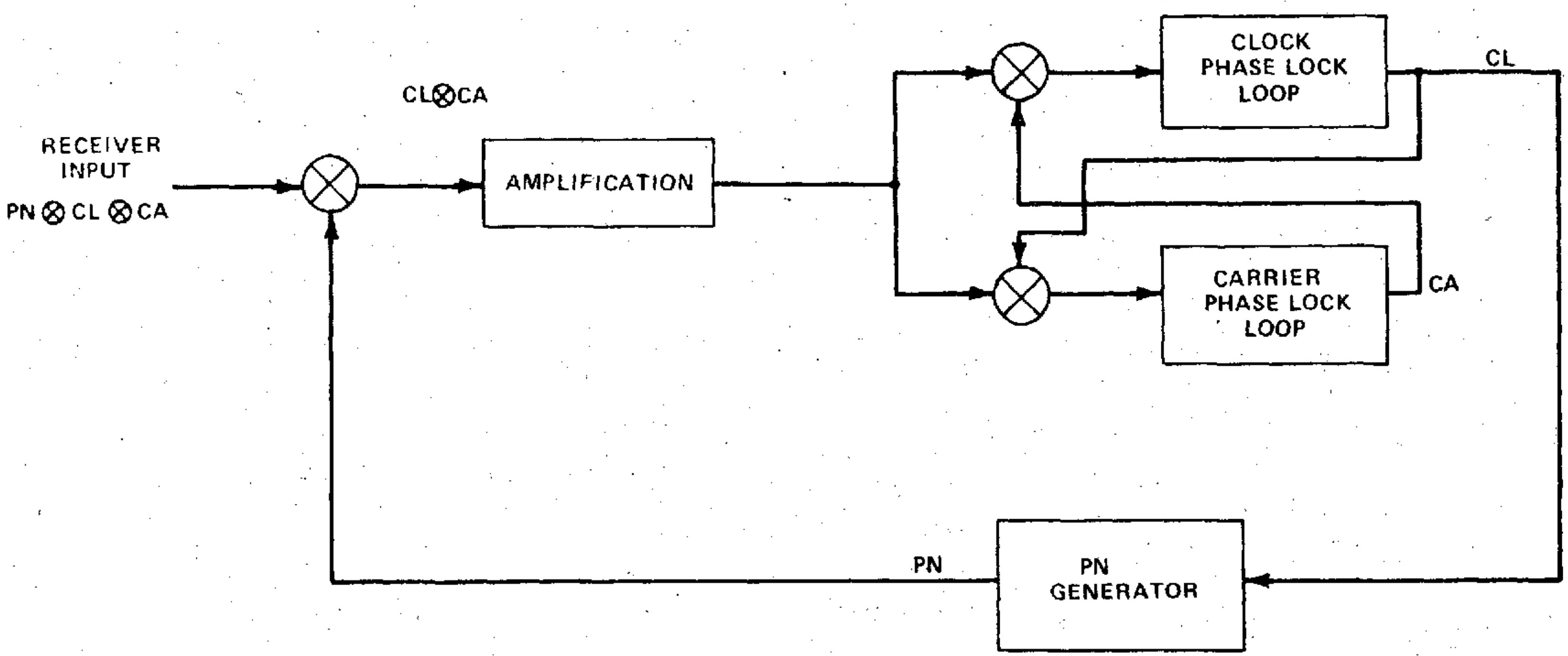


FIGURE 6-1

SIMPLIFIED SPREAD SPECTRUM RECEIVER

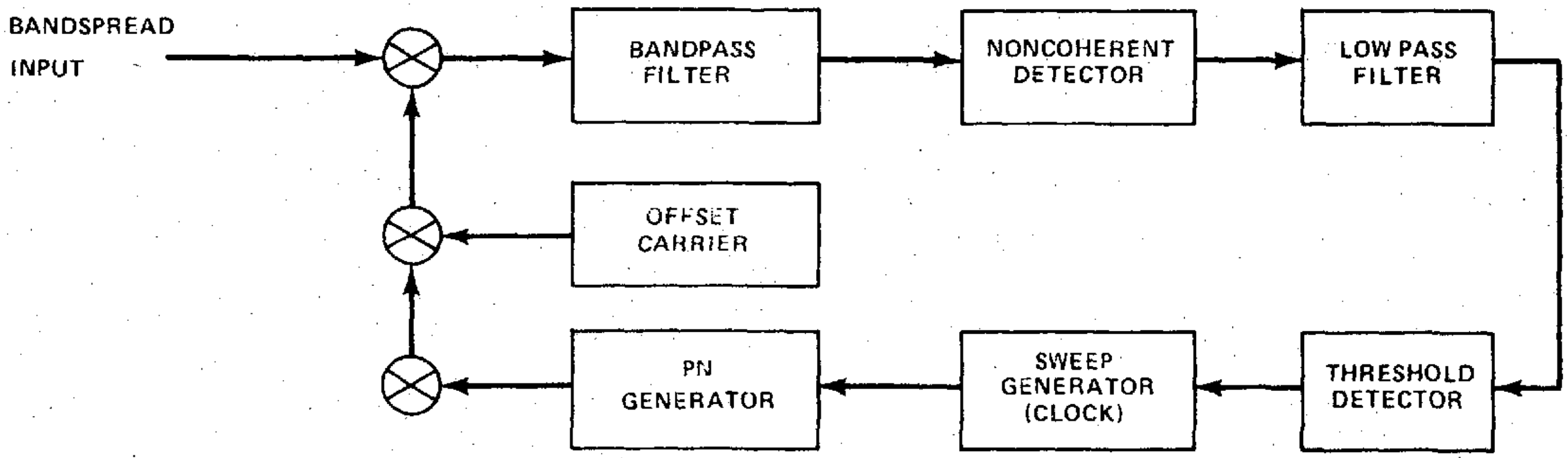


FIGURE 6-2
PN ACQUISITION SYSTEM

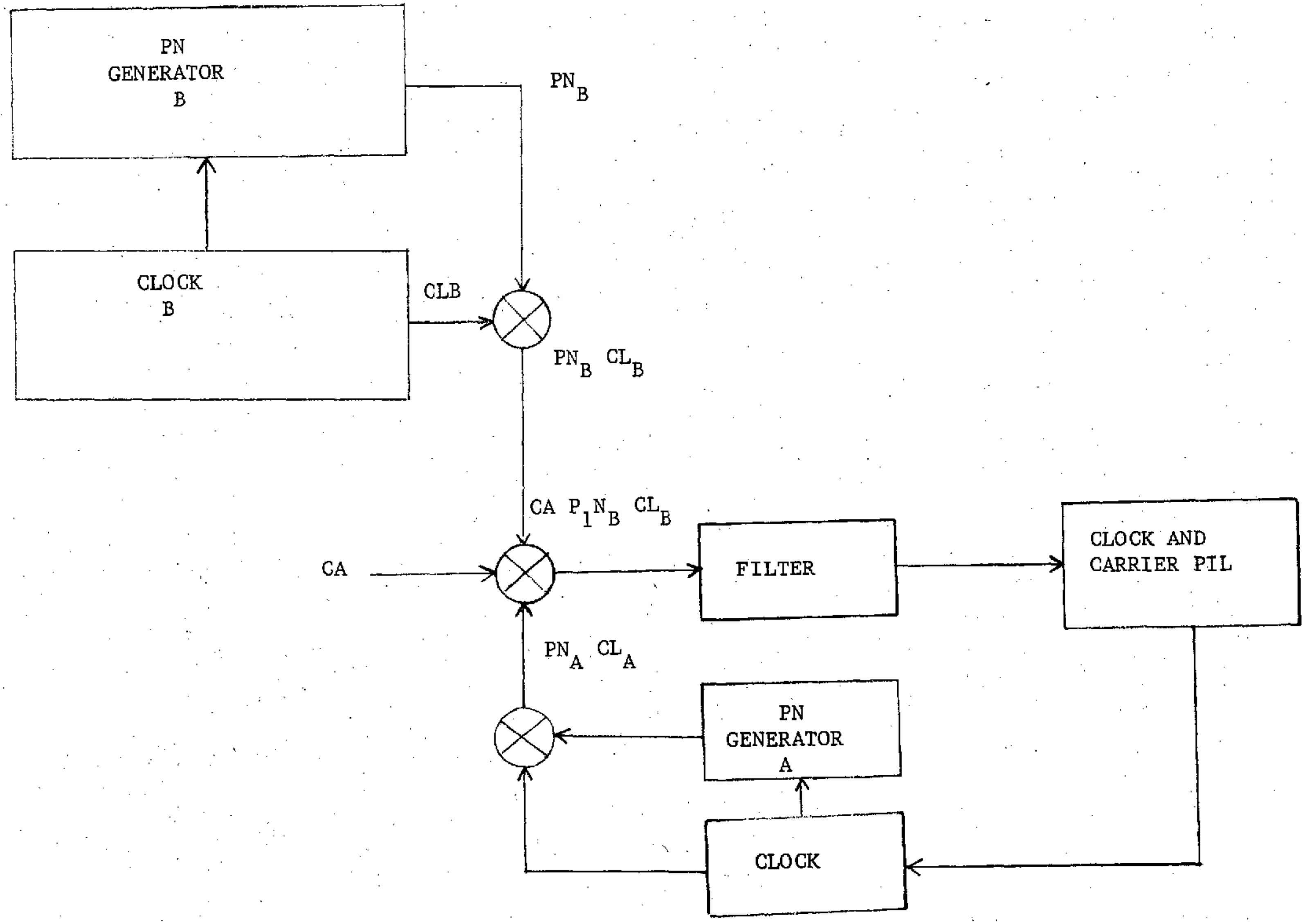
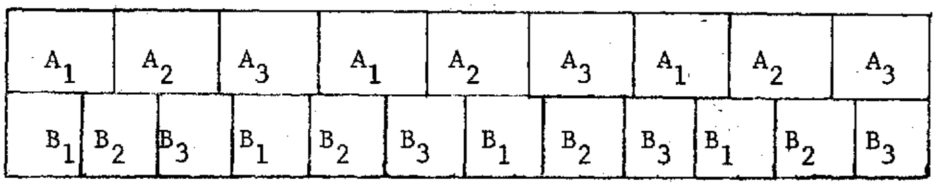


Figure 6-3 Code Interference System Model Operations



MODEL 1

A_1	A_2	A_3	A_1	A_2	A_3	A_1	A_2	A_3
B_1	B_2	B_3	B_2	B_3	B_1	B_3	B_1	B_2

MODEL 2

A_1	A_1	A_2	A_3	A_3	A_1	A_2	A_2	A_3
B_1	B_2	B_3	B_1	B_2	B_3	B_1	B_2	B_3

MODEL 3

Figure 6-4 Possible Code Models

The first model represents the realistic situation, in most cases, where code A and code B have unequal bit intervals $(CL_A \neq CL_B)$. The unequal bit intervals of model 1 present problems in the analysis, so models 2 and 3 are constructed with equal bit intervals. In model 2, sequence, B skips one bit, on a rotating basis, for each L_B bits, where L_B is the B-sequence length. In model 3, sequence A "hangs-up" for an extra clock count on one particular bit during each L_A bits, where L_A is the A-sequence length. The particular "hang-up" bit occurs on a rotating basis. This is equivalent to a skipped clock count in the A generator for each L_A-1 clock counts. Models 2 and 3 approximate the situation here $CL_A \neq CL_B$.

The analysis is to be based on Model 3.

Another relevant model, the one used in the computer simulation in the program COSIF, is to consider the code phase states represented in vector form as shown for the three bit sequence in figure 6-5.

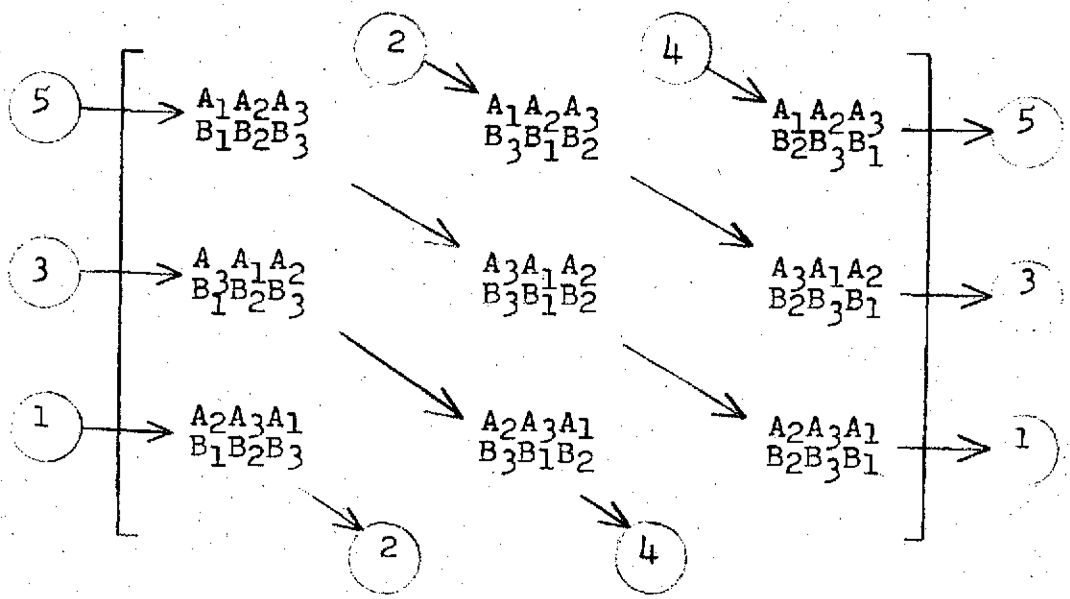


Figure 6-5

The time dependence of the model is represented as a trajectory through the matrix, indicated by the horizontal and diagonal line segments. The diagonal line segments represent a variation in this normal sequence of phase states caused by a droppedclock bit in generation A.

The forms of the filter output is slightly different for the two approaches used in the analysis and the computer simulation COSIF. This is illustrated by the sequence of filter-outputs shown in figure 6-6.

MODEL 3	COSIF MODEL
A_3 A_1	A_3 A_1
B_3 B_1	B_3 B_1
A_2 A_3	A_2 A_3
B_2 B_3	B_2 B_3
A_2 A_2	A_2 A_3
B_1 B_2	B_1 B_2
A_1 A_2	A_1 A_2
B_3 B_1	B_3 B_1
A_3 A_1	A_3 A_1
B_2 B_3	B_2 B_3
A_3 A_3	A_3 A_1
B_1 B_2	B_1 B_2
A_2 A_3	A_2 A_3
B_3 B_1	B_3 B_1
A_1 A_2	A_1 A_2
B_2 B_3	B_2 B_3
A_1 A_1	A_1 A_2
B_1 B_2	B_1 B_2

FIGURE 6-6

Notice in the above there are three bit differences in the two models. In general there would be L_A bit differences in $L_A L_B$ phase combinations. For large L_B this difference is minimal. The approximation of reality in each case is based on the fact that on the average $CL_A \neq CL_B$ even though they are equal for certain segments of the sequence. Therefore model 3 and the COSIF model approximate the realistic model 1.

In order to study the code interference properties of maximum length codes, hybrid sum codes, and Gold codes, the code generators are assumed to be structured as shown in figure 6-6. Code generator A is formed as

$$A_i = \prod_{\gamma=1}^{k_A} X_{i,\gamma} \qquad (6-2)$$

where k_a sub generators form generator A. Likewise, code generator B is formed as

$$B_i = \prod_{\gamma=k_A+1}^{k} X_{i,\gamma} \qquad (6-3)$$

where

$$k = k_A + k_B \qquad (6-4)$$

The statistical moments of the code cross correlation (interference) function can be expressed as

$$S^p = \frac{1}{L} \sum_{i=o}^{L-1} \left(\sum_{j=o}^{M-1} \prod_{\gamma=1}^{k} X_{i+j,\gamma} \right)^p \qquad (6-5)$$

where

$$L = L_A L_B \qquad (6-6)$$

is guaranteed by the condition

$$CL_A \neq CL_B$$

Now using model 3, it will be assumed that over the course of L phase states, generator A drops L_B clock bits. Therefore sequence A is a modified form of the sequence formed from the same sub-generators, but with no dropped clock bits. One important characteristic of model 3 should be noted, that is, every possible single bit phase combination A and B are cycled over L.

Evaluation of 6-5 for P=1 yields

$$S^1 = \frac{1}{L} \sum_{j=o}^{M-1} \left(\sum_{i=o}^{L_A-1} \prod_{\gamma=1}^{k_A} X_{i+j,\gamma} \right) \left(\sum_{i=o}^{L_B-1} \prod_{\gamma=k_A+1}^{k} X_{i+j,\gamma} \right) \tag{6-8}$$

Now

$$\sum_{i=o}^{L_A-1} \prod_{\gamma=1}^{k_A} X_{i+j\gamma} = \prod_{\gamma=1}^{k_A} \sum_{i=o}^{L_\gamma-1} X_{i+j\gamma} = (-1)^{k_A} \tag{6-9}$$

if code A is hybrid or maximum length. Otherwise

$$\sum_{i=o}^{L_A-1} \prod_{\gamma=1}^{k_A} X_{i+j\gamma} = A_o - A_1 \tag{6-10}$$

where Ao is the number of zeroes in the A-sequence, and A_1 is the number of ones in the A-sequence. Similar equations exist for the B-sequence.

Now define the quantity

$$H_A = \sum_{i=o}^{L_A-1} \prod_{\gamma=1}^{k_A} X_{i+j\gamma} = \begin{cases} (-1)^{k_A} & \text{HYBRID OR ML CODES} \\ [A_o-A_1] & \text{GOLD CODES} \end{cases} \tag{6-11}$$

and

$$S^1 = \frac{M}{L} H_A H_B \tag{6-12}$$

The second moment can be expressed as

$$S^2 = M + \frac{2}{L} \sum_{j=o}^{M-2} \sum_{\zeta=j+1}^{M-1} \sum_{i=o}^{L-1} \prod_{\gamma=1}^{k} X_{i+j,\gamma} X_{i+\zeta,\gamma} \tag{6-13}$$

or

$$S^2 = M + \frac{2}{L} \sum_{j=o}^{M-2} \sum_{\zeta=j+1}^{M-1} \sum_{i=o}^{L-1} \alpha i{+}i\ \alpha i{+}\zeta\ bi{+}i\ bi{+}\zeta\ . \qquad (6\text{-}14)$$

Now first consider the case of M_L or HS sequences

If $\zeta \neq j+1$

$$\sum_{i=o}^{L-1} \alpha i{+}j\ \alpha i{+}\zeta\ bi{+}j\ bi{+}\zeta = \sum_{i=o}^{L_A-1} \prod_{\gamma=1}^{k_A} X_{i+Z\gamma} \sum_{i=o}^{L_B-1} \prod_{\gamma=1}^{k_B-1} X_{i+\zeta,\gamma} \qquad (6\text{-}15)$$

If $\zeta=j+1$

$$\sum_{i=o}^{L-1} \alpha i{+}j\ \alpha i{+}\zeta\ bi{+}j\ bi{+}j$$

$$= \sum_{i=o}^{L_A-1} \sum_{\Upsilon=1}^{k_A} X_{i+Z,\gamma} \sum_{i=o}^{L_B-1} \prod X_{i+Z,\gamma} \mp H_A \pm L_A \qquad (6\text{-}16)$$

The second moment for the case becomes

$$S^2 = M + \frac{1}{L} (M^2 - 2M + 4)\ H_A H_B + \frac{2}{L} (M-1)(H_A H_B \pm (L_A - H_A)) \qquad (6\text{-}17)$$

Notice from equation 6-17, for hybrid and maximum length codes, the second moment is independent of particular codes, but only depends on M, k_A, k_B and L_A, and L_B.

For the case of codes other than ML of HS equation (6-15) becomes for $\zeta \neq j+1$

$$\sum_{i=o}^{L-1} \alpha i{+}j\ \alpha i{+}\ \ bi{+}j\ bi{+}\zeta = \sum_{i=o}^{L_A-1} \prod_{\gamma=1}^{k_A} X_{i+s,\gamma} \sum_{i=o}^{L_A-1} \prod_{\gamma=1}^{k_A-1} X_{i+s\ \gamma} \qquad (6\text{-}18)$$

where it is assumed that code B is the Gold sequence. The * indicates that code B will be modified as a result of the operation bits to a code form from the same family as B. Equation (6-19) becomes

$$S^2 = M + \frac{1}{L}(M^2-2M+4)\ H_A H_B^* + \frac{2}{L}(M-1)(H_A H_B^* \pm (L_A - H_A)). \tag{6-19}$$

H^*_B might be any H_B from the family of codes of which B is a member.

For the case in which code A and code B are both Gold sequences, and for the case $\zeta \neq j+1$

$$\sum_{i=o}^{L-1} a_{i+j}\, a_{i+\zeta}\, b_{i+j}\, b_{i+\zeta} = H_A^* H_B^* \tag{6-20}$$

Equation (19) becomes

$$S^2 \approx M + \frac{1}{L}(M^2-2M+4) H_A^* H_B^* + \frac{2}{L}(M-1)(H_A^* H_B^* \pm (L_A - H_A^*)) \tag{6-21}$$

The third moment can be expressed

$$S^3 = \frac{1}{L}\sum_{i=o}^{L-1}\left[\sum_{j=o}^{M-1}\prod_{\gamma=1}^{k} X_{i+j},\right]^3$$

$$= \frac{1}{L}\sum_{i=o}^{L-1}(3M-2)\sum_{j=o}^{M-1}\prod_{\gamma=1}^{k} X_{i+j},$$

$$= \frac{1}{L}\sum_{i=o}^{L-1} 3!\sum_{j=o}^{M-3}\sum_{V=j+1}^{M-2}\sum_{\zeta=V+1}^{M-1}\prod_{\gamma=1}^{k} X_{i+j},\ X_{i+v},\ X_{i+} \tag{6-22}$$

or

$$S^3 = \frac{3M-2}{L} \sum_{j=0}^{M-1} \sum_{i=0}^{L-1} \prod_{\gamma=1}^{k} X_{i+j,\gamma}$$

$$+ \frac{3}{L} \sum_{j=0}^{M-3} \sum_{v=j+1}^{M-2} \sum_{\zeta=v+1}^{M-1} \sum_{i=0}^{L-1} \prod_{\gamma=1}^{k} X_{i+j,\gamma} X_{i+v,\gamma} X_{i+\zeta,\gamma} \tag{6-23}$$

Now the consider the case of ML or HS sequence. If $V \neq j+1$ and $\zeta \neq V+1$

$$\sum_{i=0}^{L-1} a_{i+j}\, a_{i+v}\, a_{i+\zeta}\, b_{i+j}\, b_{i+v}\, b_{i+\zeta} =$$

$$\sum_{i=0}^{L-1} a_{i+j}\, a_{i+\phi}\, b_{i+j}\, b_{i+\theta} \tag{6-24}$$

Now if $\phi \neq j,\ \theta \neq j,\ \phi \neq j+1,\ \theta \neq j+1$

$$\sum_{i=0}^{L-1} a_{i+j}\, a_{i+\phi}\, b_{i+j}\, b_{i+\theta} = H_A H_B \tag{6-25}$$

If $\phi \neq j$, $\theta \neq j+1$ for a particular γ $\theta \neq j$, $\theta \neq j+1$ for all other γ

$$\sum_{i=0}^{L-1} a_{i+j}\, a_{i+\phi}\, b_{i+j}\, b_{i+\theta} = H_A L\, (-1)^{K_B - 1} \tag{6-26}$$

$$\text{If } \left.\begin{array}{l} \phi = j \\ \phi = j \\ \phi \neq j+1 \end{array}\right\} \begin{array}{ll} \text{for a particular } \gamma & \theta \neq j \\ \text{for all other} & \theta \neq j+1 \end{array}$$

$$\sum_{i=o}^{L-1} \alpha i+j \ \alpha i+\phi \ bi+j \ bi+\theta = H_B L_j (-1)^{k_A - 1} \qquad (6\text{-}27)$$

If $\phi = j+1$ for a particular
$\phi = j$ for all other $\theta \neq j$
$\phi \neq j+1$ $\theta \neq j+1$

$$\sum_{i=o}^{L-1} \alpha_{i+j} \ \alpha_{i+\phi} \ bi+j \ bi+\theta = H_A H_B \pm (L - H_A) \qquad (6\text{-}28)$$

Now if

$V \neq j+1, \zeta = V+1$

and

$\theta \neq j \quad \phi \neq j$

$\theta \neq j+1 \quad \phi \neq j+1$

We get

$$\sum_{i=o}^{L-1} \alpha_{i+j} \ \alpha_{i+\theta} \ bi+j \ bi+\phi = H_A H_B \pm I \qquad (6\text{-}29)$$

where

$$I = \begin{cases} 2 \\ 0 \\ -2 \end{cases}$$

depending on the particular sequence.

For

$\theta \neq j \quad \phi = j$ for a particular j

$\theta \neq j+1 \quad \left.\begin{array}{l} \phi \neq j \\ \phi \neq j+1 \end{array}\right\}$ for all other j

we get

$$\sum_{i=o}^{L-1} a_{i+j}\, a_{i+\theta}\, b_{i+j}\, b_{i+\phi} = H_A L_j (-1)^{k_B - 1} \pm I \qquad (6\text{-}30)$$

for

$\theta \neq j$ $\quad \phi = j+1 \quad$ For a particular

$\theta \neq j+1$ $\quad \phi \neq j \quad$ for all other

$\phi \neq j+1$

We get

$$\sum_{i=o}^{L-1} a_{i+j}\, a_{i+\theta}\, b_{i+j}\, b_{i+\phi} = H_A H_B \pm I \qquad (6\text{-}31)$$

for

$\theta = j$ for a particular γ $\qquad \phi \neq j$

$\theta \neq j$ $\qquad \phi \neq j+1$

$\theta \neq j$ for all other

We get

$$\sum_{i=o}^{L-1} a_{i+j}\, a_{i+\theta}\, b_{i+j}\, b_{i+\phi} = L\, (-1)^{k_A - 1}\, H_B \pm I \qquad (6\text{-}32)$$

If

$\Theta = j+1$ for a particular

$\left.\begin{matrix} \Theta \neq j \\ \Theta \neq j+1 \end{matrix}\right\}$ for all other γ $\qquad \begin{matrix} \phi \neq j \\ \phi \neq j+1 \end{matrix}$

We get

$$\sum_{i=o}^{L-1} a_{i+j}\, a_{i+\theta}\, b_{i+j}\, b_{i+\phi} = H_A H_B \mp (H_A - L\ ) \pm I \qquad (6\text{-}33)$$

If

$V = j+1$ $\qquad$ and $\zeta \neq V+1$

We get results similar to the case

$$V = j{+}1 \qquad \zeta \neq V{+}1$$

except that I should be replaced with F where

$$F \begin{cases} 4 \\ 0 \\ -4 \end{cases} \text{depending on the particular sequence}$$

Now an expression for the third moment can be formulated from the cases tabulated. The third moment is

$$S^3 = \frac{3M-2}{L} M\, H_A H_B$$

$$+ \frac{3}{L}\left\{\left[\binom{M}{3} - (M-3)(M-2) - \sum_{j}^{k_B} BB_\gamma - \sum_{j}^{k_A} BA_\gamma\right.\right.$$

$$\left. - \sum_{J}^{k_B} \overset{*}{B}B_\gamma - \sum_{j}^{k_A} \overset{*}{B}A_\gamma \right] H_A H_B$$

$$+ \sum_{j}^{k_B} B'B_\gamma \; H_A L_\gamma (-1)^{k_B - 1} + \sum_{j}^{k_A} BA_\gamma\, H_B L_\gamma (-1)^{k_A - 1}$$

$$+ \sum_{j}^{k_B} \overset{*}{B}B_\gamma \,(H_A H_B \pm (L_\gamma - H_B)) + \sum_{j}^{k_B} \overset{*}{B}A_\gamma \,(H_A H_B \pm (L_\gamma - H_A))$$

$$+ \left[2(M-3)(M-1) - 2\sum_{j}^{k_B} BB_\gamma - 2\sum_{j}^{k_A} BA_\gamma - 2\sum_{j}^{k_B} \overset{*}{B}B_\gamma - 2\sum_{j}^{k_A} \overset{*}{B}A_\gamma\right] *$$

$$[H_A H_B \pm I]$$

$$+ 2 \sum_{j}^{k_B} BB_\gamma (H_A L_\gamma (-1)^{k_B - 1} \pm I) + 2 \sum_{j}^{k_A} BA_\gamma (H_B L\ (-1)^{k_A - 1} \pm I)$$

$$+2 \sum_{j}^{k_B} \overset{*}{B}B\gamma\ (H_A H_B) \mp (H_B - L_\gamma \pm I) + 2 \sum_{j}^{k_A} \overset{*}{B}A\gamma\ ((H_A H_B) \mp (H_A - L\gamma) \pm I)$$

$$+ [(M-3) - \sum_{j}^{k_B} BB_\gamma - \sum_{j}^{k_A} BA_\gamma - \sum_{j}^{k_B} \overset{*}{B}B_\gamma$$

$$- \sum_{j}^{k_A} \overset{*}{B}A_\gamma][H_A H_B \pm F]$$

$$+ \sum_{j}^{k_B} BB_\gamma (H_A L_\gamma (-1)^{k_B - 1} \pm F) + \sum_{j}^{k_A} BA_\gamma (H_B L_\gamma (-1)^{k_A - 1} \pm F)$$

$$+ \sum_{j}^{k_B} \overset{*}{B}B_\gamma ((H_A H_B) \mp (H_B - L_\gamma) \pm F)$$

$$+ \sum_{j}^{k_A} BA_\gamma ((H_A H_B) \mp (H_A - L_\gamma) \pm F) \Bigg\} \qquad (6\text{-}34)$$

where

BA_γ : Number of characteristic trinomials of j, code A.

BB_γ : Number of characteristic trinomials of j, code B.

[Characteristic trinomials power less than or equal to M-1]

$\overset{*}{B}A_\gamma$: Number of trinomials containing characteristic of order M-2, code A.

$\overset{*}{B}B_\gamma$: Number of trinomials containing characteristic of order M-2, code B.

The expression for third moment in 6-34 is involved; however it can be observed that deviation of the third movement from that expected with a random sequence depends on BB_γ, BAj, and B^*A_γ, and B^*B_γ — the terms reflecting the number of trinomials that contain the γth sequence characteristic equation. Therefore minimizing S^3 involves minimizing these factors.

7. Code Interference Program

Code interference presents a special problem in spread spectrum transponder design. Code interference is of two types:

(1) Code cross interference

(2) Code self interference

Code interference is important because transponder performance is evaluated based upon an assumption of random noise and interference. Auto and cross correlation of pseudo random codes used in spread spectrum transponders can be significantly larger than those expected with true random codes. For this reason a computer program designed to evaluate code interference was developed.

The program basically evaluates the statistical properties of the code interference signal at point A of the simplified code tracking loop of figure 7-1.

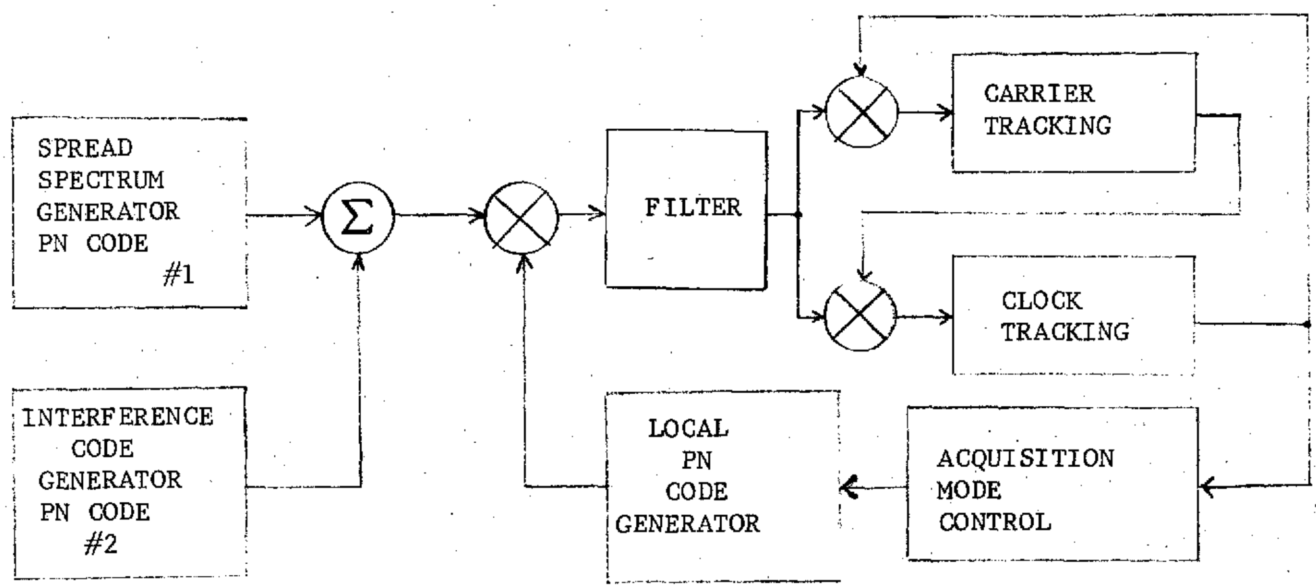


Figure 7-1 Code Interference at Input of Code Tracking Loop

The filter has a M-chip impulse response, code generator # 1 is of length L1 and code generator # 2 is of length L2. Figure 7-2 shows schematically the relation of L1, L2, and M.

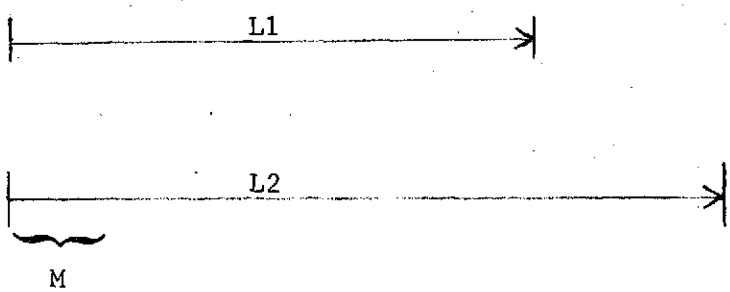


Figure 7-2 Relationship of L1, L2, and M

There are L1 L2 possible M-tuples generated at point A. The following flow diagram describer an algorithm designed to calculate the statistical properties of the error signal at A. Code generators in the spread spectrum modulaton 1 and 2 may be ML, HS, or Gold codes.

READ MIN, MFI MST — MIN: Minimum Window; MFI: Maximum Window; MST: Window Step

DO M=MIN, MFI,MST — Do over all windows

GPH1=0 — Initialize Gold Phase 1 to "0"

Read Set 1 Polynomials — CALL RPCC (NPOL1, LST1, L1, IA1, NOR1, GPH1) — NPOL1 - # of component codes; LST 1 - Code classification; L1 - Code length array; NOR1 - Code order array; GPH1 - Gold Phase

Calculate Set 1 Code Length — CALL CALSL (NPOL 1, LST 1, L1, LT1) — LT1 - Total code length, code set No. 1

GPH2=0 — Initialize gold phase 2 to "0"

Read Set 1 Polynomials — CALL RPCC (NPOL2, LST2, L2, IA2, NOR2, GPH2) — See definitions above

Calculate Set 2 Code length — CALL CALSL (NPOL2, LST2, L2, LT2) — LT2 - Total code length, code set No. 2

LT1.LT.LT2 - Otherwise error

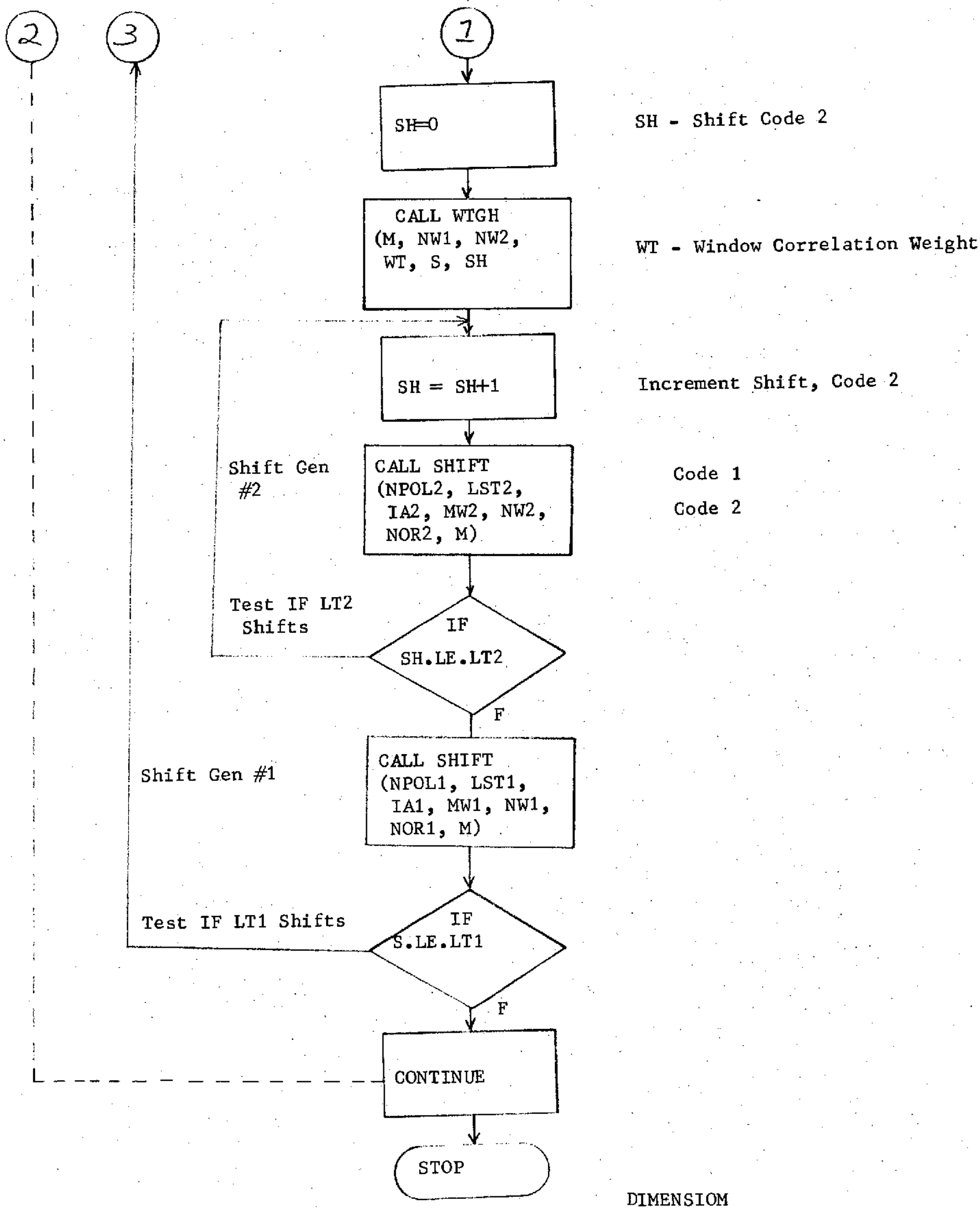
S=0 — S - Shift, code 1

Initialize Code Set 1 — CALL INTGE (NPOL1, LST1, L1, LT1, IA1, MW1, M, NOR1, GPH1, NW1) — MW1 - M - Window Array (INTIAL) from Gen. #1

S=S+1 — Increment shift, code 1

Initialize Code Set 2 — CALL INTGE (NPOL2, LST2, L2, LT2, IA2, MW2, M, NOR2, GPH2, NW2) — MW 2 - M - Window array from Gen #2

2 3 1

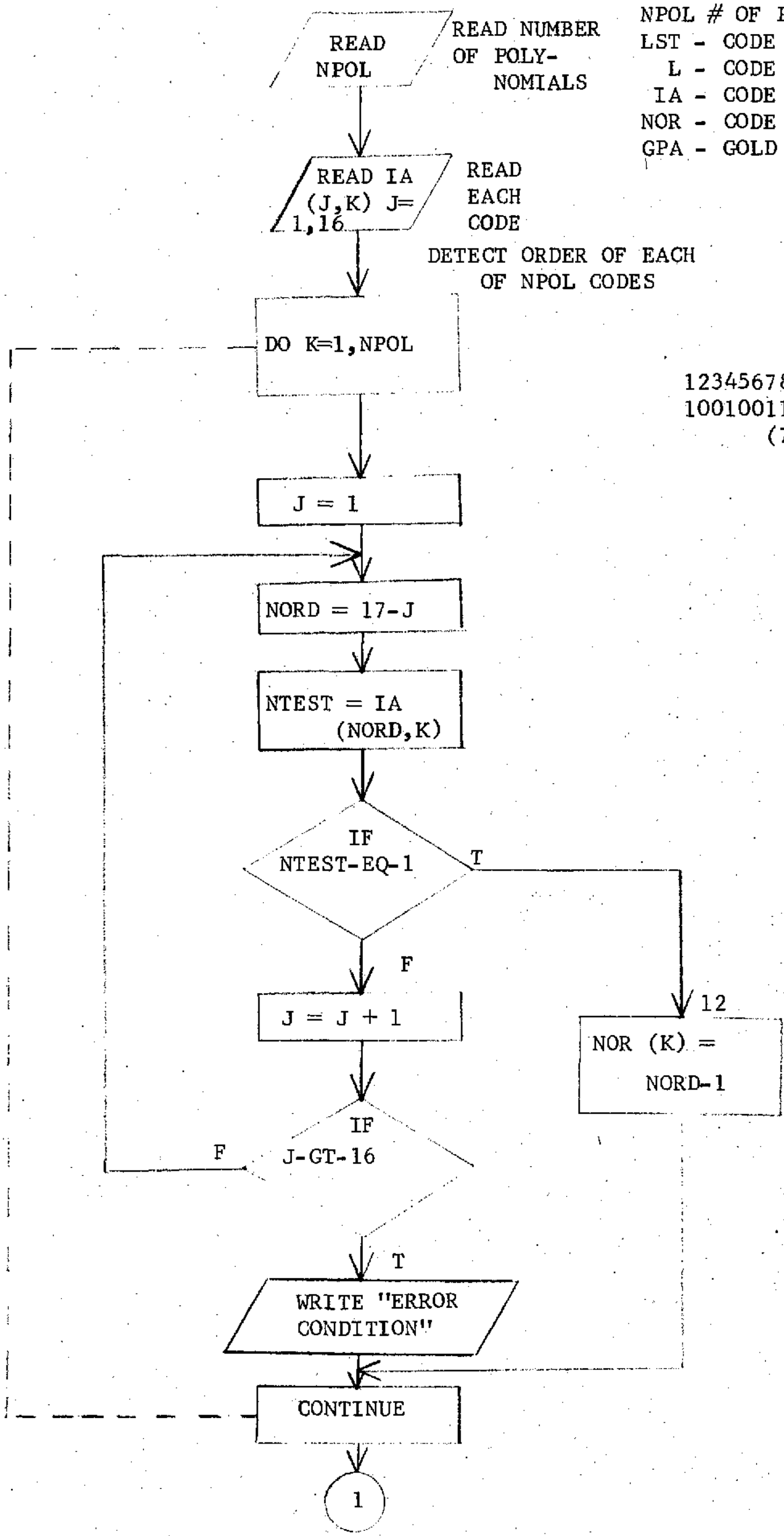


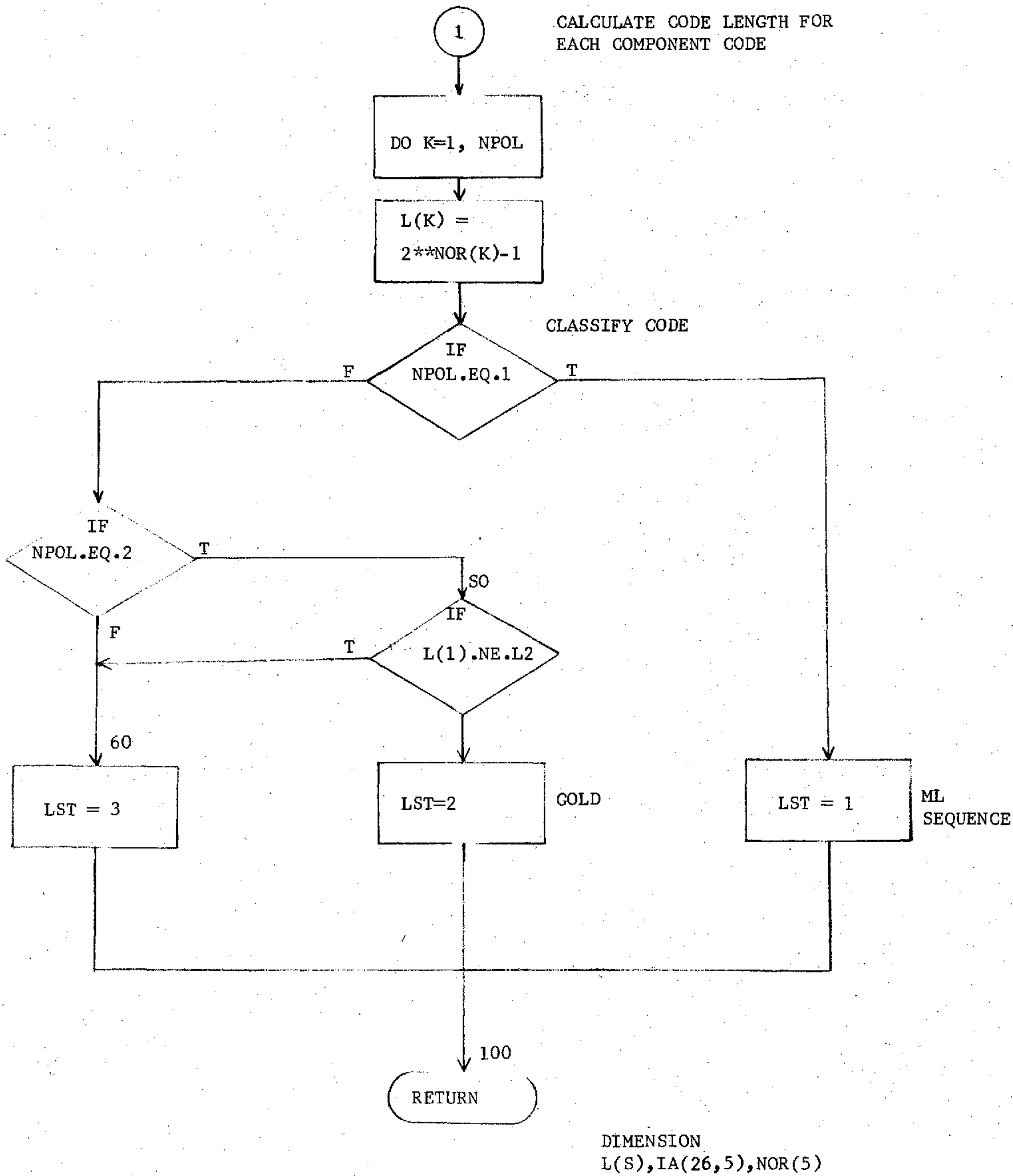
DIMENSIOM
L1(5), IA1(26,5), NOR1(5)
L2(5), IA2(26,5), NOR2(5)
MW1(500,5), MW2(500,5)
NW1(500), NW2(500)

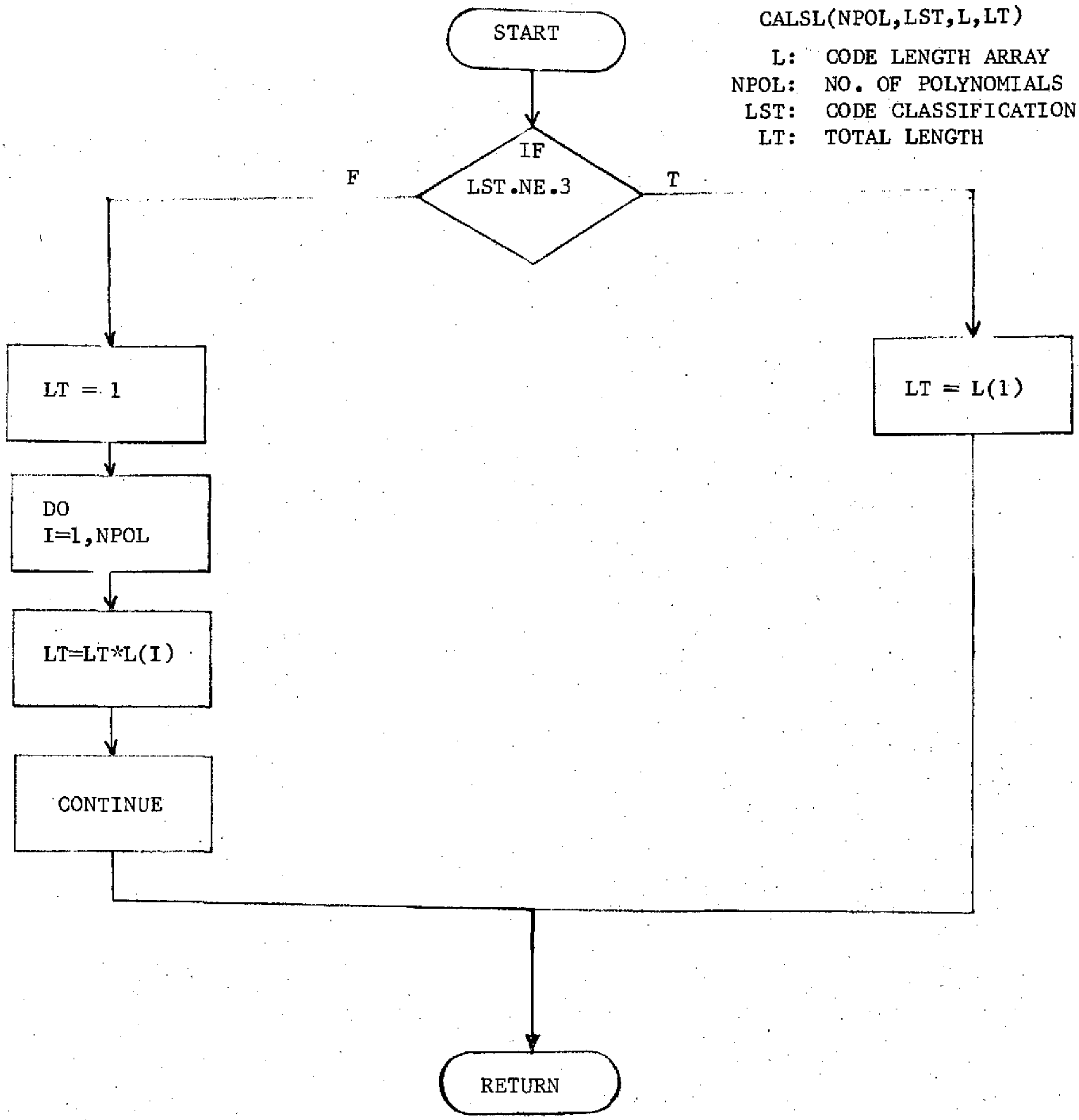
SUBROUTINE RPCC (NPOL,LST,L,IA,
NOR,GPH)
NPOL # OF POSSIBLE CODES
LST - CODE CLASSIFICATION
L - CODE LENGTH ARRAY
IA - CODE ARRAY
NOR - CODE ORDER ARRAY
GPA - GOLD PHASE

EXAMPLE:

1234567891011213141516
1001001100 0 0 0 0 0 0
(7,6,3,0)




1
CALCULATE CODE LENGTH FOR EACH COMPONENT CODE
DO K=1, NPOL
L(K) =
2**NOR(K)-1
CLASSIFY CODE
IF
NPOL.EQ.1
F
T
IF
NPOL.EQ.2
T
SO
IF
L(1).NE.L2
F
T
60
LST = 3
LST=2
GOLD
LST = 1
ML
SEQUENCE
100
RETURN
DIMENSION
L(S),IA(26,5),NOR(5)

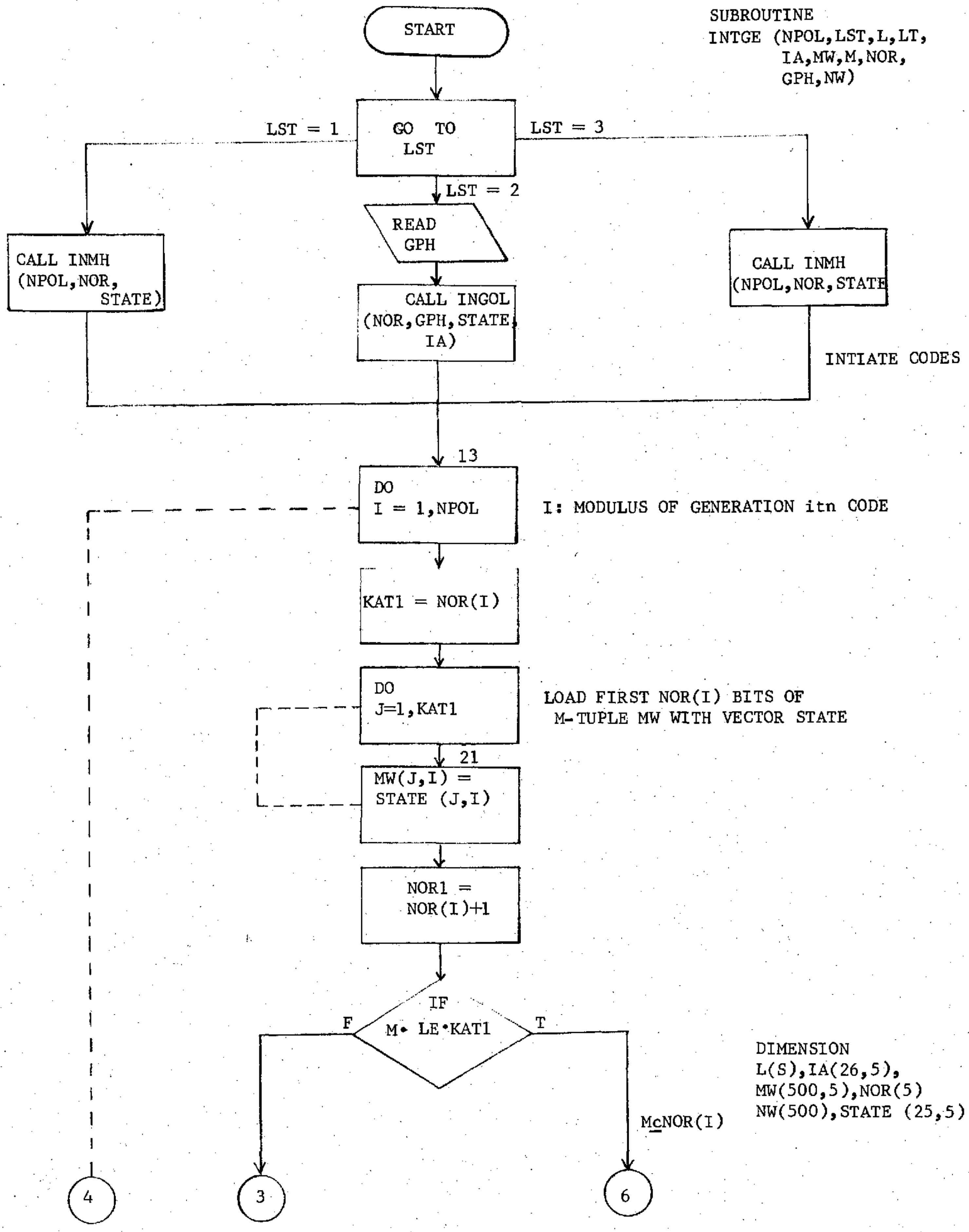


CALSL(NPOL,LST,L,LT)

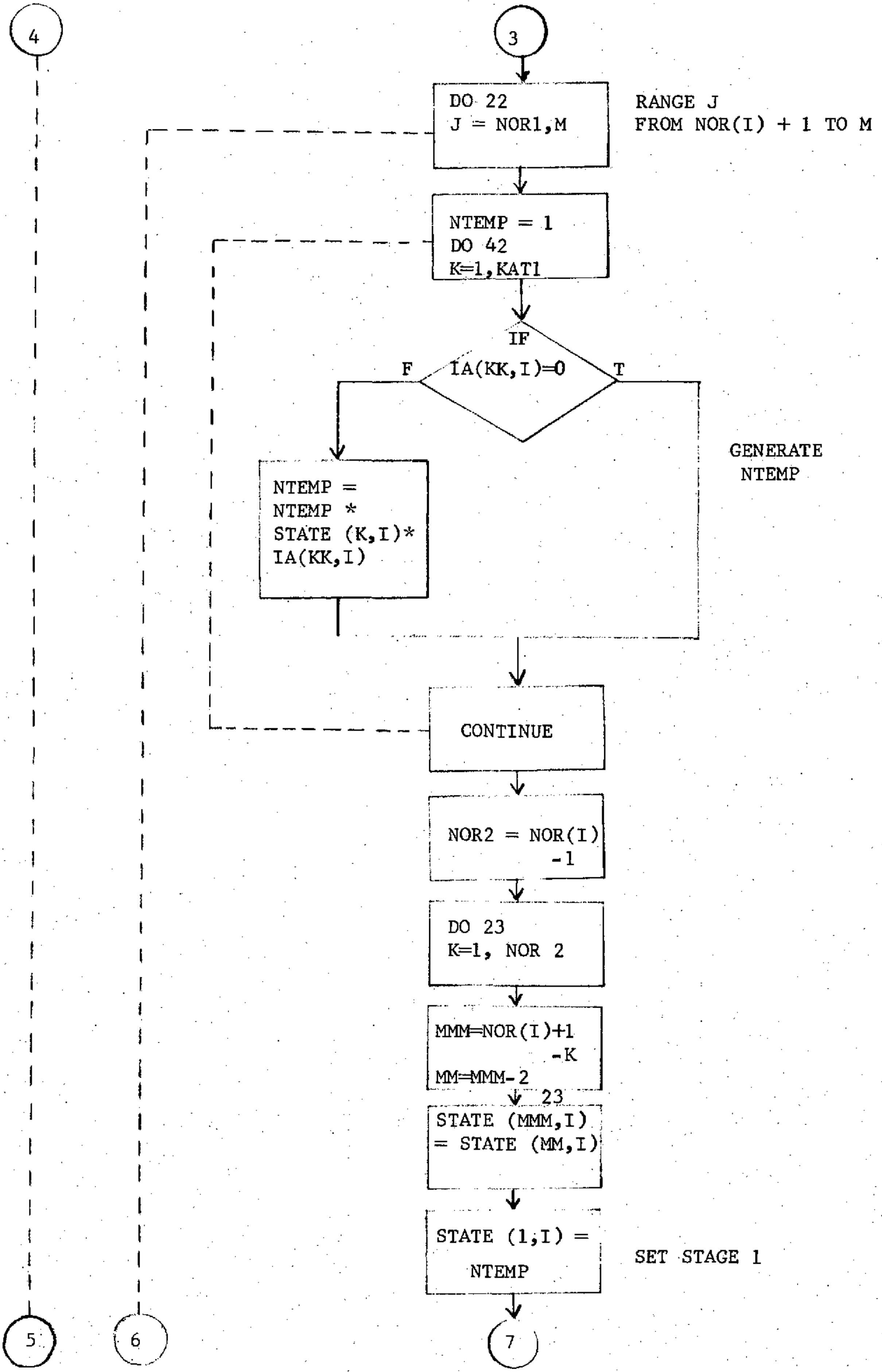
L: CODE LENGTH ARRAY
NPOL: NO. OF POLYNOMIALS
LST: CODE CLASSIFICATION
LT: TOTAL LENGTH

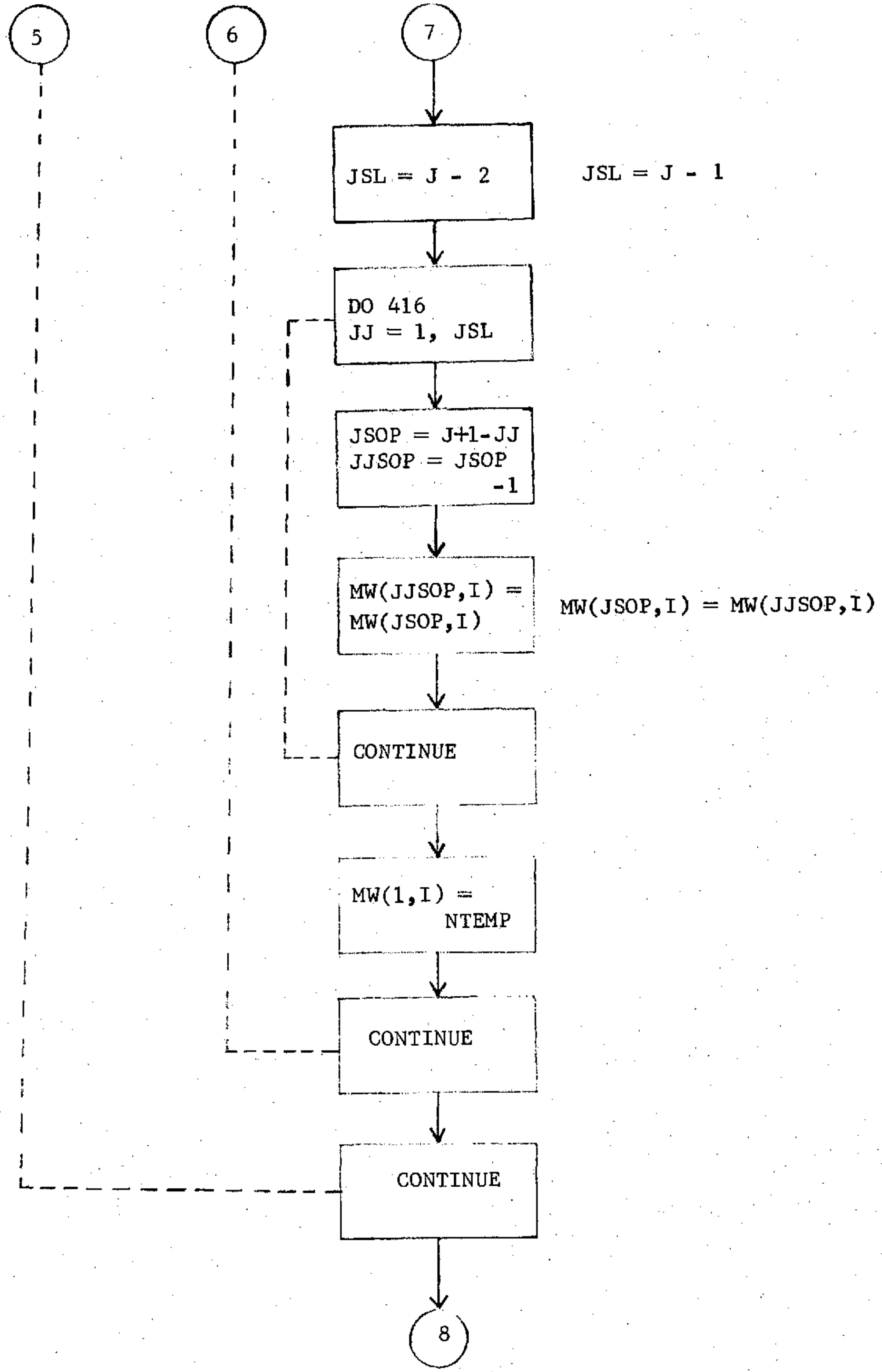
DIMENSION L(10)

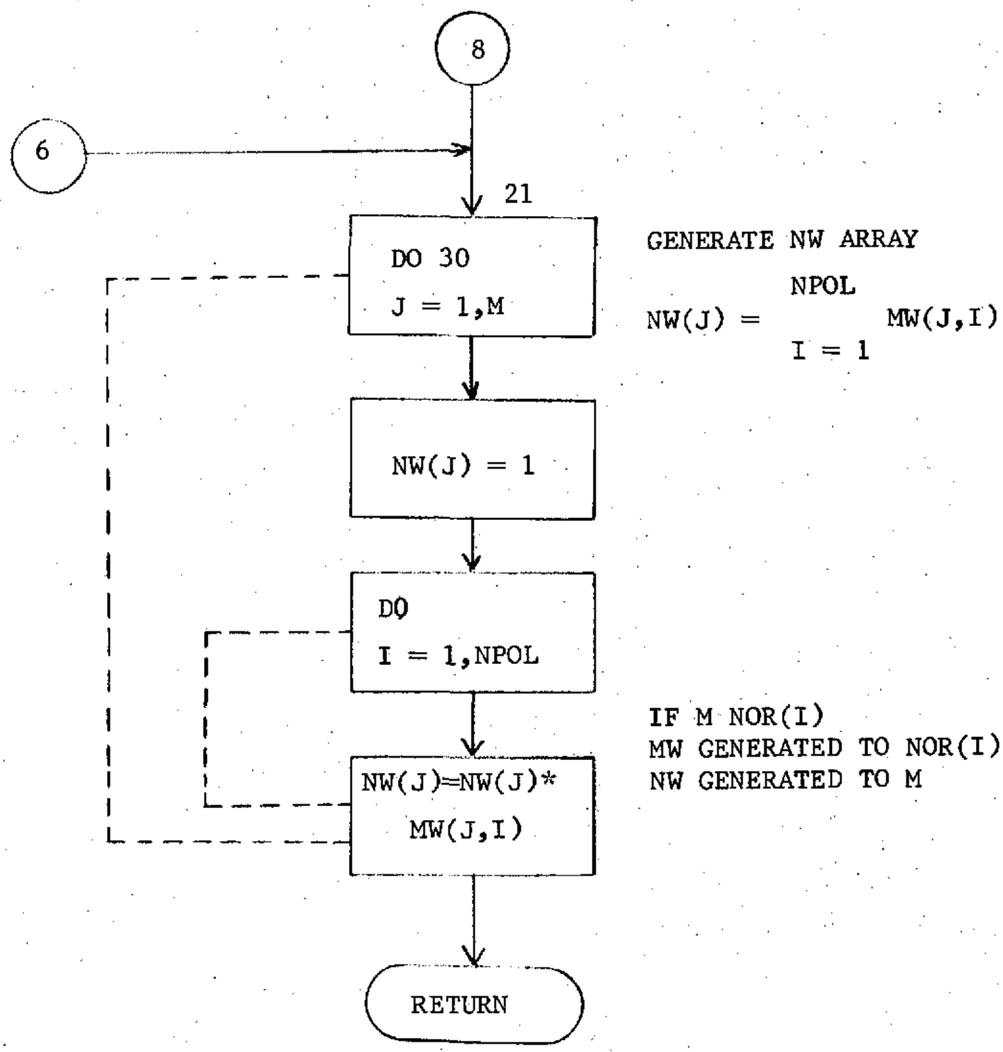
SUBROUTINE
INTGE (NPOL,LST,L,LT,
IA,MW,M,NOR,
GPH,NW)



DIMENSION
L(S),IA(26,5),
MW(500,5),NOR(5)
NW(500),STATE (25,5)

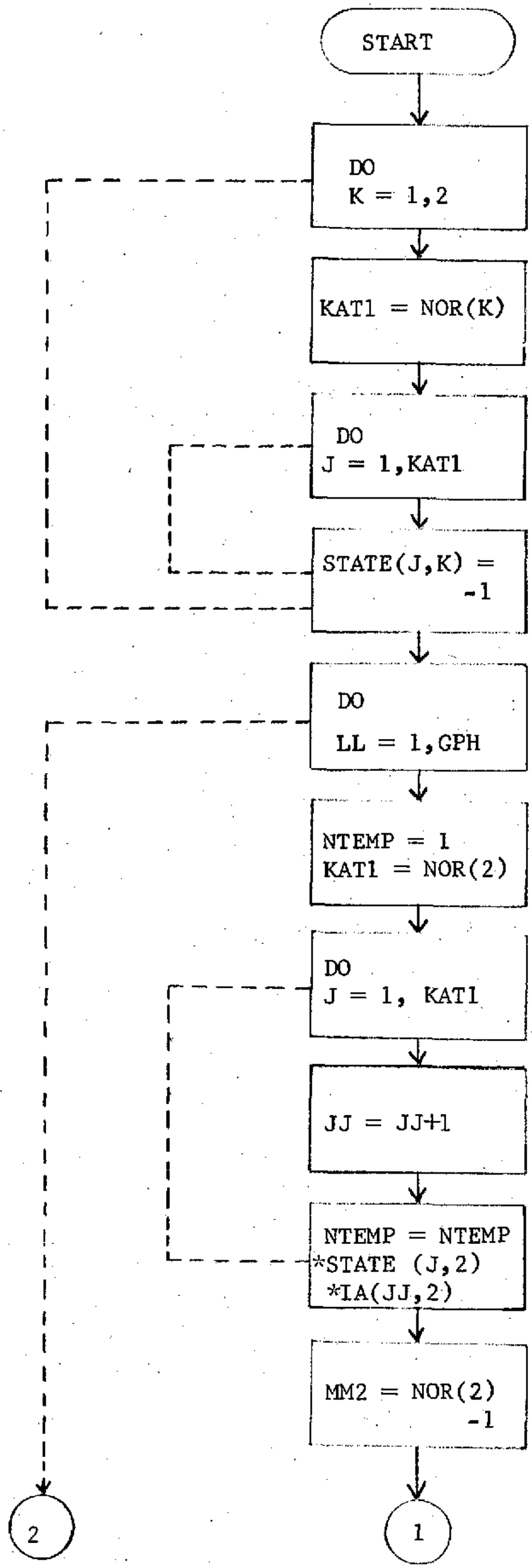

4
3
DO 22
J = NOR1,M
RANGE J
FROM NOR(I) + 1 TO M
NTEMP = 1
DO 42
K=1,KAT1
IF
IA(KK,I)=0
F
T
NTEMP =
NTEMP *
STATE (K,I)*
IA(KK,I)
GENERATE
NTEMP
CONTINUE
NOR2 = NOR(I)
-1
DO 23
K=1, NOR 2
MMM=NOR(I)+1
-K
MM=MMM-2
23
STATE (MMM,I)
= STATE (MM,I)
STATE (1,I) =
NTEMP
SET STAGE 1
5
6
7


5
6
7
JSL = J - 2
JSL = J - 1
DO 416
JJ = 1, JSL
JSOP = J+1-JJ
JJSOP = JSOP
-1
MW(JJSOP,I) =
MW(JSOP,I)
MW(JSOP,I) = MW(JJSOP,I)
CONTINUE
MW(1,I) =
NTEMP
CONTINUE
CONTINUE
8


8
6
21
DO 30
J = 1,M
GENERATE NW ARRAY
NPOL
NW(J) =
MW(J,I)
I = 1
NW(J) = 1
DO
I = 1,NPOL
IF M NOR(I)
MW GENERATED TO NOR(I)
NW GENERATED TO M
NW(J)=NW(J)*
MW(J,I)
RETURN

INGOL

SUBROUTINE INGOL (NOR,GPH,STATE,
IA)



SET BOTH REGISTERS
TO THE ALL "1" STATE

SHIFT PHASE 6F SECOND
REGISTER GPH

CALCULATE NTEMP

Example

J = 1
JJ = 2

```
                (7,6,3,0)
     1 2 3 4 5 6 7 8 9 ...
IA   1 0 0 1 0 0 1 1 0 ...
        / / / / / / /
STATE    1 1 1 1 1 1
```

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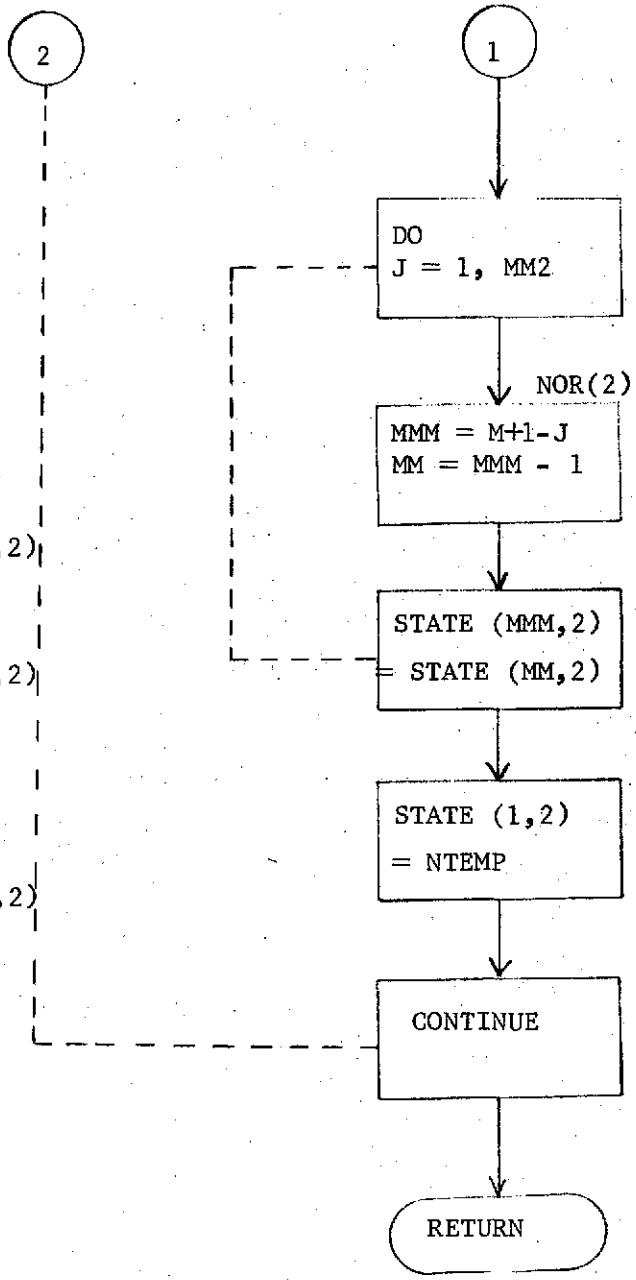
EXAMPLE
NOR(2)=7
MM2 =6

J=1
MMM=7
MM=6

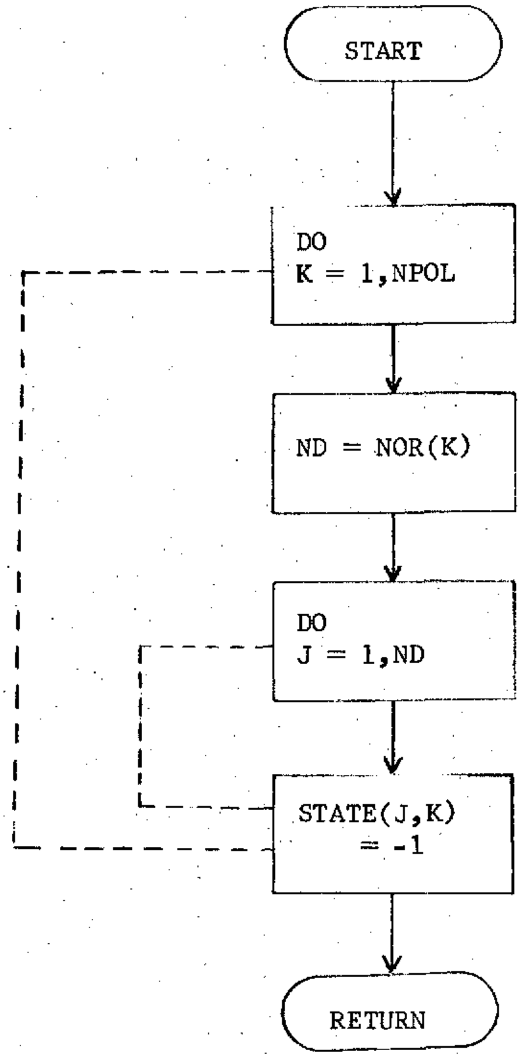
STATE(7,2)=STATE(6,2)
J=2
MMM=6
MM=5
STATE(6,2)=STATE(5,2)
.
.
.
J=6
MMM=2
MM=1
STATE(2,2)=STATE(1,2)



DIMENSION
NOR (5)
STATE (25,5)
IA (26,5)

INMH

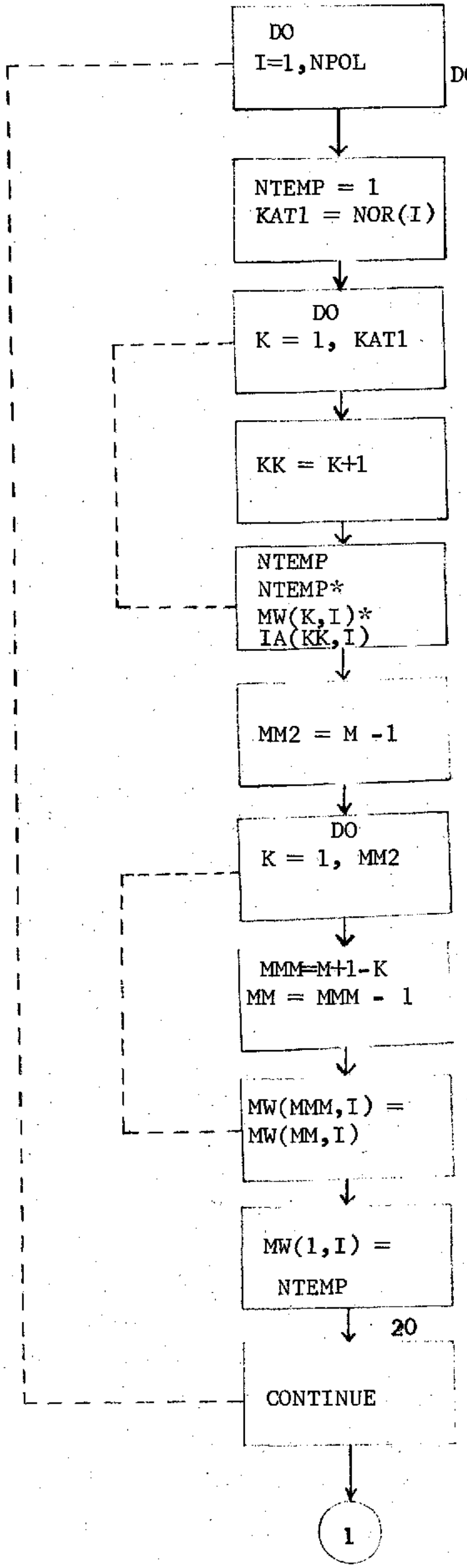
SUBROUTINE INMH
(NPOL,NOR,STATE)



DIMENSION
NOR(5)
STATE (25,5)

SHIFT

SUBROUTINE SHIFT
(NPOL,LST,IA,MW,NW,NOR,M)



DO OVER NUMBER
OF POLYNOMIALS

NTEMP:HEW INPUT BIT

GENERATE
NEW
INPUT
BIT

EXAMPLE
(7,6,3,0)
12345678910111213...
IA 10010011000000000...
///////
MW -1-1-1 1-1-1-1

NTEMP = -1

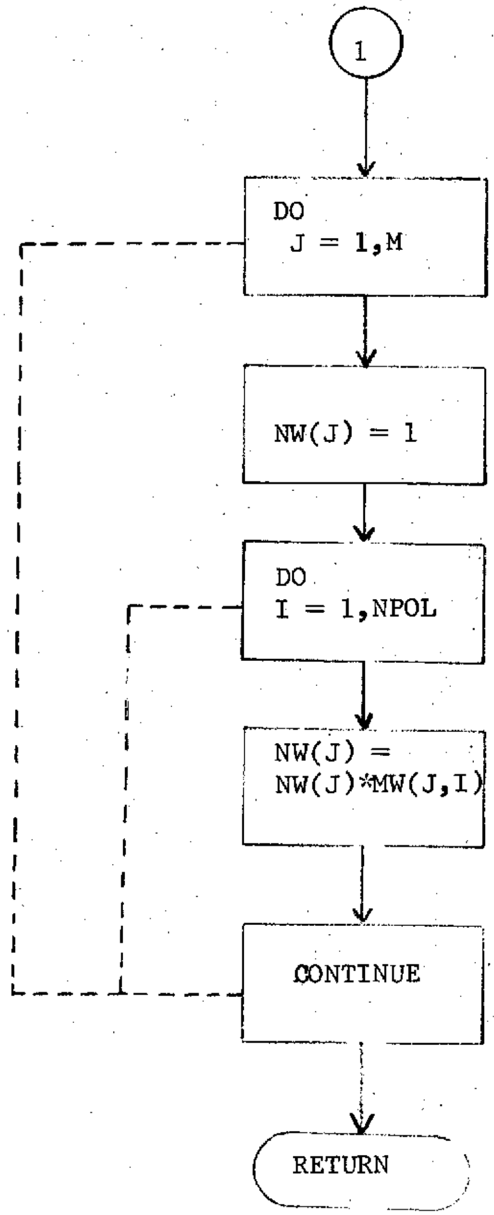
IA: CODE ARRAY
MW: M-TUPLE CONTENT ARRAY
FOR INDIVIDUAL CODE

SHIFT ENTIRE MW ARRAY

EXAMPLE
M = 500
MM2 = 499
K = 1
MMM = 500
MM = 499
MW(500,I)=MW(499,I)
K = 2
MMM = 499
MM = 498
MW(499,I)=MW(498,I)
.
.
.
.
K = 499
MMM = 2
MM = 1
MW(2,I) = MW(1,I)

LOAD NTEMP INTO 1st STAGE

SHIFT CONTD

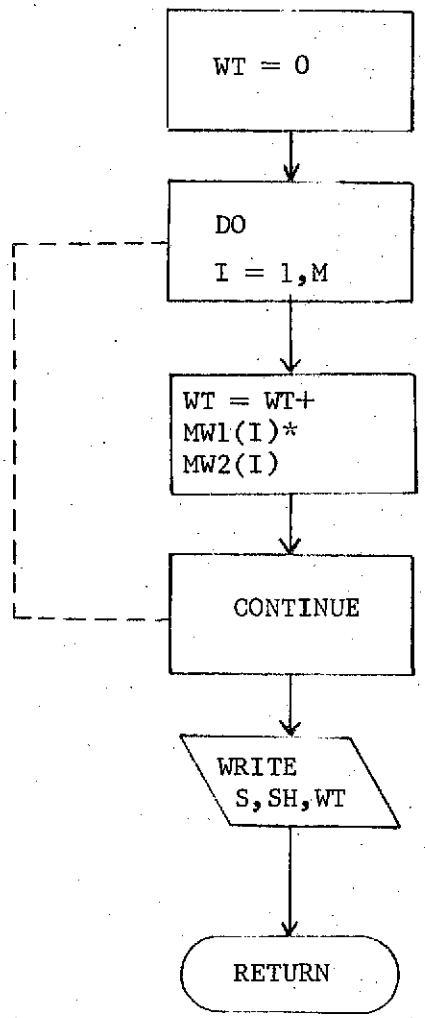


GENERATE NW ARRAY FOR GENERATOR AS MOD - 2 SUM OF MW ARRAYS OF INDIVIDUAL COMPONENT CODES

$$NW(J) = \sum_{k=1}^{NPOL} MW(J,I)$$

WTGH

SUBROUTINE WTGH
(M,MW1,MW2,WT,S,SH)
M-tuple length
MW1 - GEN. # 1 WEIGHT ARRAY
(SAME AS NW1 IN MAIN)
MW2 - GEN. # 2 WEIGHT ARRAY
(SAME AS NW2 IN MAIN)
WT - WEIGHT
S - PHASE SHIFT 1
SH - PHASE SHIFT 2



DIMENSION
MW1(500),MW2(500)

8. Generation of High-Speed Maximum Length Digital Sequences

This section is the result of a study of sampled maximum length digital sequences. The purpose of the study was to establish the mathematical basis for the design of a high speed digital PSEUDORANDOM SEQUENCE GENERATOR FOR USE IN THE HEAO-C SPREAD SPECTRUM TRANSPONDER. The proposed procedure for generating the high speed ML sequence involves sampling several slower speed ML generations. Figure 8-1 illustrates the sequence generator.

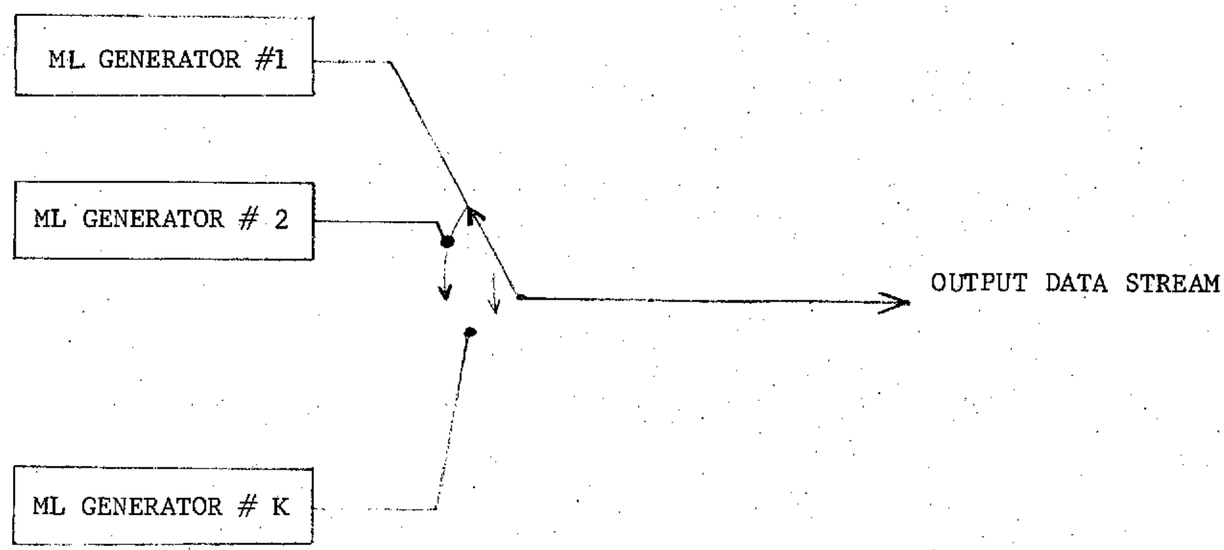


Figure 8-1 HIGH SPEED SEQUENCE GENERATOR

If there are K ML generators forming the sequence generator where $K=2^{\xi}$, ξ an integer, then the commutation rate should be K times the clock rate of ML generators. Each generator is sampled once during a clock interval, and the output data stream would consist of K digits during the clock-interval. The advantage of this configuration is that a higher speed digital bit stream can be generated with ML sequence generators operating with a clock-frequency that is only a fraction of the data-rate.

Specifically, if the data-rate is F_B bits/sec, the required clock-rate is F_B/K . For example, if it is desired to operate with a data rate of 40×10^6 Bits/sec, and assuming K=4 (four ML generators) then the generator clock rates would be 10 MHZ. This allows the use of less-expensive, more-reliable digital components from lower speed logic families. The only component required to operate at the 40 MHZ rate is the commutating switch.

Figure 8-2 shows the error checking portion of the overall system.

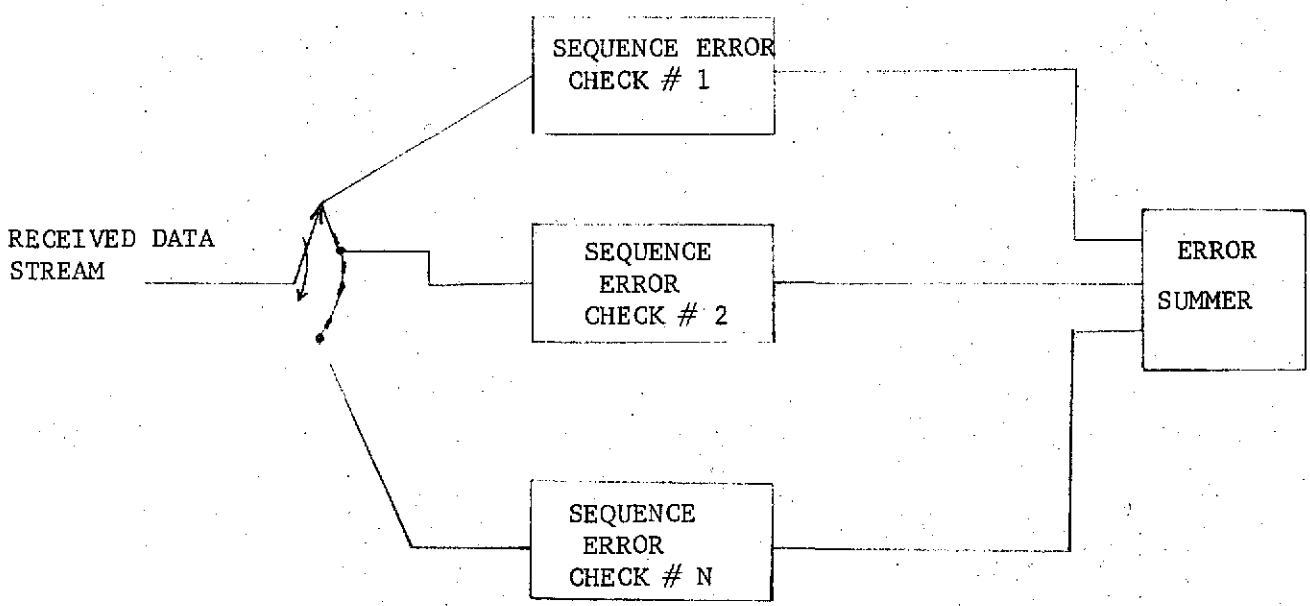


Figure 8-2. Digital Error Checking at Receiver

As in the case of the sequence generator, the error checking sub-system operates on a sampling basis. If $N=2^q$, q an integer, and if the received data stream is a maximum-length digital sequence, then sampling the data stream with a commutator operating at the rate of the incoming data stream yields a shifted version of the same ML sequence to each error check block. In this way the error check digital circuit is required to operate with a clock rate that is only F_B/N HZ. With this procedure, the same advantages mentioned for the sequence generator apply for the error check section. In the system design N is not necessarily equal to K.

Figure 8-3 shows the overall arrangement for digital communications system evaluation.

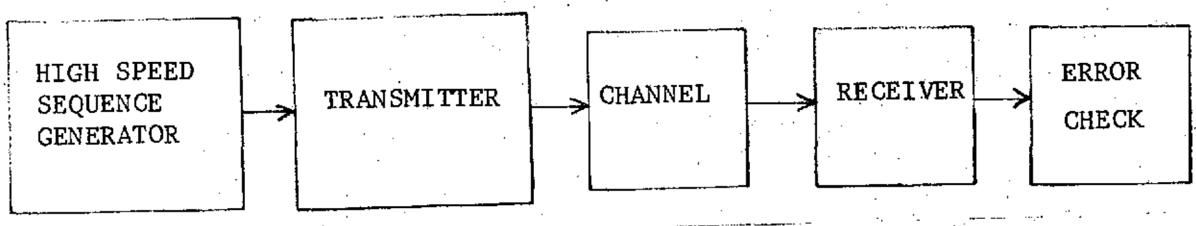


Figure 8-3. Overall Test-Configuration

Two important considerations involving the system shown in figure 8-3 are the phasing of the maximum length sequence generators shown in figure 1 to provide the desired output data stream, and the design of the sequence error check blocks shown in figure 8-2. This section deals with the phasing problems.

Sequence Generator Phasing

The phasing problem can be stated: "what initial phasing of the K ML-generators shown in figure 8-1 are required to provide the ML-sequence in the output data stream when the sampling procedure is used." The rule with K ML-generators for phasing the ith generator relative to the first generator is

1. Advance by (i-1) (L+1)/K bits

or

2. Delay be (i-1) (L-1)/K bits

The rationale for the above choice of phase relation is as follows:

1. Sampling a ML sequence provides a shifted version of the same sequence if the sampling rate is an integer power of two.

2. Consider the synthesized sequence as being reconstructed from K, K-sampled versions of itself.

3. Consecutive digits in a ML sequence must be separated by sampled sequence.

4. Arranging K_1 K-Sequences each advanced by (L+1)/K bits relative to its adjacent sequence, and sampling from each as shown in figure 1, must yield the same ML sequence.

As an example consider the ML sequence abcdefg. Sampling every other bit yields acegbdf which must be a shifted version of the same sequence. Advancing this sequencing by $(L+1)/K = 8/2 = 4$ bits yields bdfaceg. Synthesizing by sampling in-turn from the two sequence yields abcdefabcdef, which is the original ML sequence repeated twice.

As a practical illustration consider the ML sequence generator with the characteristic polynomial

$$G(z) = 1 + z + z^3 \qquad (8\text{-}1)$$

The sequence generated by this ML generator is S(z), where

$$\frac{1}{G(z)} = \frac{S(z)}{1+z^L} \qquad (8\text{-}2)$$

and where

$$L = 2^N - 1 \qquad (8\text{-}3)$$

for an N-stage shift-register generator. For the generator in question

$$S(z) = 1 + z + z^2 + z^4 \qquad (8\text{-}4)$$

which represents the sequence 1110100. Forming $S^*(z)$ by advancing the phase by

$$(L+1)/K = 4 \text{ for } K = 2 \qquad (8\text{-}5)$$

and sampling in turn from S(z) and $S^*(z)$ yields

$$\left.\begin{array}{ll} \text{SEQUENCE:} & 1110100 \\ \text{SEQUENCE} & \\ \text{Advanced} & \\ \text{BK 4 BITS:} & 1001110 \end{array}\right\} \quad 11101001110100$$

as expected. Performing the similar analysis for $K = 4$ yields

$$\left.\begin{array}{l} 1110100 \\ 1010011 \\ 1001110 \\ 0111010 \end{array}\right\} \quad 1110100111010011101001110100.$$

The output for the sampling generator is

$$S^2(z) + z\,(S(z)z^{(L-1)/2})^2 = SS(z) \qquad \text{MOD } 2L \qquad (8\text{-}6)$$

For K=2
or

$$SS(z) = S^2(z)(1+z^L) \qquad \text{MOD } 2L \qquad (8\text{-}7)$$

The sequence can also be expressed as

$$SS(z) = (1+z^L)^3/(G^2(z)) \qquad \text{MOD } 2L \qquad (8\text{-}8)$$

In general, for K component generators

$$SS(z) = \sum_{i=1}^{K} z^{i-1}\, S^K(z) z^{(i-1)(L-1)} \qquad (8\text{-}9)$$

or

$$SS(z) = \sum_{i=1}^{K} \frac{z^{(i-1)}{}_L(1-z^L)^K}{G^K(z)} \qquad \text{MOD } 2L \qquad (8\text{-}10)$$

The synthesized sequence can be expressed as a shifted version of the original ML sequences,

$$SS(z) = z^x(1+z^L)\; S\;(z) \qquad \text{MOD } 2L \qquad (8\text{-}11)$$

for the case K=2, or in general,

$$SS(z) = z^x S(z) \sum_{i=1}^{K} z^{(i-1)L} \qquad \text{MOD } 2L \qquad (8\text{-}12)$$

or

$$SS(z) = \frac{z^x(1+z^L)}{G(z)} \sum_{i=1}^{K} z^{(i-1)L} \qquad \text{MOD KL,} \qquad (8\text{-}13)$$

Equating (8-9) and (8-12) yields

$$S^K(z)\sum_{i=1}^{K} z^{(i-1)L} = z^x S(z)\sum_{i=1}^{K} z^{(i-1)L} \qquad \text{MOD KL,} \qquad (8\text{-}14)$$

or

$$S^K(z)\sum_{i=1}^{K} z^{(i-1)L} + z^x\ S(z)\sum_{i=1}^{K} z^{(i-1)L} = 0 \qquad \text{MOD KL.} \qquad (8\text{-}15)$$

For the case K=2 this reduces to

$$S^2(z)(1+z^L) + z^x S(z)(1+z^L) = 0 \qquad \text{MOD 2L} \qquad (8\text{-}16)$$

Simplification of (8-15) yields

$$S^{K-1}(z) + z^x = 0 \qquad \text{MOD KL,} \qquad (8\text{-}17)$$

which must be satisfied by the sequence.

If equations (8-17) describes the sequence generated by S(z) then

$$G(z) \mid (S^{K-1}(z) + z^K) \qquad \text{MOD KL.}$$

Now

$$\frac{S^{K-1}(z) + z^x}{G(z)} = \frac{(1+z^L)^{K-1} + z^x G^{K-1}(z)}{G^K(z)} \qquad \text{MOD KL,} \qquad (8\text{-}18)$$

or

$$G^K(z) \mid ((1+z^L)^{K-1} + z^x G^{K-1}(z)) \qquad \text{MOD KL.} \qquad (8\text{-}19)$$

For the example:

$$G(z) = 1 + z+z^3 \qquad (8\text{-}20)$$

and

$$K = 2 \qquad (8\text{-}21)$$

yields

$$X = 0 \qquad (8\text{-}22)$$

and

$$\frac{(1+z^L)^{K-1} + z^x G^{K-1}(z)}{G^K(z)} = \frac{z + z^3 + z^7}{1+z^2+z^6} = z \qquad \text{MOD } 14. \qquad (8\text{-}23)$$

A similar example for

$$G(z) = 1 + z^2 + z^3$$

where

$$S(z) = 1 + z^2 + z^3 + z^4$$

represents the sequence 1011100.

A sampling arrangement for K=2, requires a delay = (L-1)/2=3, shown below

$$\left.\begin{matrix}1011100\\1001011\end{matrix}\right\} 11001011100101$$

observe for this case X=4,

and

$$\frac{(1+z^L)^{K-1} + z^x G^{K-1}(z)}{G^K(z)} = \frac{1 + z^4 + z^6}{1 + z^4 + z^6} = 1 \qquad \text{MOD } 14. \qquad (8\text{-}24)$$

An algorithm to calculate X is as follows:

1. Starting with the all zero (N-1) - tube generate the sequence, S(z), with the characteristics equation G(z).

2. Generate $S^K(z)$ From $S(z)$ Or $G^K(z)$.

3. Find X such that $(1+z^L)^{K-1} + z^x G^{K-1}(z)$ forms a recursive relation that holds over the all zero 2(N-1) - tuple of $S^K(z)$.

Sample calculations are shown below For K=2.

Sample 1:

$$G(z) = 1 + z^2 + z^3 \qquad (8\text{-}25)$$

0 0 1 0 1 1 1 0 0 1 0 1 1 1

00 00 10 00 10 10 10 00 00 10 00 10 10 10

$$1 + z^7 + z^4\,(1+z^2+z^3) = 1 + z^4 + z^6 = G^2(z) \qquad (8\text{-}26)$$

Sample 2

$$G(z) = 1 + z + z^3 \tag{8-27}$$

0 0 1 1 1 0 1 0 0 1 1 1 0 1

00 00 10 1010001000001010100010

$$1 + z^7 + z^0\ (1+z+z^3) = z(1+z^2+z^6) = zG^2(z) \tag{8-28}$$

A similar problem involves the phase of the sequence resulting from sampling a sequence at a rate

$$r = 2^g \tag{8-29}$$

with g an interger. For example,

$$G(z) = 1 + z^2+z^3 \tag{8-30}$$

yields the sequence

1011100.

Sampling this sequence with r=2 yields

1110010

which is a phase shift corresponding to z^5.

As another example

$$G(z) = 1 + z + z^3 \tag{8-31}$$

yields the sequence

1110100.

Sampling the sequence with r=2 yields

1110100

which is a phase shift corresponding to Z^o.

A procedure for determining the phase shift can be found if an expression of the form S(z) can be found for the sequence formed as a result of sampling, and

$$f(z) = z^x S(z) \qquad \text{MOD L.} \tag{8-32}$$

If an ML sequence is sampled at a rate of

$$r=2 \tag{8-33}$$

then an adjacent bit will be sampled $(\frac{L+1}{2})$ bits after the sampled bit in the sequence formed from sampling. Extending this analysis, a sampled sequence can be expressed as

$$f(z) = \sum_{i=o}^{\frac{L-1}{2}} z^i (z^{2i} S(z)^{(L+1)/2}) \quad \text{MOD } \frac{L+1}{2}L \tag{8-34}$$

or alternately

$$f(z) = \sum_{i=o}^{\frac{L-1}{2}} z^i (z^{L-2i} \; S(z)^{\frac{L+1}{2}}) \quad \text{MOD} (\frac{L+1}{2})L \tag{8-35}$$

For example, the sequence with characteristic equation

$$G(z) = 1+z^2+z^3 \tag{8-36}$$

and

$$S(z) = 1 + z^2 + z^3 + z^4, \; 1011100 \tag{8-37}$$

with

$$S^4(z) = 1 + z^8 + z^{12} + z^{16} \quad \text{MOD } 28 \tag{8-38}$$

$$z^1(z^5)^4 S^4(z) = z + z^5+z^9+z^{21} \quad \text{MOD } 28 \tag{8-39}$$

$$z^2(z^3)^4 S^4(z) = z^2+z^{14}+z^{22}+z^{28} \quad \text{MOD } 28 \tag{8-40}$$

and

$$z^3(z)^4 S^4(z) = z^7+z^{15}+z^{19}+z^{23} \quad \text{MOD } 28 \tag{8-41}$$

yields from equation (30)

$$f(z) = 1+z+z^2+z^5+z^7+z^8+z^9+z^{12}+z^{14}+z^{15}+z^{16}+z^{19}+z^{21} + z^{22}+z^{23}+z^{26} \quad \text{MOD } 28 \tag{8-42}$$

or

1110010111001011100101110010.

Reference Sequences:			
1011100 = $z^7S(z)$	1110010 = $z^5S(z)$	1001011 = $z^3S(z)$	0101110=$zS(z)$
Sum Sequence:			
1000000010001000100000000000	= $[z^7S(z)]^4$	MOD 28	
0100010001000000000000100 0000	= $[z^5S(z)]^4$	MOD 28	
0010000000000001000000000100010	= $[z^3S(z)]^4$	MOD 28	
0000000100000000100010001000 0	= $[zS(z)]^4$	MOD 28	
1 1100101110010111001011100 10	= $f(z)$	MOD 28	

Table 8-1. Formation of synthesized sequence For the Case

$G(z) = 1 + z^2 + z^3$

The table above also illustrates the result of equation (8-30).

Now from equation (8-37).

$$\sum_{i=o}^{\frac{L-1}{2}} z^i \left[z^{L-2i} \; S(z)\right]^{\frac{L+1}{2}} = z^x S(z) \sum_{i=o}^{\frac{L-1}{2}} z^{iL} \qquad \text{MOD } \frac{L+1}{2}L \tag{8-43}$$

or the sequence must satisfy

$$[S(z)]^{\frac{L-1}{2}} \sum_{i=o}^{\frac{L-1}{2}} z^i [z^{L-2i}]^{\frac{L+1}{2}} + z^x \sum_{i=o}^{\frac{L-1}{2}} z^{iL} = 0 \tag{8-44}$$

and

$$G(z) \left| \left[(S(z))^{\frac{L-1}{2}} \sum_{i=o}^{\frac{L-1}{2}} z^i \left[z^{L-2i}\right]^{\frac{L+1}{2}} + z^x \sum_{i=o}^{\frac{L-1}{2}} z^{iL} \right] \right. \tag{8-45}$$

Also,

$$\frac{[S(z)]^{\frac{L-1}{2}} \sum_{i=o}^{\frac{L-1}{2}} z^i[z^{L-2i}]^{\frac{L+1}{2}} + z^x \sum_{i=o}^{\frac{L-1}{2}} z^{iL}}{G(z)} =$$

$$[1+z^L]^{\frac{L-1}{2}} \sum_{i=o}^{\frac{L-1}{2}} z^i[z^{L-2i}]^{\frac{L+1}{2}} + z^x G^{\frac{L-1}{2}}(z) \sum_{i=o}^{\frac{L-1}{2}} z^{iL} /[G(z)]^{(L+1)/2} \qquad (8\text{-}46)$$

For example if

$$G(z) = 1+z+z^2 \qquad (8\text{-}47)$$

$$S(z) = 1+z : 110 \qquad (8\text{-}48)$$

$$L = 3 \qquad (8\text{-}49)$$

and

$$X = 2 : 101 \qquad (8\text{-}50)$$

for

$$r = 2 \qquad (8\text{-}51)$$

Evaluating the terms in (8-41) yields:

$$(1+z^L)^{\frac{L-1}{2}} = 1 + z^3 \qquad (8\text{-}52)$$

$$\sum_{i=o}^{\frac{L-1}{2}} z^i[z^{L-2i}]^{\frac{L+1}{2}} = \sum_{i=o}^{1} z^i[z^{3-2i}]^2 = 1 + z^3 \quad \text{MOD } 2L \qquad (8\text{-}53)$$

$$[G(z)]^{\frac{L-1}{2}} = G(z) = 1 + z + z^3 \qquad (8\text{-}54)$$

$$[G(z)]^{\frac{L+1}{2}} = [G(z)]^2 = 1 + z^2 + z^4 \qquad (8\text{-}55)$$

Equation (8-46) becomes

$$\frac{(1+z^3)(1+z^3) + z^2(1+z+z^7)(1+z^3)}{1+z^2+z^4} = \frac{(1+z^3)(1+z^2+z^4)}{1+z^2+z^4} = 1+z^3 \qquad (8\text{-}56)$$

Notice that MOD $\frac{(L+1)}{2}$ L, or MOD 6, arithmetic was not used in (8-56).

As a second example, if

$$G(z) = 1 + z^2 + z^3 \qquad (8\text{-}57)$$

$$S(z) = 1 + z^2 + z^3 + z^4 \ : \ 1011100 \qquad (8\text{-}58)$$

$$L = 7 \qquad (8\text{-}59)$$

and

$$X = 5 \qquad (8\text{-}60)$$

Corresponding to the sequence 1110010 formed by sampling.

Evaluating the terms in (8-41)

$$(1+z^L)^{\frac{L-1}{2}} = (1+z^7)^3 = 1 + z^7 + z^{14} + z^{21} \qquad (8\text{-}61)$$

$$\sum_{i=o}^{\frac{L-1}{2}} z^i \, (z^{L-2i})^{\frac{L+1}{2}} = \sum_{i=o}^{3} z^i (z^{7-2i})^4$$

$$= 1+z^7+z^{14}+z^{21} \qquad (8\text{-}62)$$

$$(G(z))^{\frac{L-1}{2}} = (1+z^2+z^3)^3 = 1+z^2+z^3+z^4+z^7+z^8+z^9 \qquad (8\text{-}63)$$

$$\sum_{i=o}^{\frac{L-1}{2}} z^{iL} = \sum_{i=o}^{3} z^{i7} = 1+z^7+z^{14}+z^{21} \qquad (8\text{-}64)$$

and

$$[G(z)]^{\frac{L+1}{2}} = (1+z^2+z^3)^4 = 1+z^8+z^{12} \quad (8\text{-}65)$$

Equation (41) becomes

$$\frac{(1+z^7+z^{14}+z^{21})(1+z^7+z^{14}+z^{21})+z^5(1+z^2+z^3+z^4+z^7+z^8+z^9)(1+z^7+z^{14}+z^{21})}{1+z^8+z^{12}} =$$

$$\frac{(1+z^7+z^{14}+z^{21})(1+z^5+z^8+z^9+z^{12}+z^{13}+z^{21})}{1+z^8+z^{12}} = (1+z^7+z^{14}+z^{21}) \cdot (1+z^5+z^9) \quad (8\text{-}66)$$

Notice for these two examples, (8-46) reduces to finding X such that

$$\frac{(1+z^L)^{\frac{L-1}{2}} + z^x(G(z))^{\frac{L-1}{2}}}{[G(z)]^{\frac{L+1}{2}}} \quad (8\text{-}67)$$

is a rational fraction.

As another example, consider

$$G(z) = 1+z+z^3 \quad (8\text{-}68)$$

$$S(z) = 1+z+z^2+z^4\text{: } 1110100 \quad (8\text{-}69)$$

$$L = 7 \quad (8\text{-}70)$$

and

$$X = 0 \quad (8\text{-}71)$$

Corresponding to the sequence 110100 formed by sampling.

Evaluating the terms in (46)

$$[G(z)]^{\frac{L-1}{2}} = (1+z+z^3)^3 = 1+z+z^2+z^5+z^6+z^7+z^9 \quad (8\text{-}72)$$

and

$$(G(z))^{\frac{L+1}{2}} = 1+z^4+z^{12} \quad (8\text{-}73)$$

Equation (46) becomes

$$\frac{(1+z^7+z^{14}+z^{21})(1+z^7+z^{14}+z^{21}+z^x)(1+z+z^2+z^5+z^6+z^7+z^9)}{(1+z^4+z^{12})} \quad (8\text{-}74)$$

Now if X=0 this becomes

$$\frac{(1+z^7+z^{14}+z^{21})(z)(1+z+z^4+z^5+z^8+z^{13}+z^{20})}{1+z^4+z^{12}} =$$

$$(1+z^7+z^{14}+z^{21})(z)(1+z+z^8) \tag{8-75}$$

The algorithm for finding the shift X for a sampling rate S=2 (corresponding to sampling every other bit) is to find an integer X such that

$$\frac{(1+z^L)^{\frac{L-1}{2}} + z^x\,(G(z))^{\frac{L-1}{2}}}{(G(z))^{\frac{L+1}{2}}}$$

is rational.

Now for the general case

$$S=2^{\vartheta} \tag{8-76}$$

ϑ an integer,

the generalization of (43) becomes

$$\sum_{i=o}^{\frac{L+1}{S}-1} z^i[z^{L-Si}(S(z)]^{\frac{L+1}{3}} = z^x S(z) \sum_{i=o}^{\frac{L+1}{S}} z^{iL} \qquad \text{MOD } \frac{L+1}{S}L \tag{8-77}$$

For example with

$$G(z) = 1+z^2+z^3 \tag{8-78}$$

and

$$S(z) = 1+z^2+z^3+z^4 \; : \; 1011100 \tag{8-79}$$

Sampling at the rate

$$S=4 \tag{8-80}$$

yields

$$1100101: \; 1+z+z^4+z^6$$

or

$$X=4. \tag{8-81}$$

Equation (8-77) becomes

$$\sum_{i=o}^{1} z^i[z^{7-4i}][S(z)]^2 = z^x S(z)\sum_{i=o}^{1} z^{7i} \tag{8-82}$$

Evaluating the term in (8-82),

$$S^2(z) = 1 + z + z^4 + z^6 \tag{8-83}$$

$$\sum_{i=o}^{1} z^i\ [z^{7-4i}]^2 = z^{14}+z^7 = 1 + z^7 \tag{8-84}$$

$$\sum_{i=o}^{1} z^{7i} = 1 + z^7 \tag{8-85}$$

Substituting into (8-82)

$$(1+z+z^4+z^6)(1+z^7) = z^4(1+z^2+z^3+z^4)(1+z^7) \quad \text{MOD}(7)2 \tag{8-86}$$

or

$$(1+z+z^4+z^6) = z^4(1+z^2+z^3+z^4) \quad \text{MOD}(7) \tag{8-87}$$

Equation (8-77) reduces in general to

$$[S(z)]^{\frac{L+1}{S}} = z^x S(z) \quad \text{MOD L} \tag{8-88}$$

in general

or

$$G(z) \,|\, ((S(z))^{\frac{L+1}{S}} + z^x S(z))$$

and

$$G(z) \,|\, ((S(z))^{\frac{L+1}{S}-1} + z^x)$$

Now

$$\frac{[S(z)]^{\frac{L+1}{S}-1} + z^x}{G(z)} = \frac{(1+z^L)^{\frac{L-S+1}{S}} + z^x(G(z))^{\frac{L-S+1}{S}}}{[G(z)]^{\frac{L+1}{S}}} \tag{8-89}$$

and the fraction in the latter portion of (8-89) must be rational.

For example,

$$G(z) = 1 + z^2 + z^3 \tag{8-90}$$

$S=4$

$$\frac{(1+z^7)^1 + z^4(G(z))}{(G(z))^2} = \frac{1+z^4+z^6}{1+z^4+z^6} = 1 \tag{8-91}$$

The table below is intended to summarize the use of equation (8-89) for several example sequences.

G(z)	L	S	X	Ratio From (89)
$1 + z + z^2$	3	2	2	1
$1 + z^2 + z^3$	7	2	5	$(1+z^5+z^9)$
$1 + z^2 + z^3$	7	4	4	1
$1 + z + z^3$	7	2	0	$z(1+z+z^8)$
$1 + z + z^3$	7	4	0	z
$1 + z + z^4$	15	2	8	—
$1 + z + z^4$	15	4	12	$1+z^4+z^8+z^{13}+z^{14}+z^{17}+z^{29}$
$1 + z + z^4$	15	8	14	$1+z^2+z^4+z^6+z^{10}$
$1 + z^3 + z^4$	15	2	9	—
$1 + z^3 + z^4$	15	4	6	$1+z^6+z^9+z^{10}+z^{14}+z^{17}+z^{21}+z^{25}+z^{29}$
$1 + z^3+z^4$	15	8	12	$1+z^6+z^8$

9. Channel Capacity and Signal Spectral Density

The distribution of signal energy in a communication channel affects the channel capacity of the system. The design of the TDRS-HEAO-C modem should maximize channel capacity by controlling signal spectral density.

The channel capacity for a channel with signal power spectral density S(f) is

$$C = \int_0^{\infty} \log_2 \left[1 + \frac{S(f)}{N(f)}\right] df \qquad \text{bits/sec} \qquad (9\text{-}1)$$

where N(f) is the one-sided noise power spectral density. If the signal is confined to a finite bandwidth B, and constrained to

$$\int_0^B S(f)df = S \qquad (9\text{-}2)$$

where S is the total signal power. For a general N(f), the calculus of variations provides an approach to bounding the function, or class of functions, S(f), constrained by (9-2) to maximize the integral in (9-1).

Define

$$\rho(f) = \frac{S(f)}{N(f)} \qquad (9\text{-}3)$$

where $\rho(f)$ is the continuous contrast ratio. An extrema can be found by forming an auxiliary function of the form of the Hamiltonian

$$H = \log_2[1+\rho(f)] + \lambda\rho(f)\, N(f) \qquad (9\text{-}4)$$

Substituting (9-4) into the Euler differential equation

$$\frac{d}{df}\left(\frac{\partial H}{\partial \rho}\right) - \frac{H}{\partial \rho} = 0 \qquad (9\text{-}5)$$

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or from (9-4)

$$\frac{\log_2 e}{1+\rho(f)} + \lambda N(f) = 0 \tag{9-6}$$

or

$$\rho(f) = -1 - \frac{\log_2 e}{\lambda N(f)} \tag{9-7}$$

Now to evaluate λ

$$\int_0^B N(f)\rho(f)df = \int_0^B -\left(N(f) + \frac{\log_2(e)}{\lambda}\right)df = S \tag{9-8}$$

or

$$S = -\int_0^B N(f)df - \frac{\log_2(e)}{\lambda} B \; . \tag{9-9}$$

Now define the integral

$$\int_0^B N(f)df = N \tag{9-10}$$

where N is the total noise power in bandwidth B.

Then

$$S = -N - \frac{\log_2(e)}{\lambda} B \tag{9-11}$$

or

$$\lambda = \frac{-\log_2(e)B}{(S+N)} \tag{9-12}$$

The equation for contrast ratio is from (9-7) and (9-12)

$$g(f) = -1 + \frac{S+N}{BN(f)} \qquad (9\text{-}13)$$

or

$$S(f) = \frac{S+N}{B} - N(f) \qquad (9\text{-}14)$$

For the case

$$N(f) = N_o \qquad (9\text{-}15)$$

equation (9-14) becomes

$$S(f) = \frac{S}{B}$$

and the maximum channel capacity becomes

$$C = \log_2\left[1 + \frac{S}{N}\right] B \qquad \text{bits/sec.} \qquad (9\text{-}16)$$

In the case of equation (9-14) the maximum channel capacity is

$$C = \int_o^B \log_2 \left[\frac{S+N}{BN(f)}\right] df \qquad \text{bits/sec.} \qquad (9\text{-}17)$$

Now assuming a constrant on the noise energy in finite bandwith B,

$$\int_o^B N(f) df = N \qquad (9\text{-}18)$$

The calculus of variations approach can be used to find an extrema of the variational problem described by (9-17) and the constrant (9-18). The auxilary equation can be formed as

$$H = \log_2\left[\frac{S+N}{BN(f)}\right] + \lambda N(f) \qquad (9\text{-}19)$$

Now define

$$P = \frac{S+N}{B} \qquad (9\text{-}20)$$

as the average energy per unit bandwidth in the communications channel of bandwidth B. This includes both signal and noise energy. Substitution of (9-19) into Euler's differential equation yields

$$\frac{N(f)}{P}\left(\frac{-P}{N(f)^2}\right)+\lambda=0 \tag{9-21}$$

or

$$N(f) = \frac{1}{\lambda} \tag{9-22}$$

The Lagrange multiplier can be evaluated by substituting (9-22) into (9-18).

$$\int_0^B \frac{df}{\lambda} = N \tag{9-23}$$

or

$$\lambda = \frac{B}{N} \tag{9-24}$$

and

$$N(f) = \frac{N}{B} \tag{9-25}$$

is an extreme of the integral in (9-17).

The form for S(f) from (9-14) is given for an example noise spectral density N(f) in figure (9-1)

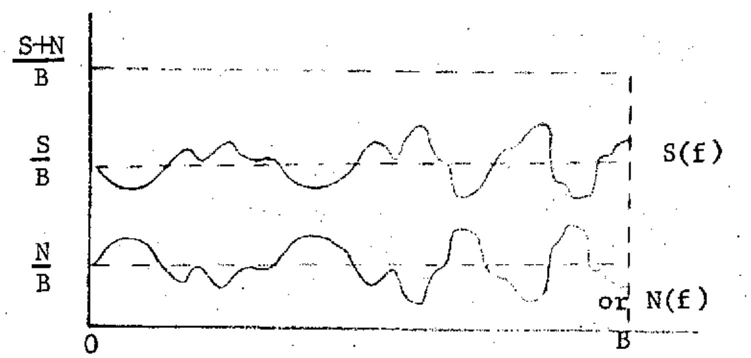


Figure 9-1 Signal and noise spectral density illustrating the reflection effect in the relation of N(f) to S(f) for maximum channel capacity.

To show that the calculation resulting by equation (9-25) is a minimum extrema consider the following example :

Case 1: B = 1MHZ, $N(f) = .5 \times 10^{-8}$ WATTS/HZ

$$N = \int_{o}^{B} N(f)df = .5 \times 10^{-2} = 5 \times 10^{-3} \text{ WATTS}$$

Assume $S = 20 \times 10^{-3}$ WATTS

$$S(f) = \frac{25 \times 10^{-3}}{10^{+6}} - N(f) \qquad \text{from (14)}$$

or

$$S(f) = 25 \times 10^{-9} - 5 \times 10^{-9} = 20 \times 10^{-9} \text{ WATTS/HZ}$$

From (9-17)

$$C = \int_{o}^{B} log_2[\frac{S+N}{B} \cdot \frac{1}{NH}]df \qquad \text{bit/sec}$$

or

$$C = \int_{o}^{B} log_2 [\frac{25 \times 10^{-9}}{.5 \times 10^{-8}}]df = \int_{o}^{B} lo_2 [50 \times 10^{-1}] df$$

$$= log_2(5) \times 10^6 \text{ bits/sec}$$

$$\frac{loge(5)}{loge(2)} \times 10^{-6} = \frac{1.61}{.693} \times 10^6 = 2.32 \times 10^6 \text{ bits/sec}$$

Case 2 ssume some total noise power and total signal power in a different distribution across the same bandwidth as in Case 1.

$N = 5 \times 10^{-3}$ WATTS, $S = 20 \times 10^{-3}$ WATTS

$N(f) = .1 \times 10^{-8} + f(8x10^{-15})$ WATTS/HZ

$$\int_{o}^{10^6} (.1 \times 10^{-8} + 8 \times 10^{-15} f)df = .1 \times 10^{-8}(10^6) + \frac{8 \times 10^{-15}(10^{12})}{2}$$

$$= 1 \times 10^{-3} + 4 \times 10^{-3}$$

$$= 5 \times 10^{-3}$$

From (9-14)

$$S(f) = \frac{25 \times 10^{-3}}{10^{6}} - N(f)$$

$$= 25 \times 10^{-9} - 1 \times 10^{-9} - f(8x10^{-15})$$

$$= 24 \times 10^{-9} - f\ (8x10^{-15})$$

From (9-17)

$$C = \int_{o}^{10^6} \log_2 \left[\frac{25 \times 10^{-9}}{(1+8x10^{-6})10^{-9}} \right] df$$

$$= \int_{o}^{10^6} \log_2 \left[\frac{25}{(1+8x10^{-6}f)} \right] df$$

$$= \int_{o}^{10^6} \frac{\log e \dfrac{25}{1+8x10^{-6}f} \ df}{\log_e 2}$$

$$= \frac{1}{\log_2 e} \int_{o}^{10^6} \log e \left[\frac{25}{1+8x10^{-6}f} \right] df$$

$$= K \int_{o}^{B} \log e \left(\frac{a}{1+bf}\right) df$$

Let $V = \dfrac{a}{1+bf}$

$$dv = \frac{-ab}{(1+bf)^2} = \frac{-b}{a} V^2 df$$

Then

$$C = \frac{-ak}{b} \int_{B'}^{B''} \frac{\log e(v)du}{v^2}$$

where

$$B'' = \frac{a}{1+bB}$$

$$B' = a$$

or

$$C = \frac{-ak}{b}\left[v^{-1}\left(\frac{\log e(v)}{-1} - 1\right)\right]\Bigg|_{B'}^{B''}$$

$$= \frac{-ak}{b}\left[\frac{1}{v}\ (\log e(v)+1)\right]\Bigg|_{B'}^{B''}$$

$$= \frac{ak}{b}\left[\frac{1}{B''}\ (\log e(B'')+1) - \frac{1}{B'}\ (\log e(B')+1)\right]$$

Channel capacity is then

$$B'' = \frac{25}{1+8x10^{-6}(10^{+6})} = \frac{25}{9}$$

$$B' = 25$$

$$\frac{25}{\log e^2(8x10^{-6})}\left[\frac{9}{25}\ (\log e\ (\frac{25}{9})+1)\right.$$

$$\left. - \frac{1}{25}\ (\log e(25)+1)\right]$$

$$= \frac{25x10^6}{[\log e(2)][8]} \left[\frac{9}{25}[2.02] - \frac{1}{25}[4.22]\right]$$

$$= \frac{10^6}{8\log e(2)} [18.18 - 4.22]$$

$$= \frac{10^6}{8\ (.694)} [13.96]$$

$$= 2.52 \times 10^6 \quad \text{bits/sec}$$

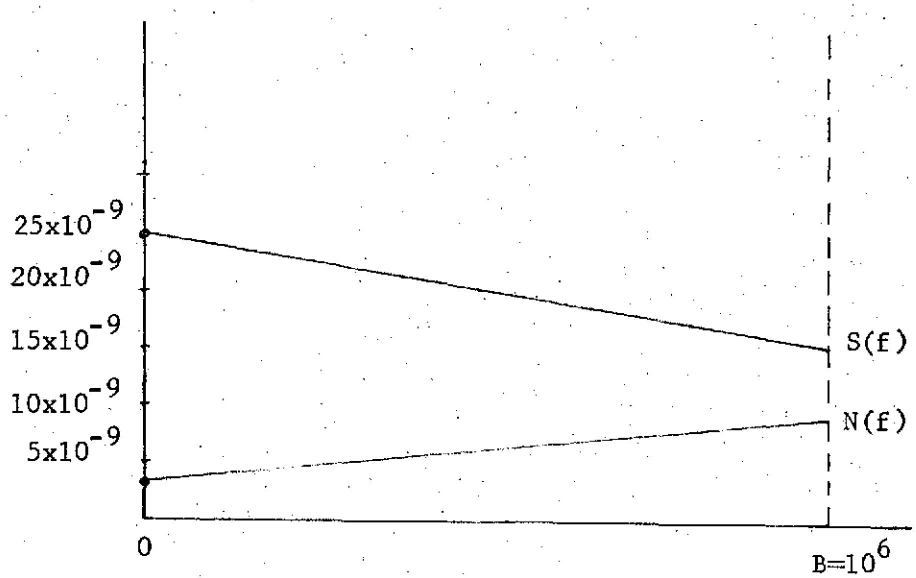
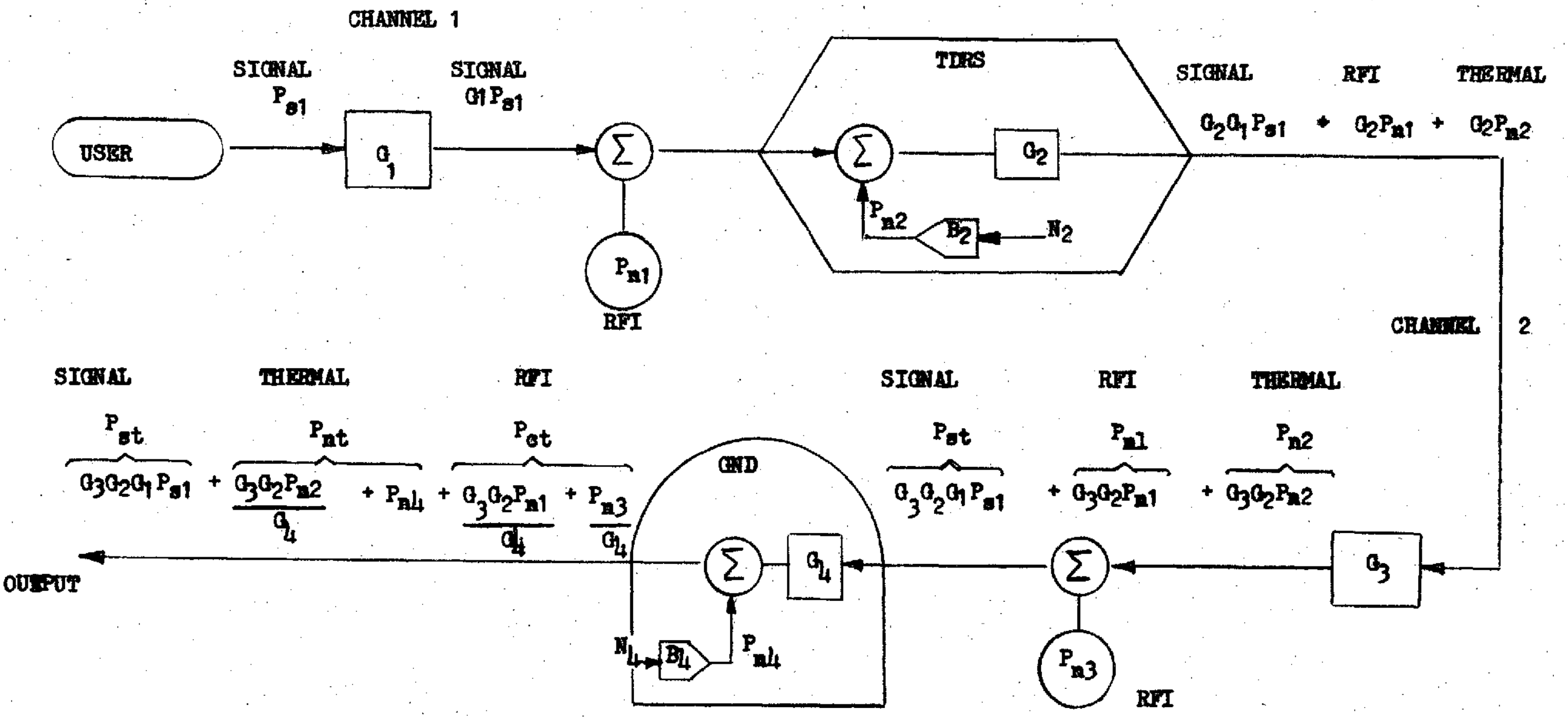


Figure 9-2 Example Results

10 Power Budget Program

A computer program designed to perform power budget calculations for the HEAO-C-TDRS- Ground Return Link has been prepared. The schematic diagram of the link along with definitions of parameter values is given in figure 10-1. This computer program is expected to be used extensively in further HEAO-C-TDRS communication study efforts.

FIGURE 10-1 SIMPLIFIED RETURNED LINK MODEL: USER ——→ TDRS ——→ GND



P_{s1} = Transmitting Signal Power of User

G_1 = Gain (or loss) of Channel 1

P_{n1} = Received RFI and Multipath Power at TDRS

G_2 = Gain (or Loss) of TDRS

B_2 = Receiver Bandwidth of TDRS

N_2 = Receiver Thermal Noise Spectral Density of TDRS

$P_{n2} = N_2 B_2$ = Receiver Thermal Noise Power of TDRS

G_3 = Gain (or Loss) of Channel 2

P_{n3} = Received RFI and Multipath Power at GND Station

G_4 = Processing gain of GND Receiver

B_4 = Receiver Bandwidth of GND Station

N_4 = Receiver Thermal Noise Spectral Density of GND Station

$P_{n4} = N_4 B_4$ = Receiver Thermal Noise Power of GND Station

P_{st} = Total Received (And Outputted) Signal Power of GND Station

P_{nt} = Total Thermal Noise Power Output of GND Station

P_{ot} = Receiver output RFI Spread Interference

POWER BUDGET CALCULATIONS

```
      PROGRAM LAWEA(INPUT,OUTPUT,TAPE1=INPUT,TAPE3=OUTPUT)
      COMMON  PAR(100),TCL(100),NCS
      DIMENSION A(59,16), B(59,12), C(59), D(59,6)
      DIMENSION MINUS(13)
      REAL  NPV,NT
      INTEGER C
C
C
C
C     NCASE1 =NRFI,NSS,NPG
C     NCASE2 =NRFI,SS,PG
C     NCASE3 =RFI,NSS,NPG
C     NCASE4 =RFI,SS,PG
C          FOUR CONFIGURATION CASES  ARE CONSIDERED . EACH CASE REPRESENTS SOME
C          COMBINATION OF BASIC PARAMETERS AND VARIABLES WHICH INCLUDE RADIO
C          FREQUENCY INTEREFERENCE (RFI OR NRFI),SPREAD SPECTRUM (SS OR NSS),
C          PROCESSING GAIN (PG OR NPG)
C
C
      DO 7555 K=1,13
 7555 MINUS(K)=10H----------
      WRITE(3,9994) (MINUS(N),N=1,13)
      DO 6 I=1,59
    6 READ(1,101)(A(I,J),J=1,16)
  101 FORMAT(16A4)
      DO 7 I = 1,59
    7 READ(1,102)(B(I,J),J=1,12)
  102 FORMAT(11A4,A3)
      READ(1,103)C
  103 FORMAT(59(I1))
      DO 9 I = 1,59
    9 READ(1,104)(D(I,J),J=1,6)
  104 FORMAT(5A4,A2)
      WRITE(3,9995)
      DO 8000 K=1,5
 8000 WRITE(3,9999)
 9999 FORMAT(1H )
      WRITE(3,9998)
 9998 FORMAT(1H ,50X,≠SELECTED SATELLITE RELAY CASES≠,
     1/,51X,≠FOR THE USER TELEMETRY CHANNEL≠)
      DO 8001 I=1,5
 8001 WRITE(3,9999)
      WRITE(3,5)
    5 FORMAT(1H ,55X,≠CASE 1 = NRFI,NSS,NPG≠,///,56X,≠CASE 2 = NRFI,SS,P
     1G≠,///,56X,≠CASE 3 = RFI,NSS,NPG≠,///,56X,≠CASE 4 = RFI,SS,PG≠)
      DO 8002 K=1,10
 8002 WRITE(3,9999)
      WRITE(3,9997)
 9997 FORMAT(1H ,61X,≠PREPARED BY≠,/,56X,≠ALABAMA A-M UNIVERSITY≠)
      READ(1,10)NC,NS,NP,NPV,NT
   10 FORMAT (I1,I2,I2,F8.2,F8.2)
   15 NCS =NC
      NSS =NS
      PAR(NP) =NPV
      TOL(NP) =NT
   16 READ(1,20)NC,NS,NP,NPV,NT
   20 FORMAT(I1,I2,I2,F8.2,F8.2)
```

```
      IF(NCS-NC)40,25,40
   25 IF(NSS-NS)40,35,40
   35 PAR(NP) =NPV
      TOL(NP) = NT
      GO TO 16
 40   WRITE(3,9995)
      DO 8003 K=1,10
 8003 WRITE(3,9999)
      WRITE(3,9996)
 9996 FORMAT(1H ,56X,≠KEY PARAMETER VALUES≠,/,51X,≠FOR THE USER TELEMETR
     1Y CHANNEL≠)
      WRITE(3,50) NCS,NSS,PAR(13),PAR(16),PAR(35)
   50 FORMAT(///,44X,≠CASE≠,4X,≠SET≠,4X,≠PARAMETER(13)≠,4X,≠PARAMETER(16
     1)≠,4X,≠PARAMETER(35)≠,///,44X,I1,7X,I2,8X,F8.2,7X,F8.2,5X,F8.2)
      IPAGE=1
      CALL USERLK
      CALL  TDRSLK
      CALL GRDLK
      WRITE(3,9995)
 9995 FORMAT(1H1)
      WRITE(3,9994)(MINUS(N),N=1,13)
 9994 FORMAT(13A10,/)
C  PLACE MINUS IN HEADERS
C
      WRITE(3,90) IPAGE,NCS,NSS
   90 FORMAT(1H ,60X,≠PAGE≠,1X,I1,//≠ CASE≠,1X,I1,≠.≠,3X,≠POWER BUDGET C
     1ALCULATIONS FOR RETURN LINK PERFORMANCE   USER---TDRS---GROUND≠//
     2≠ SET≠,2X,I1,≠.≠,3X,≠CONDITIONS    NRFI,NSS,NPG≠///6X,≠HIGH DATA RA
     3TE(HDR)≠,9X,≠USER TELEMETRY≠/////≠ I.D.≠,21X,≠PARAMETER≠,31X,≠PARA
     4METER≠,7X,≠PARAMETER≠,13X,≠TOLERANCE≠,10X,≠TOLERANCE≠/≠ CODE≠,
     561X,≠NUMBER≠,10X,≠VALUE≠,17X,≠NUMBER≠,13X,≠VALUE≠)
      WRITE(3,9994)  (MINUS(N),N=1,13)
      WRITE(3,9993)
 9993 FORMAT(1H ,9X,≠***  USERLK  ***≠,/)
      DO 500 I=1,59
      IF(PAR(I)-999.) 60,95,60
 60   IF(TOL(I)-5000.) 80,70,80
 80   IF(C(I)-1.)210,201,210
 201  JFLAG=NCS*10000+NSS*100+I
      WRITE(3,1000)  (JFLAG,(A(I,J),J=1,16),PAR(I),(D(I,J),J=1,6),
     1TOL(I),(B(I,J),J=1,12))
 1000 FORMAT(1H ,I5,3X,16A4,8X,F8.1,1X,5A4,A2,7X,F8.1,1X,≠DB≠/9X,
     1 11A4,A3//)
      GO TO 800
 210  JFLAG=NCS*10000+NSS*100+I
      WRITE(3,2000)  (JFLAG,(A(I,J),J=1,16),PAR(I),(D(I,J),J=1,6),
     1 TOL(I))
 2000 FORMAT(1H ,I5,3X,16A4,8X,F8.1,1X,5A4,A2,7X,F8.1,1X,≠DB≠//)
      GO TO 800
 70   JFLAG=NCS*10000+NSS*100+I
      IF(C(I)-1.) 310,300,310
 300  CONTINUE
      WRITE(3,3000)  (JFLAG,(A(I,J),J=1,16),PAR(I),(D(I,J),J=1,6),
     1 (B(I,J),J=1,12))
 3000 FORMAT(1H ,I5,3X,16A4,8X,F8.1,1X,5A4,A2/9X,11A4,A3//)
      GO TO 800
 310  CONTINUE
```

```
      WRITE(3,4000) JFLAG,(A(I,J),J=1,16),PAR(I),(D(I,J),J=1,6)
 4000 FORMAT(1H ,I5,3X,16A4,8X,F8.1,1X,5A4,A2//)
      GO TO 800
   95 JFLAG=NCS*10000+NSS*100+I
      IF(C(I)-1.) 610,600,610
  600 CONTINUE
      DSAVE=D(I,1)
      ESAVE=D(I,2)
      D(I,1)=10H
      D(I,2)=10H
      WRITE(3,5000)(JFLAG,(A(I,J),J=1,16),(D(I,J),J=1,6),
     1   (B(I,J),J=1,12))
      D(I,1)=DSAVE
      D(I,2)=ESAVE
      IF(JFLAG.EQ.30151) WRITE(3,9980)
 9980 FORMAT(1H ,≠ITEM 51 REPRESENTS THE SUM OF THERMAL NOISE SPECTRAL D
     1ENSITY AND SPREAD COHERENT INTERFERENCE SPECTRAL DENSITY. FOR THE
     2REASONS GIVEN≠,/,1X,≠ IN APPENDIX C (NOTE 4) ITEM 51 IS NOT APPLIC
     3ABLE. AS A RESULT, ITEMS 53, 57, AND 59 ARE ALSO NOT APPLICABLE. T
     4HIS DOES NOT PRECLUDE≠,/,1X,≠A CASE-BY-CASE ANALYSIS TO TAKE INTO
     5ACCOUNT THE EFFECT OF RFI.≠)
 5000 FORMAT(1H ,I5,3X,16A4,9X,8H   NA    ,1X,5A4,A2,/
     18X,11A4,A3//)
      GO TO 800
  610 CONTINUE
      DSAVE=D(I,1)
      ESAVE=D(I,2)
      D(I,1)=10H
      D(I,2)=10H
      WRITE(3,6000)(JFLAG,(A(I,J),J=1,16),(D(I,J),J=1,6))
      D(I,1)=DSAVE
      D(I,2)=ESAVE
      IF(JFLAG.EQ.30151) WRITE(3,9980)
 6000 FORMAT(1H ,I5,3X,16A4,9X,8H   NA    ,1X,5A4,A2//)
  800 IF(I-9)7000,900,810
  810 IF(I-14)7000,910,820
  820 IF(I-20) 7000,900,830
  830 IF(I-33) 7000,900,840
  840 IF(I-40) 7000,920,850
  850 IF(I-42) 7000,900,860
  860 IF(I-51) 7000,900,7000
  900 IPAGE=IPAGE+1
      WRITE(3,9995)
      WRITE(3,9994) (MINUS(N),N=1,13)
      WRITE(3,950) IPAGE
  950 FORMAT(1H ,60X,≠PAGE≠,1X,I1,//,≠ I.D.≠,21X,≠PARAMETER≠,31X,
     1≠PARAMETER≠,7X,≠PARAMETER≠,13X,≠TOLERANCE≠,10X,≠TOLERANCE≠,/,
     2≠ CODE≠
     2,91X,≠NUMBER≠,10X,≠VALUE≠,17X,≠NUMBER≠,13X,≠VALUE≠,
     3)
      WRITE(3,9994) (MINUS(N),N=1,13)
      IF(IPAGE-6) 500,500,450
  450 GO TO 160
  910 WRITE(3,951)
  951 FORMAT(1H ,///,9X,≠***  TDRSLK  ***≠,/)
      GO TO 500
  920 WRITE(3,952)
```

```
  952 FORMAT(1H ,///,9X,≠*** GRDLK ****≠,/)
 7000 CONTINUE
  500 CONTINUE
      IF(PAR(59)-999.)107,180,107
  107 IF(PAR(59)-5.)130,110,110
  110 WRITE(3,125)NCS,NSS
  125 FORMAT(///,≠ PERFORMANCE MARGIN FOR NCS=  ≠,I1,3X,≠NSS=  ≠,I2,2X,
     1 ≠IS ADEQUATE≠)
      GO TO 160
  130 WRITE(3,150)NCS,NSS
  150 FORMAT(///,≠ PERFORMANCE MARGIN FOR NCS=  ≠,I1,3X,≠NSS=  ≠,I2,2X,
     1≠IS NOT ADEQUATE≠)
      GO TO 160
  180 WRITE(3,190)NCS,NSS
  190 FORMAT(///,≠ PERFORMANCE MARGIN FOR NCS=  ≠,I1,3X,≠NSS=  ≠,I2,2X,
     1 ≠IS NOT APPLICABLE≠)
  160 IF(NC -5)15,200,200
  200 STOP
      END
```

```
      SYMBOLIC REFERENCE MAP (R=1)

 ENTRY POINTS
  4102  LAWEA

 VARIABLES       SN  TYPE            RELOCATION
  5644  A             REAL      ARRAY                   7524  B           REAL       ARRAY
 11030  C             INTEGER   ARRAY                  11123  D           REAL       ARRAY
  5642  DSAVE         REAL                              5643  ESAVE       REAL
  5632  I             INTEGER                           5640  IPAGE       INTEGER
  5633  J             INTEGER                           5641  JFLAG       INTEGER
  5630  K             INTEGER                          11665  MINUS       INTEGER    ARRAY
  5631  N             INTEGER                           5634  NC          INTEGER
   310  NCS           INTEGER              / /          5636  NP          INTEGER
  5626  NPV           REAL                              5635  NS          INTEGER
  5637  NSS           INTEGER                           5627  NI          REAL
     0  PAR           REAL      ARRAY      / /           144  TOL         REAL       ARRAY     / /

 FILE NAMES           MODE
     0  INPUT                2036  OUTPUT                       0  TAPE1     FMT          2036  TAPE3     FMT

 EXTERNALS           TYPE    ARGS
        GRBLK                  0                          TORSLK              0
        USERLK                 0

 STATEMENT LABELS
  5014  5        FMT                        0  6                                0  7
     0  9                                5056  10      FMT                   4221  15
  4227  16                               5072  20      FMT                      0  25             INACTIVE
     0  35           INACTIVE            4241  40                            5126  50      FMT
     0  60           INACTIVE            4411  70                               0  80             INACTIVE
  5164  90       FMT                     4512  95                            4727  101     FMT
  4740  102      FMT                     4747  103     FMT                   4760  104     FMT
     0  107          INACTIVE               0  110         INACTIVE          5553  125     FMT
  4702  130                              5570  150     FMT                   4707  160
  4705  180                              5606  190     FMT                      0  200            INACTIVE
     0  201          INACTIVE            4352  210                              0  300            INACTIVE
  4461  310                                 0  450         INACTIVE          4671  500
     0  600          INACTIVE            4572  610                           4632  800
     0  610          INACTIVE               0  820         INACTIVE             0  830            INACTIVE
     0  840          INACTIVE               0  850         INACTIVE             0  860            INACTIVE
  4650  900                              4664  910                           4667  920
  5503  950      FMT                     5533  951     FMT                   5542  952     FMT
  5263  1000     FMT                     5306  2000    FMT                   5331  3000    FMT
  5352  4000     FMT                     5441  5000    FMT                   5463  6000    FMT
  4671  7000                                0  7555                             0  8000
     0  8001                                0  8002                             0  8003
  5375  9980     FMT                     5240  9993    FMT                   5154  9994    FMT
  5146  9995     FMT                     5106  9996    FMT                   5037  9997    FMT
  4775  9998     FMT                     4770  9999    FMT

 COMMON BLOCKS    LENGTH
         / /        201

 STATISTICS
   PROGRAM LENGTH    5606B     2950
   BUFFER LENGTH     4074B     2108
```

123A

STATISTICS

BLANK COMMON 3116 201

```
      SUBROUTINE  USERLK
      COMMON  PAR(100), TOL(100)
      REAL  NPV,NT
      F = 10357.
      D = 35800.
      PAR(5) = -(32.5+20.*ALOG10(F)+20.*ALOG10(D))
      TOL(5) = 2.7
      PAR(11) = -(TOL(2)+TOL(3)+TOL(4)+TOL(5)+TOL(6)+TOL(7)+TOL(8)+
     1 TOL(9)+TOL(10))
      PAR(12) = PAR(2)+PAR(3)+PAR(4)+PAR(5)+PAR(6)+PAR(7)+PAR(8)+PAR(9)+
     1  PAR(10)+PAR(11)
      PAR(14) = PAR(1)+PAR(12)
      RETURN
      END
```

```
      SYMBOLIC REFERENCE MAP (R=1)

ENTRY POINTS
    1  USERLK

VARIABLES        SN   TYPE              RELOCATION
   52  D              REAL                                   51  F          REAL
   47  NPV            REAL        *UNDEF                     50  NT         REAL        *UNDEF
    0  PAR            REAL         ARRAY    / /             144  TOL        REAL         ARRAY    / /

EXTERNALS             TYPE    ARGS
      ALOG10      REAL         1 LIBRARY

COMMON BLOCKS    LENGTH
      / /          200

STATISTICS
  PROGRAM LENGTH        53B        43
    BLANK COMMON       310B       200
```

```
      SUBROUTINE TDRSLK
      COMMON  PAR(100),TOL(100)
      REAL  NPV,NT,K
      K = -228.6
      T = 29.5
      PAR(15) = K+T
      PAR(17) = PAR(15)+PAR(16)
      PAR(18) = PAR(14)-PAR(15)
      PAR(20) = PAR(14)+PAR(19)
      PAR(21) = PAR(13)+PAR(19)
      PAR(22) = PAR(17)+PAR(19)
      F = 8250.
      D = 38000.
      PAR(27) = -(32.5+20.*ALOG10(F)+20.*ALOG10(D))
      TOL(27) = 0.2
      PAR(33) = -(TOL(23)+TOL(24)+TOL(25)+TOL(26)+TOL(27)+TOL(28)+
     1 TOL(29)+TOL(30)+TOL(31)+TOL(32))
      PAR(34) = PAR(23)+PAR(24)+PAR(25)+PAR(26)+PAR(27)+PAR(28)+PAR(29)+
     1   PAR(30)+PAR(31)+PAR(32)+PAR(33)
      PAR(36) = PAR(20)+PAR(34)
      PAR(37) = PAR(21)+PAR(34)
           X   = 10.**(PAR(37)/10.)
           Y   = 10.**(PAR(38)/10.)
           Z = X+Y
      PAR(39) = 10.*ALOG10(Z)
      PAR(40) = PAR(22)+PAR(34)
      RETURN
      END
```

```
SYMBOLIC REFERENCE MAP (R=1)

ENTRY POINTS
    1  TORSLK

VARIABLES      SN  TYPE          RELOCATION
  113  D           REAL                          112  F       REAL
  110  K           REAL                          106  KPV     REAL     *UNDEF
  107  NT          REAL      *UNDEF                0  PAR     REAL      ARRAY   / /
  111  T           REAL                          144  TOL     REAL      ARRAY   / /
  114  X           REAL                          115  Y       REAL
  116  Z           REAL

EXTERNALS          TYPE    ARGS
       ALOG10      REAL       1 LIBRARY

COMMON BLOCKS   LENGTH
      / /          200

STATISTICS
  PROGRAM LENGTH     117B     79
   BLANK COMMON      310B    200
```

1208A

```
      SUBROUTINE  GRDLK
      COMMON  PAR(100),TOL(100),NCS
      REAL  NPV,NT,K
      K = -228.6
      T = 20.0
      PAR(41) = K+T
      PAR(43) = PAR(41)+PAR(42)
         X    = PAR(40)-PAR(35)
         Y = 10.**(X/10.)
         Z    = 10.**(PAR  (43)/10.)
         W = Y+Z
      PAR(44) = 10.*ALOG10(W)
      PAR(45) = PAR(44)-PAR(42)
      PAR(46) = PAR(36)-PAR(45)
      PAR(47) = PAR(36)-PAR(44)
      PAR(49) = PAR(46)+PAR(48)
      NTEST = NCS-3
      IF(NTEST-0)50,150,50
150   PAR(50) = 999.
151   PAR(51) = 999.
      GO TO 251
50    PAR(50)=PAR(39)-(PAR(42)+PAR(35))
         P    = 10.**(PAR(45)/10.)
         Q    = 10.**(PAR(50)/10.)
         R = P+Q
      PAR(51) = 10.*ALOG10(R)
251   IF(NTEST-0)52,152,52
152   PAR(52) = PAR(36)+(PAR(48)-PAR(39))
153   PAR(53) = 999.
      GO TO 154
52    PAR(52) = 999.
      PAR(53) = PAR(36)+(PAR(48)-PAR(51))
154   PAR(55)=PAR(49)-PAR(54)
      IF(NTEST-0)156,56,156
56    PAR(56) = PAR(52)-PAR(54)
      GO TO 157
156   PAR(56) = 999.
      PAR(57) = PAR(53)-PAR(54)
      GO TO 158
157   PAR(57) = 999.
158   IF(NTEST-0) 159,59,159
159   PAR(59) = PAR(57)-PAR(58)
      GO TO 200
59    PAR(59) = 999.
200   RETURN
      END
```

```
        SYMBOLIC REFERENCE MAP (R=1)

    ENTRY POINTS
      1   GRDLK

    VARIABLES      SN  TYPE            RELOCATION
     124  K            REAL                                   310  NCS        INTEGER            / /
     122  NPV          REAL       *UNDEF                      123  NT         REAL       *UNDEF
     132  NTEST        INTEGER                                133  P          REAL
       0  PAR          REAL        ARRAY     / /              134  Q          REAL
     135  R            REAL                                   125  T          REAL
     144  TOL          REAL        ARRAY     / /              131  W          REAL
     126  X            REAL                                   127  Y          REAL
     130  Z            REAL

    EXTERNALS           TYPE    ARGS
          ALOG10   REAL          1 LIBRARY

    STATEMENT LABELS
      37  5[illegible]                          64  52                          0  56         INACTIVE
     111  5[illegible]                           0  150        INACTIVE         0  151        INACTIVE
       0  152        INACTIVE                    0  153        INACTIVE        71  154
      77  156                                  103  157                       105  158
       0  159        INACTIVE                  113  200                        55  251

    COMMON BLOCKS   LENGTH
          / /           201

    STATISTICS
      PROGRAM LENGTH      136B      94
        BLANK COMMON      311B     201
```

FWA OF THE LOAD 101
LWA+1 OF THE LOAD 25442

356628 WORDS WERE REQUIRED FOR LOADING

TRANSFER ADDRESS -- LAWEA 4203

NO. TABLE MOVES 54

PROGRAM AND BLOCK ASSIGNMENTS.

BLOCK	ADDRESS	LENGTH	FILE	PREFIX TABLE CONTENTS
LAWEA	101	11702	LGO	03/27/74 15.23.17. SCOPE 3.4 FTN 4.0P357 666X I OPT=1
USERLK	12003	53	LGO	03/27/74 15.23.22. SCOPE 3.4 FTN 4.0P357 666X I OPT=1
TORSLK	12056	117	LGO	03/27/74 15.23.25. SCOPE 3.4 FTN 4.0P357 666X I OPT=1
GRBLK	12175	136	LGO	03/27/74 15.23.30. SCOPE 3.4 FTN 4.0P357 666X I OPT=1
FORSYS=	12333	1032	SL-FORTRAN	09/03/73 15.38.17. SCOPE 3.4 COMPASS 3.73244 FORTRAN OBJECT LIBRARY UTILITIES.
GETFIL=	13365	33	SL-FORTRAN	09/03/73 15.38.37. SCOPE 3.4 COMPASS 3.73244 LOCATE AN FIT GIVEN A FILE NAME.
INPC=	13420	233	SL-FORTRAN	09/03/73 15.38.52. SCOPE 3.4 COMPASS 3.73244 FORMATTED READ FORTRAN RECORD.
KODER=	13653	1441	SL-FORTRAN	09/03/73 15.39.12. SCOPE 3.4 COMPASS 3.73244 OUTPUT FORMAT INTERPRETER.
KRAKER=	15314	1540	SL-FORTRAN	09/03/73 15.39.26. SCOPE 3.4 COMPASS 3.73244 FORMAT CRACKER FOR INPUT OPERATIONS.
OUTC=	17054	205	SL-FORTRAN	09/03/73 15.40.36. SCOPE 3.4 COMPASS 3.73244 FORMATTED WRITE FORTRAN RECORD.
ALOG=	17261	37	SL-FORTRAN	09/03/73 15.42.08. SCOPE 3.4 COMPASS 3.73244 COMPUTE THE LOG AND LOG_{10} OF X.
EXP=	17320	44	SL-FORTRAN	09/03/73 15.43.43. SCOPE 3.4 COMPASS 3.73244 EXPONENTIAL FUNCTION.
XTOY=	17364	7	SL-FORTRAN	09/03/73 15.44.32. SCOPE 3.4 COMPASS 3.73244 REAL TO REAL EXPONENTIATION.
/JMPS.RM/	17373	14		
LBUF.SG	17407	133	SL-SYSIO	09/03/73 16.44.11. SCOPE 3.4 COMPASS 3.73244
/CON.RM/	17542	6		
CIC.RM	17550	25	SL-SYSIO	09/03/73 16.45.12. SCOPE 3.4 COMPASS 3.73244
/ADR.RM/	17575	10		
MOVE.RM	17605	64	SL-SYSIO	09/03/73 16.45.18. SCOPE 3.4 COMPASS 3.73244
PCT.RM	17671	233	SL-SYSIO	09/03/73 16.45.21. SCOPE 3.4 COMPASS 3.73244
/OPEN.FO/	20124	7		
OPEN.RM	20133	301	SL-SYSIO	09/03/73 16.45.36. SCOPE 3.4 COMPASS 3.73244
OPEN.SG	20434	236	SL-SYSIO	09/03/73 16.45.50. SCOPE 3.4 COMPASS 3.73244
/PUT.RT/	20672	11		
RLEQ.RM	20703	42	SL-SYSIO	09/03/73 16.46.13. SCOPE 3.4 COMPASS 3.73244
WAR.SQ	20745	244	SL-SYSIO	09/03/73 16.46.55. SCOPE 3.4 COMPASS 3.73244
/CLSF.FO/	21211	7		
CLSF.RM	21220	23	SL-SYSIO	09/03/73 16.47.23. SCOPE 3.4 COMPASS 3.73244
/GET.RT/	21243	11		
Z.SQ	21254	76	SL-SYSIO	09/03/73 16.48.26. SCOPE 3.4 COMPASS 3.73244

BLOCK	ADDRESS	LENGTH	FILE	PREFIX TABLE CONTENTS
FSU.SQ	01352	111	SL-SYSIO	09/03/73 16.48.45. SCOPE 3.4 COMPASS 3.73244
ERR.RM	01463	367	SL-SYSIO	09/03/73 16.45.13. SCOPE 3.4 COMPASS 3.73244
CHW?.SG	02052	7	SL-SYSIO	09/03/73 16.45.20. SCOPE 3.4 COMPASS 3.73244
OPEX.SG	02061	21	SL-SYSIO	09/03/73 16.46.02. SCOPE 3.4 COMPASS 3.73244
/PUT.FO/	02102	7		
PUT.RM	02111	2	SL-SYSIO	09/03/73 16.46.12. SCOPE 3.4 COMPASS 3.73244
PUT.SQ	02113	1112	SL-SYSIO	09/03/73 16.46.16. SCOPE 3.4 COMPASS 3.73244
CLSF.SG	03225	136	SL-SYSIO	09/03/73 16.47.24. SCOPE 3.4 COMPASS 3.73244
/CLSV.FO/	03363	7		
CLSV.SQ	03372	127	SL-SYSIO	09/03/73 16.47.34. SCOPE 3.4 COMPASS 3.73244
/GET.FO/	03521	7		
/GET.RT/	03530	5		
GET.SQ	03535	727	SL-SYSIO	09/03/73 16.47.54. SCOPE 3.4 COMPASS 3.73244
RTRT.SG	04464	117	SL-SYSIO	09/03/73 16.48.51. SCOPE 3.4 COMPASS 3.73244
LXER.SG	04903	210	SL-SYSIO	09/03/73 16.49.25. SCOPE 3.4 COMPASS 3.73244
/SKFL.FO/	05013	7		
SKFL.SG	05022	50	SL-SYSIO	09/03/73 16.49.49. SCOPE 3.4 COMPASS 3.73244
SYS.RM	05072	37	SL-NUCLEUS	09/03/73 16.23.58. SCOPE 3.4 COMPASS 3.73244 PROCESS SYSTEM REQUEST.
//	05131	311		

2.035 CP SECONDS LOAD TIME

232

SELECTED SATELLITE RELAY CASES
FOR THE USER TELEMETRY CHANNEL

CASE 1 = NRFI,NSS,NPG

CASE 2 = NRFI,SS,PG

CASE 3 = RFI,NSS,NPG

CASE 4 = RFI,SS,PG

PREPARED BY
ALABAMA A-M UNIVERSITY

KEY PARAMETER VALUES
FOR THE USER TELEMETRY CHANNEL

CASE	SET	PARAMETER(13)	PARAMETER(16)	PARAMETER(35)
1	1	-300.00	66.00	-0.00

CASE 1. POWER BUDGET CALCULATIONS FOR RETURN LINK PERFORMANCE USER---TDRS---GROUND

SET 1. CONDITIONS NRFI,NSS,NPG

HIGH DATA RATE(HDR) USER TELEMETRY

I.D. CODE	PARAMETER		PARAMETER NUMBER	PARAMETER VALUE	TOLERANCE NUMBER	TOLERANCE VALUE
	*** USERLN ***					
10101	TOTAL TRANSMITTER POWER (1[illegible].0W)	(USER)	PAR(1)	10.0 DBW	TOL(1)	
10102	TRANSMITTING CIRCUIT LOSS	(USER)	PAR(2)	-1.0 DB	TOL(2)	0.0 DB
10103	TRANSMITTING ANTENNA GAIN (3-FOOT DISH)	(USER)	PAR(3)	37.9 DB	TOL(3)	.6 DB
10104	TRANSMITTING ANTENNA POINTING LOSS (1.0 DEG)	(USER)	PAR(4)	-2.5 DB	TOL(4)	.1 DB
10105	SPACE LOSS (F=1[illegible],357 MHZ, D=35 800 KM)	(RF LINK)	PAR(5)	-203.9 DB	TOL(5)	2.7 DB
10106	POLARZATION LOSS	(RF LINK)	PAR(6)	-0.0 DB	TOL(6)	.1 DB
10107	RECEIVING ANTENNA GAIN ([illegible]-FT DISH)	(TDRS)	PAR(7)	46.5 DB	TOL(7)	.6 DB
10108	RECEIVING ANTENNA POINTING LOSS (0.3 DEG)	(TDRS)	PAR(8)	-1.7 DB	TOL(8)	.1 DB
10109	RECEIVING ANTENNA SCAN LOSS	(TDRS)	PAR(9)	-0.0 DB	TOL(9)	1.0 DB

I.D. CODE	PARAMETER		PARAMETER NUMBER	PARAMETER VALUE	TOLERANCE NUMBER	TOLERANCE VALUE
10110	RECEIVING CIRCUIT LOSS	(TDRS)	PAR(10)	-1.0 DB	TOL(10)	.1 DB
10111	SOME OF ADVERSE TOLERANCES (2 THRU 10)	(TDRS)	PAR(11)	-5.3 DB	TOL(11)	
10112	NET CIRCUIT LOSS (SUM 2 THRU 11)	(TDRS)	PAR(12)	-131.0 DB	TOL(12)	
10113	RECEIVED RFI POWER (COHERENT)	(TDRS)	PAR(13)	-300.0 DB	TOL(13)	
10114	RECEIVED SIGNAL POWER (1+12)	(TDRS)	PAR(14)	-121.0 DBW	TOL(14)	
	*** TDRSLK ***					
10115	RECEIVER THERMAL NOISE SPECTRAL DENSITY (TS = 900 DEG K,NF =6)	(TDRS)	PAR(15)	-199.1 DBW/HZ	TOL(15)	
10116	RECEIVER BANDWIDTH (4.0 MHZ)	(TDRS)	PAR(16)	66.0 DB/HZ	TOL(16)	
10117	RECEIVER NOISE POWER (15+16)	(TDRS)	PAR(17)	-133.1 DBW	TOL(17)	
10118	RECEIVED SIGNAL POWER-TO-THERMAL NOISE SPECTRAL DENSITY (14-15)	(TDRS)	PAR(18)	78.1 DB	TOL(18)	
10119	TDRS GAIN	(TDRS)	PAR(19)	120.8 DB	TOL(19)	
10120	TRANSMITTED SIGNAL POWER (1.0W) (14+19)	(TDRS)	PAR(20)	-.2 DBW	TOL(20)	

I.D. CODE	PARAMETER		PARAMETER NUMBER	PARAMETER VALUE	TOLERANCE NUMBER	TOLERANCE VALUE
10121	TDRS TRANSMITTED COHERENT INTERFERENCE (13+19)	(TDRS)	PAR(21)	-179.2 DBW	TOL(21)	
10122	TDRS TRANSMITTED THERMAL NOISE POWER (17+19)	(TDRS)	PAR(22)	-12.3 DBW	TOL(22)	
10123	TRANSMITTING CIRCUIT LOSS	(TDRS)	PAR(23)	-.5 DB	TOL(23)	.1 DB
10124	TRANSMITTING ANTENNA GAIN (1-FT DISH)	(TDRS)	PAR(24)	26.4 DB	TOL(24)	.6 DB
10125	TRASMITTING ANTENNA POINTING LOSS	(TDRS)	PAR(25)	-0.0 DB	TOL(25)	.1 DB
10126	TRANSMITTING ANTENNA SCAN LOSS	(TDRS)	PAR(26)	-0.0 DB	TOL(26)	0.0 DB
10127	SPACE LOSS (F=6.250 MHZ, D=38,000 KM)	(RF LINK)	PAR(27)	-202.4 DB	TOL(27)	.2 DB
10128	POLARIZATION LOSS	(RF LINK)	PAR(28)	-0.0 DB	TOL(28)	.1 DB
10129	RECEIVING ANTENNA GAIN	(GRD)	PAR(29)	65.0 DB	TOL(29)	.2 DB
10130	RECEIVING ANTENNA POINTING LOSS	(GRD)	PAR(30)	-0.0 DB	TOL(30)	.1 DB
10131	RECEIVING ANTENNA SCAN LOSS	(GRD)	PAR(31)	-0.0 DB	TOL(31)	0.0 DB
10132	RECEIVING CIRCUIT LOSS	(GRD)	PAR(32)	-0.0 DB	TOL(32)	0.0 DB
10133	SOME OF ADVERSE TOLERANCES (23 THRU 32)	(GRD)	PAR(33)	-1.4 DB	TOL(33)	

I.D. CODE	PARAMETER		PARAMETER NUMBER	PARAMETER VALUE	TOLERANCE NUMBER	TOLERANCE VALUE
10134	NET CIRCUIT LOSS (SUM 23 THRU 33)	(GRD)	PAR(34)	-112.9 DB	TOL(34)	
10135	PROCESSING GAIN	(GRD)	PAR(35)	-0.0 DB	TOL(35)	
10136	TOTAL RECEIVED SIGNAL POWER (20+34)	(GRD)	PAR(36)	-113.1 DBW	TOL(36)	
10137	RECEIVED COHERENT INTERFERENCE FROM TDRS (21+34)	(GRD)	PAR(37)	-292.1 DBW	TOL(37)	
10138	RECEIVED COHERENT INTERFERENCE TDRS TO GND LINK	(GRD)	PAR(38)	-300.0 DBW	TOL(38)	
10139	TOTAL RFI POWER (COHERENT) (37+38) (NOTE.. DBW CAN NOT BE ADDED DIRECTLY)	(GRD)	PAR(39)	-291.5 DBW	TOL(39)	
10140	RECEIVED THERMAL NOISE POWER (22+34)	(GRD)	PAR(40)	-125.2 DBW	TOL(40)	
	*** GRDLK ***					
10141	RECEIVER THERMAL NOISE SPECTRAL DENSITY (TS =100 DEG K), (NF=DB)	(GRD)	PAR(41)	-208.6 DBW/HZ	TOL(41)	
10142	RECEIVER BANDWIDTH (4.0 MHZ)	(GRD)	PAR(42)	66.0 DB/HZ	TOL(42)	

I.D. CODE	PARAMETER		PARAMETER NUMBER	PARAMETER VALUE	TOLERANCE NUMBER	TOLERANCE VALUE
10143	RECEIVER THERMAL NOISE POWER (41+42)	(GRD)	PAR(43)	-142.6 DBW	TOL(43)	
10144	TOTAL THERMAL NOISE POWER OUTPUT (4[illegible]-35+43) (NOTE.. DBW CAN NOT BE ADDED DIRECTLY)	(GRD)	PAR(44)	-125.1 DBW	TOL(44)	
10145	ACTUAL RECEIVER THERMAL NOISE SPECTRAL DENSITY (44-42)	(GRD)	PAR(45)	-191.1 DBW/HZ	TOL(45)	
10146	RECEIVED SIGNAL POWER-TO-THERMAL NOISE SPECTRAL DENSITY (36-45)	(GRD)	PAR(46)	78.0 DB	TOL(46)	
10147	TOTAL RECEIVED SIGNAL POWER-TO-TOTAL THERMAL NOISE POWER (34-44)	(GRD)	PAR(47)	12.0 DB	TOL(47)	
10148	MODULATION LOSS (MODULATION LOSS INCLUDES ADVERSE TOLERANCE)	(GRD)	PAR(48)	-1.0 DB	TOL(48)	-.5 DB
10149	RECEIVED DATA SUBCARRIER POWER-TO-THERMAL NOISE SPECTRAL DENSITY (46+48)	(GRD)	PAR(49)	77.0 DB	TOL(49)	
10150	RECEIVER OUTPUT RFI SPREAD INTERFERENCE SPECTRAL DENSITY (39-42-35)	(GRD)	PAR(50)	-357.5 DBW/HZ	TOL(50)	
10151	TOTAL UNCORRELATED SPECTRAL DENSITY (45+50) (NOTE.. DBW IS NOT ADDED DIRECTLY)	(GRD)	PAR(51)	-191.1 DBW/HZ	TOL(51)	

I.D. CODE	PARAMETER		PARAMETER NUMBER	PARAMETER VALUE	TOLERANCE NUMBER	TOLERANCE VALUE
10152	RECEIVED DATA SUBCARRIER POWER-TO-COHERENT RFI POWER (36+48-39)	(GRD)	PAR(52)	NA	TOL(52)	
10153	RECEIVER DATA SUBCARRIER POWER-TO-TOTAL UNCORRELATED NOISE SPECTRAL DENSITY (36+48-51)	(GRD)	PAR(53)	77.0 DB	TOL(53)	
10154	CHANNEL BIT RATE (2.0M BDS)	(GRD)	PAR(54)	63.0 DB/BPS	TOL(54)	
10155	AVAILABLE ENERGY-TO-THERMAL NOISE SPECTRAL DENSITY (49-54)	(GRD)	PAR(55)	14.0 DB	TOL(55)	
10156	AVAILABLE ENERGY-TO-RFI POWER (COHERENT) (52-54)	(GRD)	PAR(56)	NA	TOL(56)	
10157	AVAILABLE ENERGY-TO-TOTAL UNCORRELATED NOISE SPECTRAL DENSITY (53-54)	(GRD)	PAR(57)	14.0 DB	TOL(57)	
10158	REQUIRED ENERGY-TO-NOISE SPECTRAL DENSITY (FOR BIT PE=1*10**-3)	(GRD)	PAR(58)	6.8 DB	TOL(58)	
10159	CHANNEL PERFORMANCE MARGIN (ABOVE THE SUM OF THE ADVERSE TOLERANCES)(57-58	(GRD)	PAR(59)	7.2 DB	TOL(59)	

PERFORMANCE MARGIN FOR NCS= 1 NSS= 1 IS ADEQUATE

KEY PARAMETER VALUES
FOR THE USER TELEMETRY CHANNEL

CASE	SET	PARAMETER(13)	PARAMETER(16)	PARAMETER(35)
2	1	-300.00	80.00	14.00

CASE 2. POWER BUDGET CALCULATIONS FOR RETURN LINK PERFORMANCE USER---TDRS---GROUND

SET 1. CONDITIONS RFI,NSS,NPG

HIGH DATA RATE(HDR) USER TELEMETRY

I.D. CODE	PARAMETER		PARAMETER NUMBER	PARAMETER VALUE	TOLERANCE NUMBER	TOLERANCE VALUE
	*** USERLK ***					
20101	TOTAL TRANSMITTED POWER (10.0W)	(USER)	PAR(1)	10.0 DBW	TOL(1)	
20102	TRANSMITTING CIRCUIT LOSS	(USER)	PAR(2)	-1.0 DB	TOL(2)	0.0 DB
20103	TRANSMITTING ANTENNA GAIN (3-FOOT DISH)	(USER)	PAR(3)	37.9 DB	TOL(3)	.6 DB
20104	TRANSMITTING ANTENNA POINTING LOSS (1.0 DEG)	(USER)	PAR(4)	-2.5 DB	TOL(4)	.1 DB
20105	SPACE LOSS (F=10.357 MHZ, D=35 800 KM)	(RF LINK)	PAR(5)	-203.9 DB	TOL(5)	2.7 DB
20106	POLARZATION LOSS	(RF LINK)	PAR(6)	-0.0 DB	TOL(6)	.1 DB
20107	RECEIVING ANTENNA GAIN (8-FT DISH)	(TDRS)	PAR(7)	46.5 DB	TOL(7)	.6 DB
20108	RECEIVING ANTENNA POINTING LOSS (1.3 DEG)	(TDRS)	PAR(8)	-1.7 DB	TOL(8)	.1 DB
20109	RECEIVING ANTENNA SCAN LOSS	(TDRS)	PAR(9)	-0.0 DB	TOL(9)	1.0 DB

I.D. CODE	PARAMETER		PARAMETER NUMBER	PARAMETER VALUE	TOLERANCE NUMBER	TOLERANCE VALUE
20110	RECEIVING CIRCUIT LOSS	(TDRS)	PAR(10)	-1.0 DB	TOL(10)	.1 DB
20111	SOME OF ADVERSE TOLERANCES (2 THRU 10)	(TDRS)	PAR(11)	-5.3 DB	TOL(11)	
20112	NET CIRCUIT LOSS (SUM 2 THRU 11)	(TDRS)	PAR(12)	-131.0 DB	TOL(12)	
20113	RECEIVED RFI POWER (COHERENT)	(TDRS)	PAR(13)	-300.0 DB	TOL(13)	
20114	RECEIVED SIGNAL POWER (1+12)	(TDRS)	PAR(14)	-121.0 DBW	TOL(14)	
	*** TDRSLK ***					
20115	RECEIVER THERMAL NOISE SPECTRAL DENSITY (TS = 900 DEG K,NF =6)	(TDRS)	PAR(15)	-199.1 DBW/HZ	TOL(15)	
20116	RECEIVER BANDWIDTH (4.0 MHZ)	(TDRS)	PAR(16)	80.0 DB/HZ	TOL(16)	
20117	RECEIVER NOISE POWER (15+16)	(TDRS)	PAR(17)	-119.1 DBW	TOL(17)	
20118	RECEIVED SIGNAL POWER-TO-THERMAL NOISE SPECTRAL DENSITY (14-15)	(TDRS)	PAR(18)	78.1 DB	TOL(18)	
20119	TDRS GAIN	(TDRS)	PAR(19)	120.8 DB	TOL(19)	
20120	TRANSMITTED SIGNAL POWER (1.0W) (14+19)	(TDRS)	PAR(20)	-.2 DBW	TOL(20)	

I.D. CODE	PARAMETER		PARAMETER NUMBER	PARAMETER VALUE	TOLERANCE NUMBER	TOLERANCE VALUE
20121	TDRS TRANSMITTED COHERENT INTERFERENCE (13+19)	(TDRS)	PAR(21)	-179.2 DBW	TOL(21)	
20122	TDRS TRANSMITTED THERMAL NOISE POWER (17+19)	(TDRS)	PAR(22)	1.7 DBW	TOL(22)	
20123	TRANSMITTING CIRCUIT LOSS	(TDRS)	PAR(23)	-.5 DB	TOL(23)	.1 DB
20124	TRANSMITTING ANTENNA GAIN (1-FT DISH)	(TDRS)	PAR(24)	26.4 DB	TOL(24)	.6 DB
20125	TRASMITTING ANTENNA POINTING LOSS	(TDRS)	PAR(25)	-0.0 DB	TOL(25)	.1 DB
20126	TRANSMITTING ANTENNA SCAN LOSS	(TDRS)	PAR(26)	-0.0 DB	TOL(26)	0.0 DB
20127	SPACE LOSS (F=8.250 MHZ, D=38,000 KM)	(RF LINK)	PAR(27)	-202.4 DB	TOL(27)	.2 DB
20128	POLARIZATION LOSS	(RF LINK)	PAR(28)	-0.0 DB	TOL(28)	.1 DB
20129	RECEIVING ANTENNA GAIN	(GRD)	PAR(29)	65.0 DB	TOL(29)	.2 DB
20130	RECEIVING ANTENNA POINTING LOSS	(GRD)	PAR(30)	-0.0 DB	TOL(30)	.1 DB
20131	RECEIVING ANTENNA SCAN LOSS	(GRD)	PAR(31)	-0.0 DB	TOL(31)	0.0 DB
20132	RECEIVING CIRCUIT LOSS	(GRD)	PAR(32)	-0.0 DB	TOL(32)	0.0 DB
20133	SOME OF ADVERSE TOLERANCES (23 THRU 32)	(GRD)	PAR(33)	-1.4 DB	TOL(33)	

I.D. CODE	PARAMETER		PARAMETER NUMBER	PARAMETER VALUE	TOLERANCE NUMBER	TOLERANCE VALUE
20134	NET CIRCUIT LOSS (SUM 23 THRU 33)	(GRD)	PAR(34)	-112.9 DB	TOL(34)	
20135	PROCESSING GAIN	(GRD)	PAR(35)	14.0 DB	TOL(35)	
20136	TOTAL RECEIVED SIGNAL POWER (2 +34)	(GRD)	PAR(36)	-113.1 DBW	TOL(36)	
20137	RECEIVED COHERENT INTERFERENCE FROM TDRS (21+34)	(GRD)	PAR(37)	-292.1 DBW	TOL(37)	
20138	RECEIVED COHERENT INTERFERENCE TDRS TO GND LINK	(GRD)	PAR(38)	-300.0 DBW	TOL(38)	
20139	TOTAL RFI POWER (COHERENT) (37+38) (NOTE.. DBW CAN NOT BE ADDED DIRECTLY)	(GRD)	PAR(39)	-291.5 DBW	TOL(39)	
20140	RECEIVED THERMAL NOISE POWER (22+34)	(GRD)	PAR(40)	-111.2 DBW	TOL(40)	
	*** GROLK ***					
20141	RECEIVER THERMAL NOISE SPECTRAL DENSITY (TS =100 DEG K), (NF=DB)	(GRD)	PAR(41)	-208.6 DBW/HZ	TOL(41)	
20142	RECEIVER BANDWIDTH (4.0 MHZ)	(GRD)	PAR(42)	66.0 DB/HZ	TOL(42)	

I.D. CODE	PARAMETER		PARAMETER NUMBER	PARAMETER VALUE	TOLERANCE NUMBER	TOLERANCE VALUE
20143	RECEIVER THERMAL NOISE POWER (41+42)	(GRD)	PAR(43)	-142.6 DBW	TOL(43)	
20144	TOTAL THERMAL NOISE POWER OUTPUT (42-35+43) (NOTE. DBW CAN NOT BE ADDED DIRECTLY)	(GRD)	PAR(44)	-125.1 DBW	TOL(44)	
20145	ACTUAL RECEIVER THERMAL NOISE SPECTRAL DENSITY (44-42)	(GRD)	PAR(45)	-191.1 DBW/HZ	TOL(45)	
20146	RECEIVED SIGNAL POWER-TO-THERMAL NOISE SPECTRAL DENSITY (36-45)	(GRD)	PAR(46)	78.0 DB	TOL(46)	
20147	TOTAL RECEIVED SIGNAL POWER-TO-TOTAL THERMAL NOISE POWER (34-44)	(GRD)	PAR(47)	12.0 DB	TOL(47)	
20148	MODULATION LOSS (MODULATION LOSS INCLUDES ADVERSE TOLERANCE)	(GRD)	PAR(48)	-1.0 DB	TOL(48)	.5 DB
20149	RECEIVED DATA SUBCARRIER POWER-TO-THERMAL NOISE SPECTRAL DENSITY (46+48)	(GRD)	PAR(49)	77.0 DB	TOL(49)	
20150	RECEIVER OUTPUT SFI SPREAD INTERFERENCE SPECTRAL DENSITY (39-42-35)	(GRD)	PAR(50)	-371.5 DBW/HZ	TOL(50)	
20151	TOTAL UNCORRELATED SPECTRAL DENSITY (45+50) (NOTE.. DBW IS NOT ADDED DIRECTLY)	(GRD)	PAR(51)	-191.1 DBW/HZ	TOL(51)	

I.D. CODE	PARAMETER		PARAMETER NUMBER	PARAMETER VALUE	TOLERANCE NUMBER	TOLERANCE VALUE
20152	RECEIVED DATA SUBCARRIER POWER-TO-COHERENT RFI POWER (36+48-39)	(GRD)	PAR(52)	NA	TOL(52)	
20153	RECEIVER DATA SUBCARRIER POWER-TO-TOTAL UNCORRELATED NOISE SPECTRAL DENSITY (36+48-51)	(GRD)	PAR(53)	77.0 DB	TOL(53)	
20154	CHANNEL BIT RATE (2.0M BPS)	(GRD)	PAR(54)	63.0 DB/BPS	TOL(54)	
20155	AVAILABLE ENERGY-TO-THERMAL NOISE SPECTRAL DENSITY (49-54)	(GRD)	PAR(55)	14.0 DB	TOL(55)	
20156	AVAILABLE ENERGY-TO-RFI POWER (COHERENT) (52-54)	(GRD)	PAR(56)	NA	TOL(56)	
20157	AVAILABLE ENERGY-TO-TOTAL UNCORRELATED NOISE SPECTRAL DENSITY (53-54)	(GRD)	PAR(57)	14.0 DB	TOL(57)	
20158	REQUIRED ENERGY-TO-NOISE SPECTRAL DENSITY (FOR BIT PE=1*10**-3)	(GRD)	PAR(58)	6.8 DB	TOL(58)	
20159	CHANNEL PERFORMANCE MARGIN (ABOVE THE SUM OF THE ADVERSE TOLERANCES) (57-58	(GRD)	PAR(59)	7.2 DB	TOL(59)	

PERFORMANCE MARGIN FOR NCS= 2 NSS= 1 IS ADEQUATE

KEY PARAMETER VALUES
FOR THE USER TELEMETRY CHANNEL

CASE	SET	PARAMETER(13)	PARAMETER(16)	PARAMETER(35)
3	1	-125.80	66.00	0.00

CASE 3. POWER BUDGET CALCULATIONS FOR RETURN LINK PERFORMANCE USER---TDRS---GROUND

SET 1. CONDITIONS NRFI,NSS,NPG

HIGH DATA RATE(HDR) USER TELEMETRY

I.D. CODE	PARAMETER		PARAMETER NUMBER	PARAMETER VALUE	TOLERANCE NUMBER	TOLERANCE VALUE
	*** USERLK ***					
30101	TOTAL TRANSMITTER POWER (10.0W)	(USER)	PAR(1)	10.0 DBW	TOL(1)	
30102	TRANSMITTING CIRCUIT LOSS	(USER)	PAR(2)	-1.0 DB	TOL(2)	0.0 DB
30103	TRANSMITTING ANTENNA GAIN (3-FOOT DISH)	(USER)	PAR(3)	37.9 DB	TOL(3)	.6 DB
30104	TRANSMITTING ANTENNA POINTING LOSS (1.0 DEG)	(USER)	PAR(4)	-2.5 DB	TOL(4)	.1 DB
30105	SPACE LOSS (F=10.357 MHZ, D=35 800 KM)	(RF LINK)	PAR(5)	-203.9 DB	TOL(5)	2.7 DB
30106	POLARZATION LOSS	(RF LINK)	PAR(6)	-0.0 DB	TOL(6)	.1 DB
30107	RECEIVING ANTENNA GAIN (8-FT DISH)	(TDRS)	PAR(7)	46.5 DB	TOL(7)	.6 DB
30108	RECEIVING ANTENNA POINTING LOSS (0.3 DEG)	(TDRS)	PAR(8)	-1.7 DB	TOL(8)	.1 DB
30109	RECEIVING ANTENNA SCAN LOSS	(TDRS)	PAR(9)	-0.0 DB	TOL(9)	1.0 DB

I.D. CODE	PARAMETER		PARAMETER NUMBER	PARAMETER VALUE	TOLERANCE NUMBER	TOLERANCE VALUE
30110	RECEIVING CIRCUIT LOSS	(TDRS)	PAR(10)	-1.0 DB	TOL(10)	.1 DB
30111	SOME OF ADVERSE TOLERANCES (2 THRU 10)	(TDRS)	PAR(11)	-5.3 DB	TOL(11)	
30112	NET CIRCUIT LOSS (SUM 2 THRU 11)	(TDRS)	PAR(12)	-131.0 DB	TOL(12)	
30113	RECEIVED RFI POWER (COHERENT)	(TDRS)	PAR(13)	-125.8 DB	TOL(13)	
30114	RECEIVED SIGNAL POWER (1+12)	(TDRS)	PAR(14)	-121.0 DBW	TOL(14)	
	*** TDRSLK ***					
30115	RECEIVER THERMAL NOISE SPECTRAL DENSITY (TS = 900 DEG K, NF =6)	(TDRS)	PAR(15)	-199.1 DBW/HZ	TOL(15)	
30116	RECEIVER BANDWIDTH (4.0 MHZ)	(TDRS)	PAR(16)	66.0 DB/HZ	TOL(16)	
30117	RECEIVER NOISE POWER (15+16)	(TDRS)	PAR(17)	-133.1 DBW	TOL(17)	
30118	RECEIVED SIGNAL POWER-TO-THERMAL NOISE SPECTRAL DENSITY (14-15)	(TDRS)	PAR(18)	78.1 DB	TOL(18)	
30119	TDRS GAIN	(TDRS)	PAR(19)	120.8 DB	TOL(19)	
30120	TRANSMITTED SIGNAL POWER (1.0H) (14+19)	(TDRS)	PAR(20)	-.2 DBW	TOL(20)	

I.D. CODE	PARAMETER		PARAMETER NUMBER	PARAMETER VALUE	TOLERANCE NUMBER	TOLERANCE VALUE
30121	TDRS TRANSMITTED COHERENT INTERFERENCE (13+19)	(TDRS)	PAR(21)	-5.0 DBW	TOL(21)	
30122	TDRS TRANSMITTED THERMAL NOISE POWER (17+19)	(TDRS)	PAR(22)	-12.3 DBW	TOL(22)	
30123	TRANSMITTING CIRCUIT LOSS	(TDRS)	PAR(23)	-.5 DB	TOL(23)	.1 DB
30124	TRANSMITTING ANTENNA GAIN (1-FT DISH)	(TDRS)	PAR(24)	26.4 DB	TOL(24)	.6 DB
30125	TRASMITTING ANTENNA POINTING LOSS	(TDRS)	PAR(25)	-0.0 DB	TOL(25)	.1 DB
30126	TRANSMITTING ANTENNA SCAN LOSS	(TDRS)	PAR(26)	-0.0 DB	TOL(26)	0.0 DB
30127	SPACE LOSS (F=8,250 MHZ, D=38,000 KM)	(RF LINK)	PAR(27)	-202.4 DB	TOL(27)	.2 DB
30128	POLARIZATION LOSS	(RF LINK)	PAR(28)	-0.0 DB	TOL(28)	.1 DB
30129	RECEIVING ANTENNA GAIN	(GRD)	PAR(29)	65.0 DB	TOL(29)	.2 DB
30130	RECEIVING ANTENNA POINTING LOSS	(GRD)	PAR(30)	-0.0 DB	TOL(30)	.1 DB
30131	RECEIVING ANTENNA SCAN LOSS	(GRD)	PAR(31)	-0.0 DB	TOL(31)	0.0 DB
30132	RECEIVING CIRCUIT LOSS	(GRD)	PAR(32)	-0.0 DB	TOL(32)	0.0 DB
30133	SOME OF ADVERSE TOLERANCES (23 THRU 32)	(GRD)	PAR(33)	-1.4 DB	TOL(33)	

I.D. CODE	PARAMETER		PARAMETER NUMBER	PARAMETER VALUE	TOLERANCE NUMBER	TOLERANCE VALUE
30134	NET CIRCUIT LOSS (SUM 23 THRU 33)	(GRD)	PAR(34)	-112.9 DB	TOL(34)	
30135	PROCESSING GAIN	(GRD)	PAR(35)	0.0 DB	TOL(35)	
30136	TOTAL RECEIVED SIGNAL POWER (20+34)	(GRD)	PAR(36)	-113.1 DBW	TOL(36)	
30137	RECEIVED COHERENT INTERFERENCE FROM TDRS (21+34)	(GRD)	PAR(37)	-117.9 DBW	TOL(37)	
30138	RECEIVED COHERENT INTERFERENCE TDRS TO GND LINK	(GRD)	PAR(38)	-300.0 DBW	TOL(38)	
30139	TOTAL RFI POWER (COHERENT) (37+38) (NOTE.. DBW CAN NOT BE ADDED DIRECTLY)	(GRD)	PAR(39)	-117.9 DBW	TOL(39)	
30140	RECEIVED THERMAL NOISE POWER (22+34)	(GRD)	PAR(40)	-125.2 DBW	TOL(40)	
	*** GRDLK ***					
30141	RECEIVER THERMAL NOISE SPECTRAL DENSITY (TS =100 DEG K), (NF=DB)	(GRD)	PAR(41)	-208.6 DBW/HZ	TOL(41)	
30142	RECEIVER BANDWIDTH (4.0 MHZ)	(GRD)	PAR(42)	66.0 DB/HZ	TOL(42)	

153A

I.D. CODE	PARAMETER		PARAMETER NUMBER	PARAMETER VALUE	TOLERANCE NUMBER	TOLERANCE VALUE
30143	RECEIVER THERMAL NOISE POWER (41+42)	(GRD)	PAR(43)	-142.6 DBW	TOL(43)	
30144	TOTAL THERMAL NOISE POWER OUTPUT (4?-35+43)(NOTE.,DBW CAN NOT BE ADDED DIRECTLY)	(GRD)	PAR(44)	-125.1 DBW	TOL(44)	
30145	ACTUAL RECEIVER THERMAL NOISE SPECTRAL DENSITY (44-42)	(GRD)	PAR(45)	-191.1 DBW/HZ	TOL(45)	
30146	RECEIVED SIGNAL POWER-TO-THERMAL NOISE SPECTRAL DENSITY (36-45)	(GRD)	PAR(46)	78.0 DB	TOL(46)	
30147	TOTAL RECEIVED SIGNAL POWER-TO-TOTAL THERMAL NOISE POWER (34-44)	(GRD)	PAR(47)	12.0 DB	TOL(47)	
30148	MODULATION LOSS (MODULATION LOSS INCLUDES ADVERSE TOLERANCE)	(GRD)	PAR(48)	-1.0 DB	TOL(48)	+5 DB
30149	RECEIVED DATA SUBCARRIER POWER-TO-THERMAL NOISE SPECTRAL DENSITY (46+48)	(GRD)	PAR(49)	77.0 DB	TOL(49)	
30150	RECEIVER OUTPUT RFI SPREAD INTERFERENCE SPECTRAL DENSITY (39-42-35)	(GRD)	PAR(50)	NA	TOL(50)	
30151	TOTAL UNCORRELATED SPECTRAL DENSITY (45+50) (NOTE.. DBW IS NOT ADDED DIRECTLY)	(GRD)	PAR(51)	NA	TOL(51)	

ITEM 51 REPRESENTS THE SUM OF THERMAL NOISE SPECTRAL DENSITY AND SPREAD COHERENT INTERFERENCE SPECTRAL DENSITY. FOR THE REASONS GIVEN IN APPENDIX C (NOTE 4) ITEM 51 IS NOT APPLICABLE. AS A RESULT, ITEMS 53, 57, AND 59 ARE ALSO NOT APPLICABLE. THIS DOES NOT PRECLUDE A CASE-BY-CASE ANALYSIS TO TAKE INTO ACCOUNT THE EFFECT OF RFI.

I.D. CODE	PARAMETER		PARAMETER NUMBER	PARAMETER VALUE	TOLERANCE NUMBER	TOLERANCE VALUE
30152	RECEIVED DATA SUBCARRIER POWER-TO-COHERENT RFI POWER (36+48-39)	(GRD)	PAR(52)	3.8 DB	TOL(52)	
30153	RECEIVER DATA SUBCARRIER POWER-TO-TOTAL UNCORRELATED NOISE SPECTRAL DENSITY (36+48-51)	(GRD)	PAR(53)	NA	TOL(53)	
30154	CHANNEL BIT RATE (2.0M BPS)	(GRD)	PAR(54)	63.0 DB/BPS	TOL(54)	
30155	AVAILABLE ENERGY-TO-THERMAL NOISE SPECTRAL DENSITY (49-54)	(GRD)	PAR(55)	14.0 DB	TOL(55)	
30156	AVAILABLE ENERGY-TO-RFI POWER (COHERENT) (52-54)	(GRD)	PAR(56)	-59.2 DB	TOL(56)	
30157	AVAILABLE ENERGY-TO-TOTAL UNCORRELATED NOISE SPECTRAL DENSITY (53-54)	(GRD)	PAR(57)	NA	TOL(57)	
30158	REQUIRED ENERGY-TO-NOISE SPECTRAL DENSITY (FOR BIT PE=1*10**-3)	(GRD)	PAR(58)	6.8 DB	TOL(58)	
30159	CHANNEL PERFORMANCE MARGIN (ABOVE THE SUM OF THE ADVERSE TOLERANCES)(57-58	(GRD)	PAR(59)	NA	TOL(59)	

PERFORMANCE MARGIN FOR NCS= 3 NSS= 1 IS NOT APPLICABLE

KEY PARAMETER VALUES
FOR THE USER TELEMETRY CHANNEL

CASE	SET	PARAMETER(13)	PARAMETER(16)	PARAMETER(35)
4	1	-125.80	80.00	14.00

CASE 4. POWER BUDGET CALCULATIONS FOR RETURN LINK PERFORMANCE USER---TDRS---GROUND

SET 1. CONDITIONS NRFI,NSS,NPG

HIGH DATA RATE(HDR) USER TELEMETRY

I.D. CODE	PARAMETER		PARAMETER NUMBER	PARAMETER VALUE	TOLERANCE NUMBER	TOLERANCE VALUE
	*** USERLK **					
40101	TOTAL TRANSMITTED POWER (10.0W)	(USER)	PAR(1)	10.0 DBW	TOL(1)	
40102	TRANSMITTING CIRCUIT LOSS	(USER)	PAR(2)	-1.0 DB	TOL(2)	0.0 DB
40103	TRANSMITTING ANTENNA GAIN (3-FOOT DISH)	(USER)	PAR(3)	37.9 DB	TOL(3)	.6 DB
40104	TRANSMITTING ANTENNA POINTING LOSS (1.0 DEG)	(USER)	PAR(4)	-2.5 DB	TOL(4)	.1 DB
40105	SPACE LOSS (F=15,357 MHZ, D=35 800 KM)	(RF LINK)	PAR(5)	-203.9 DB	TOL(5)	2.7 DB
40106	POLARZATION LOSS	(RF LINK)	PAR(6)	-0.0 DB	TOL(6)	.1 DB
40107	RECEIVING ANTENNA GAIN (8-FT DISH)	(TDRS)	PAR(7)	46.5 DB	TOL(7)	.6 DB
40108	RECEIVING ANTENNA POINTING LOSS (0.3 DEG)	(TDRS)	PAR(8)	-1.7 DB	TOL(8)	.1 DB
40109	RECEIVING ANTENNA SCAN LOSS	(TDRS)	PAR(9)	-0.0 DB	TOL(9)	1.0 DB

I.D. CODE	PARAMETER		PARAMETER NUMBER	PARAMETER VALUE	TOLERANCE NUMBER	TOLERANCE VALUE
40110	RECEIVING CIRCUIT LOSS	(TDRS)	PAR(10)	-1.0 DB	TOL(10)	.1 DB
40111	SOME OF ADVERSE TOLERANCES (2 THRU 10)	(TDRS)	PAR(11)	-5.3 DB	TOL(11)	
40112	NET CIRCUIT LOSS (SUM 2 THRU 11)	(TDRS)	PAR(12)	-131.0 DB	TOL(12)	
40113	RECEIVED RFI POWER (COHERENT)	(TDRS)	PAR(13)	-125.8 DB	TOL(13)	
40114	RECEIVED SIGNAL POWER (1+12)	(TDRS)	PAR(14)	-121.0 DBW	TOL(14)	
	*** TDRSLK ***					
40115	RECEIVER THERMAL NOISE SPECTRAL DENSITY (TS = 900 DEG K,NF =6)	(TDRS)	PAR(15)	-199.1 DBW/HZ	TOL(15)	
40116	RECEIVER BANDWIDTH (4.0 MHZ)	(TDRS)	PAR(16)	80.0 DB/HZ	TOL(16)	
40117	RECEIVER NOISE POWER (15+16)	(TDRS)	PAR(17)	-119.1 DBW	TOL(17)	
40118	RECEIVED SIGNAL POWER-TO-THERMAL NOISE SPECTRAL DENSITY (14-15)	(TDRS)	PAR(18)	78.1 DB	TOL(18)	
40119	TDRS GAIN	(TDRS)	PAR(19)	120.8 DB	TOL(19)	
40120	TRANSMITTED SIGNAL POWER (1.0?) (14,19)	(TDRS)	PAR(20)	-.2 DBW	TOL(20)	

157

I.D. CODE	PARAMETER		PARAMETER NUMBER	PARAMETER VALUE	TOLERANCE NUMBER	TOLERANCE VALUE
40121	TDRS TRANSMITTED COHERENT INTERFERENCE (13+19)	(TDRS)	PAR(21)	-5.0 DBW	TOL(21)	
40122	TDRS TRANSMITTED THERMAL NOISE POWER (17+19)	(TDRS)	PAR(22)	1.7 DBW	TOL(22)	
40123	TRANSMITTING CIRCUIT LOSS	(TDRS)	PAR(23)	-.5 DB	TOL(23)	.1 DB
40124	TRANSMITTING ANTENNA GAIN (?-FT DISH)	(TDRS)	PAR(24)	26.4 DB	TOL(24)	.6 DB
40125	TRASMITTING ANTENNA POINTING LOSS	(TDRS)	PAR(25)	-0.0 DB	TOL(25)	.1 DB
40126	TRANSMITTING ANTENNA SCAN LOSS	(TDRS)	PAR(26)	-0.0 DB	TOL(26)	0.0 DB
40127	SPACE LOSS (F=8,250 MHZ, D=38,000 KM)	(RF LINK)	PAR(27)	-202.4 DB	TOL(27)	.2 DB
40128	POLARIZATION LOSS	(RF LINK)	PAR(28)	-0.0 DB	TOL(28)	.1 DB
40129	RECEIVING ANTENNA GAIN	(GRD)	PAR(29)	65.0 DB	TOL(29)	.2 DB
40130	RECEIVING ANTENNA POINTING LOSS	(GRD)	PAR(30)	-0.0 DB	TOL(30)	.1 DB
40131	RECEIVING ANTENNA SCAN LOSS	(GRD)	PAR(31)	-0.0 DB	TOL(31)	0.0 DB
40132	RECEIVING CIRCUIT LOSS	(GRD)	PAR(32)	-0.0 DB	TOL(32)	0.0 DB
40133	SUM OF ADVERSE TOLERANCES (23 THRU 32)	(GRD)	PAR(33)	-1.4 DB	TOL(33)	

I.D. CODE	PARAMETER		PARAMETER NUMBER	PARAMETER VALUE	TOLERANCE NUMBER	TOLERANCE VALUE
40134	NET CIRCUIT LOSS (SUM 23 THRU 33)	(GRD)	PAR(34)	-112.9 DB	TOL(34)	
40135	PROCESSING GAIN	(GRD)	PAR(35)	14.0 DB	TOL(35)	
40136	TOTAL RECEIVED SIGNAL POWER (2?+34)	(GRD)	PAR(36)	-113.1 DBW	TOL(36)	
40137	RECEIVED COHERENT INTERFERENCE FROM TDRS (21+34)	(GRD)	PAR(37)	-117.9 DBW	TOL(37)	
40138	RECEIVED COHERENT INTERFERENCE TDRS TO GND LINK	(GRD)	PAR(38)	-300.0 DBW	TOL(38)	
40139	TOTAL RFI POWER (COHERENT) (37+38) (NOTE.. DBW CAN NOT BE ADDED DIRECTLY)	(GRD)	PAR(39)	-117.9 DBW	TOL(39)	
40140	RECEIVED THERMAL NOISE POWER (22+34)	(GRD)	PAR(40)	-111.2 DBW	TOL(40)	
	*** GHDLK ***					
40141	RECEIVER THERMAL NOISE SPECTRAL DENSITY (TS =100 DEG K), (NF=DB)	(GRD)	PAR(41)	-208.6 DBW/HZ	TOL(41)	
40142	RECEIVER BANDWIDTH (4.0 MHZ)	(GRD)	PAR(42)	66.0 DB/HZ	TOL(42)	

I.D. CODE	PARAMETER		PARAMETER NUMBER	PARAMETER VALUE	TOLERANCE NUMBER	TOLERANCE VALUE
40143	RECEIVER THERMAL NOISE POWER (41+42)	(GRD)	PAR(43)	-142.6 DBW	TOL(43)	
40144	TOTAL THERMAL NOISE POWER OUTPUT (45-35+43)(NOTE..DBW CAN NOT BE ADDED DIRECTLY)	(GRD)	PAR(44)	-125.1 DBW	TOL(44)	
40145	ACTUAL RECEIVER THERMAL NOISE SPECTRAL DENSITY (44-42)	(GRD)	PAR(45)	-191.1 DBW/HZ	TOL(45)	
40146	RECEIVED SIGNAL POWER-TO-THERMAL NOISE SPECTRAL DENSITY (36-45)	(GRD)	PAR(46)	78.0 DB	TOL(46)	
40147	TOTAL RECEIVED SIGNAL POWER-TO-TOTAL THERMAL NOISE POWER (36-44)	(GRD)	PAR(47)	12.0 DB	TOL(47)	
40148	MODULATION LOSS (MODULATION LOSS INCLUDES ADVERSE TOLERANCE)	(GRD)	PAR(48)	-1.0 DB	TOL(48)	.5 DB
40149	RECEIVED DATA SUBCARRIER POWER-TO-THERMAL NOISE SPECTRAL DENSITY (46+48)	(GRD)	PAR(49)	77.0 DB	TOL(49)	
40150	RECEIVER OUTPUT RFI SPREAD INTERFERENCE SPECTRAL DENSITY (39-42-35)	(GRD)	PAR(50)	-197.9 DBW/HZ	TOL(50)	
40151	TOTAL UNCORRELATED SPECTRAL DENSITY (45+50) (NOTE.. DBW IS NOT ADDED DIRECTLY)	(GRD)	PAR(51)	-190.3 DBW/HZ	TOL(51)	

I.D. CODE	PARAMETER		PARAMETER NUMBER	PARAMETER VALUE	TOLERANCE NUMBER	TOLERANCE VALUE
40152	RECEIVED DATA SUBCARRIER POWER-TO-COHERENT RFI POWER (36+48-49)	(GRD)	PAR(52)	NA	TOL(52)	
40153	RECEIVER DATA SUBCARRIER POWER-TO-TOTAL UNCORRELATED NOISE SPECTRAL DENSITY (36+48-51)	(GRD)	PAR(53)	76.2 DB	TOL(53)	
40154	CHANNEL BIT RATE (2.0M BPS)	(GRD)	PAR(54)	63.0 DB/BPS	TOL(54)	
40155	AVAILABLE ENERGY-TO-THERMAL NOISE SPECTRAL DENSITY (49-54)	(GRD)	PAR(55)	14.0 DB	TOL(55)	
40156	AVAILABLE ENERGY-TO-RFI POWER (COHERENT) (52-54)	(GRD)	PAR(56)	NA	TOL(56)	
40157	AVAILABLE ENERGY-TO-TOTAL UNCORRELATED NOISE SPECTRAL DENSITY (53-54)	(GRD)	PAR(57)	13.2 DB	TOL(57)	
40158	REQUIRED ENERGY-TO-NOISE SPECTRAL DENSITY (FOR BIT PE=1*10**-3)	(GRD)	PAR(58)	6.8 DB	TOL(58)	
40159	CHANNEL PERFORMANCE MARGIN (ABOVE THE SUM OF THE ADVERSE TOLERANCES)(57-58	(GRD)	PAR(59)	6.4 DB	TOL(59)	

PERFORMANCE MARGIN FOR NCS= 4 NSS= 1 IS ADEQUATE

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 03/27/74  SAFSCOM COMP CNTR  PSR L357  09/01/73
15.22.59.VP540AC  FROM *.
15.23.01.IP  00002624 WORDS - FILE INPUT  , DC 00
15.23.01.VP500.
15.23.01.          WEALAW                   12      $S
15.23.01.BOXC(HOTSEAT,JEFFREYS)
15.23.01.)*****************************************
15.23.01. *                                       *
15.23.01. *            H O T S E A T              *
15.23.01. *           J E F F R E Y S             *
15.23.01. *                                       *
15.23.01. ****************************************
15.23.01.FTN.
15.23.33.     4.283 CP SECONDS COMPILATION TIME
15.23.33.LGC.
15.23.47.    13.615 RT SECONDS LOAD TIME
15.23.59.    STOP
15.23.59.     4.357 CP SECONDS EXECUTION TIME
15.23.59.OP  00010112 WORDS - FILE OUTPUT , DC 40
15.23.59.CPA      11.037 SEC.       11.037 ADJ.
15.23.59.IO        2.395 SEC.        2.395 ADJ.
15.23.59.CM      211.600 KWS.       12.877 ADJ.
15.23.59.SS                         26.310
15.23.59.PP       18.606 SEC.     DATE 03/27/74
15.23.59.EJ  END OF JOB, **
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