CR-171 861
c.1

FINAL REPORT

SHUTTLE COMMUNICATIONS AND TRACKING SYSTEMS MODELING AND TDRSS LINK SIMULATION STUDIES

(NASA-CR-171861) SHUTTLE COMMUNICATIONS AND TRACKING SYSTEMS MODELING AND TDRSS LINK SIMULATIONS STUDIES Final Report (LinCom Corp., Los Angeles, Calif.) 313 p HC A14/MF A01 CSCL 22B G3/16

N85-25369

Unclas 21197

PREPARED FOR

NASA/LYNDON B. JOHNSON SPACE CENTER

HOUSTON, TX 77058

TECHNICAL MONITOR: WILLIAM E. TEASDALE

CONTRACT NO. NAS 9-17152

MAY, 1985

LinCom Corporation

P.O BOX 15897, LOS ANGELES, CALIFORNIA 90015

TR-8411-1

LinCom

FINAL REPORT

SHUTTLE COMMUNICATIONS AND TRACKING SYSTEMS MODELING AND TDRSS LINK SIMULATIONS STUDIES

PREPARED FOR

NASA/LYNDON B. JOHNSON SPACE CENTER
HOUSTON, TX 77058

TECHNICAL MONITOR: WILLIAM E. TEASDALE

CONTRACT NO. NAS 9-17152

PREPARED BY

C. M. CHIE
K. DESSOUKY
W. C. LINDSEY
C. S. TSANG
Y. T. SU

LINCOM CORPORATION
P.O. BOX 15897
LOS ANGELES, CA 90015-0897

MAY, 1985

TR-8411-1

LinCom

TABLE OF CONTENTS

1.0 SUMMARY

1.1 Introduction

This document represents the final report for the "Space Shuttle Program Communication and Tracking Systems Modeling and TDRSS Link Simulations," performed under Contract No. NAS9-17152. It represents a portion of the work accomplished during the reporting period June 1, 1984 through May 1,1985.

Since August 1976 LinCom has been under contract to NASA/JSC to modify and extend an analytical simulation package (LinCsim) which allows the analytical verification of data transmission performance through TDRSS satellites [1,2,3]. The work involved the modeling of the user transponder, TDRS, TDRS ground terminal, and link dynamics for forward and return links based on the TDRSS performance specifications [4] and the critical design reviews conducted by WU/TRW/HESD contractors [5,6]. The scope of this effort has recently been expanded to include the effects of radio frequency interference (RFI) on the bit error rate (BER) performance of the S-band return links. The RFI environment and the modified TDRSS satellite and ground station hardware are being modeled in accordance with their description in the applicable documents [7].

Since 1976 LinCom has been expanding this program capability to make it applicable to Shuttle unique/ground station capabilities. Now that both the TDRSS and Shuttle programs have completed the conceptual design phases and most of the design/development activities (especially the TDRSS), some specific interface problems have been identified which affect Shuttle capabilities. Of major importance is the RFI determined to be present on the S-band return link. In addition, a number of areas (user constraints) of incompatibility between TDRSS and Shuttle orbiter exist which need to be resolved through analysis and simulation. In order to evaluate the technical and operational problem areas and provide a recommendation, the NASA/JSC Statement of Work (SOW) provided for a study to evaluate the performance of the Orbiter Ku-band and S-band TDRSS Communications links through simulation and analysis. The RFI models, as determined and supplied by Goddard Space Flight Center, has been updated into LinCsim as the RFI environment became better known. Additionally based on the above tasks, LinCom provided recommendations for system and subsystem level tests and analysis.

Under a separate continuing contract to NASA/GSFC [3], LinCom was developing an extensive computer simulation model for analytical verification of bit error performance and tracking service performances for user satellite MA, SSA and KSA services through the Tracking and Data Relay Satellite (TDRS). The above mentioned simulation was used and modified to incorporate detailed simulation of the Shuttle Orbiter/TDRSS Ku-band and RFI/S-band link. Using this new link simulation LinCom evaluated the performance of the Orbiter Ku-band/S-band link under varying system RFI parameter conditions. The results of these simulations served as a data base for support of Orbiter Ku-band/S-band system design reviews. In addition, LinCom made recommendations for Ku-band and S-band system and subsystem level tests to support ESTL RFI system performance verification and evaluation tests and analyzed the test results required.

2.0 SUMMARY OF WORK ACCOMPLISHED

This final report documents the analysis, modeling, simulation and evaluation performed by LinCom under three tasks contained in the

Statement of Work. During this contract period, LinCom has continued evaluation/simulation of the Shuttle/TDRSS S-band and Ku-band forward and return link performance using updated Shuttle and TDRSS performance parameters/characteristics. Effects of TDRSS failures/operational constraints on Shuttle link performance has been evaluated. Problem areas encountered during missions have been analyzed and potential viable approaches to resolving problems have been identified.

LinCom continued to develop, expand and refine simulation models to emulate the effects of RFI environments on digital communications links. GSFC RFI model updates were incorporated into the simulation capability to make it available to unique RF links.

LinCom continued refinement of analytical simulations models for Shuttle/TDRSS links to reflect updated TDRS performance characteristics/operational constraints. LinCom continued development of generalized C&T system mathematical models and integrated software modules for the NASA/JSC vax computer system, allowing simulation and evaluation of performance of one-way and two-way communication and tracking links. Development was done using a modular approach that facilitates integration of future simulation requirements.

In addition to the above mentioned efforts, LinCom personnel was directed by the NASA/JSC technical monitor, Mr. William Teasdale, on various supporting studies, analyses, model development and link simulations to ensure that elements of the Shuttle C&T Systems meet Space Shuttle Program (SSP) requirements. The results of these efforts are documented in Appendices A-N.

REFERENCES

[1] "Tracking and Data Relay Satellite System (TDRSS) Users' Guide, Revision 4," STDN No. 101.2, NASA Goddard Space Flight Center, January 1980.

[2] Chie, C. M., "A Modified Barrett-Lampard Expansion and Its Application to Bandpass Nonlinearities with Both AM-AM and AM-PM Conversion," IEEE Trans. on Comm., Vol. COM-28, November 1980.

[3] Gagliardi, R. M., et al., "Phase Noise Analysis of the TDRSS," Proc. NTC '77, Los Angeles, CA, 1977.

[4] Braun, W. R., "Comparison Between Variable and Fixed Dwell-Time PN Acquisition Algorithms," Proc. ICC 1981, Denver, CO, 1981.

[5] Braun, W. R., Performance Analysis for Variable Dwell Time Expanding Serach PN Acquisition Algorithm," IEEE Trans. on Comm., Vol. COM-30, March 1982.

[6] Davisson, L. D., and Milstein, L. B., "On the Performance of Digital Communications Systems with Bandpass Limiters--Part I," IEEE Trans. on Comm., Vol. COM-20, October 1972.

[7] Massey, J. L., "Coding and Modulation in Digital Communications," Proceedings, Int'l. Zurich Seminar on Digital Communications, Switzerland, March 12-15, 1974.

LinCom

APPENDIX A

KU-BAND SHUTTLE MRD I-Q CHANNEL

REVERSAL STUDY

LinCom

TABLE OF CONTENTS

LIST OF FIGURES

LIST OF TABLES

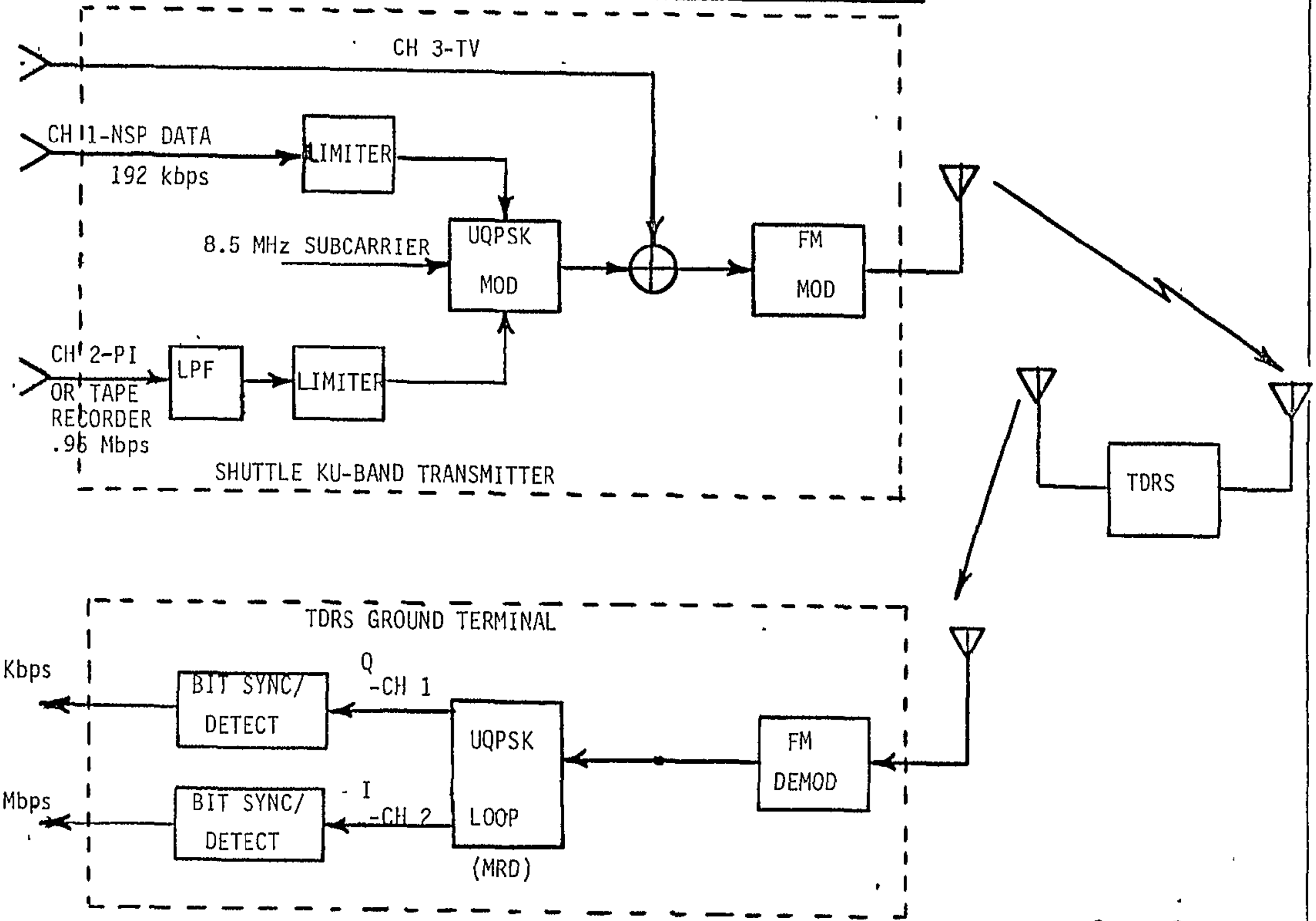
1. Introduction

There are three channels (1, 2, and 3) in the Space Shuttle Ku-Band return link subsystem. Channel 1 (192 Kbps) is primarily for voice and telemetry data. Channel 2 (maximum 2 Mbps) is designed for tape recorder playback or Payload Interrogator bent-pipe mode transmission. Channel 3 allows for Mode 1 a maximum data rate of 50 Mbps, or analog signals (TV) for Mode 2. Channels 1 and 2 are QPSK modulating a 8.5 MHz subcarrier which in turn along with channel 3 Quadrature Double Sideband (Mode 1) or Frequency (Mode 2) modulating a carrier. The composite signal is sent back to the White Sands Ground Terminal (WSGT) via the TDRS.

The scenario addressed here is illustrated in Figure 1. The channel 2 is either a tape recorder playback of 0.96 Mbps Biphase (Bi-ϕ) signal or a bent-pipe signal. The Channel 1 contains a 192 kbps Biphase signal. These two signals are unbalanced QPSK (UQPSK) modulated into a 8.5 MHz subcarrier. The power ratio between the two signals is 4:1 with the Channel 2 being a stronger channel. In the WSGT receiver, the subcarrier is extracted from the FM signal and input it to the medium rate demodualtor (MRD).

The MRD demodulates the subcarrier into I (Channel 2) and Q (Channel 1) channels. Each channel contains a bit synchronizer and detector of the appropriate data rate. Under certain conditions (to be discussed below) the I & Q channels output are reversed. In this situation the data will be lost since they are processed by the bit synchronizers and detectors of different data rates. It is the purpose of this report to investigate the cause of the I-Q channel reversal problem and proposes possible fixes for the phenonmenon.

Fig. 1 PROBLEM STATEMENT WITH TAPE RECORDER PRESENT



1.1 Problem Scenario

It has been demonstrated by tests performed by the NASA JSC Electronic System Test Laboratory (ESTL) that I-Q reversals occur when the Operational Data Tape Recorder (OPS Recorder) starts-up and wraps around. During this time, the recorder generates noise which is broadband relative to the data rate of interest. For the purpose of analyzing this problem, the noise can be treated as white Gaussian noise.

2. Executive Summary

Analysis and simulation predict that the broadband noise spectrum generated during the OPS Recorder startup and wraparound will cause MRD I-Q channels reversals. It is the reason that the signal output from the transmitter hard limiter (constant power output) contains a large portion of power of high frequency components relative to the data rate. Part of this high frequency components will be filtered out in the MRD carrier loop arm filters (1.8 MHz RC low pass filter). The signal power from Channel 1 (192 Kbps), however, will be practically unchanged by the 1.8 MHz arm filter. These filterings will change the power ratio of the I-Q channels at the output of the arm filters from the designed 4:1, which in turn will shift the stable lock point of the loop S curve away from zero degrees. In case the stable lock point is close to 90°, I-Q reversal will occur. The shape of the S curves, however, is a function of E_b/N_0. It is also a steady state statistical performance of the loop. Thus from the loop S curves, one can tell statistically how the loop will perform in the steady state.

Tests performed by ESTL demonstrate that inserting a 2.5 MHz 7 pole Butterworth low pass filter can prevent the I-Q channel reversal problem. The analytic simulation result verified this phenonmenon. In summary, the following results are found in the analytic simulation:

- If channel 2 operates normally with random data in it, I-Q reversal is highly improbable for E_b^*/N_0>10 dB, whether or not the 2.5 MHz 7 pole Butterworth filter is present in the transmitter.
- With white Gaussian noise in channel 2 transmitter input
  - In the absence of 2.5 MHz LPF
    - There will be I-Q reversal with high probability for E_b/N_0 < 25 dB
  - In the presence of 2.5 MHz LPF
    - The I-Q reversal will most probably be unobserved for E_b/N_0 > 13 dB
    - The I-Q reversal will somewhat be probable for E_b/N_0 < 8 dB
- With a square wave serves as an input to the Channel 2 transmitter and in the presence of the 2.5 MHz LPF
  - I-Q reversal with high probability occurs if the square wave frequency exceeds 1.7 MHz
  - I-Q reversal will most probably not occur if the square wave frequency is less than 1.7 MHz.

Some of the results above were verified by the ESTL tests and some are to be verified. The results with comments are tabulated in Table 1.

A proposed Motorola MRD software fix is also provided which will update the tracking loop with the current stable lock point.

*E_b refers to the bit energy in channel 2.

LinCom

Table 1 CONCLUSIONS

TAPE RECORDER SIGNAL MODEL	FILTER (2.5MHz)	ANALYTICAL SIMULATION	ESTL TESTS	COMMENTS
NOMINAL (FLAT RANDOM DATA)	ABSENT OR PRESENT	I-Q REVERSAL HIGHLY IMPROBABLE FOR $E_b/N_0>$ 10 dB IN CH 2	AGREEMENT WITH ANALYTICAL SIMULATION	THEORY-PRACTICE AGREE
FILTERED WHITE GAUSSIAN NOISE	ABSENT	1. I-Q REVERSAL WITH HIGH PROB. FOR $E_b/N_0<25$ dB 2. NO I-Q REVERSAL WITH HIGH PROB. FOR $E_b/N_0>30$dB	1. OBSERVED IN ESTL 2. TEST CONDITION NOT SPECIFICALLY ESTABLISHED	1. THEORY-PRACTICE AGREE 2. RECOMMEND ESTL TEST
	PRESENT	1. I-Q REVERSAL MOST PROBABLY UNOBSERVED FOR $E_b/N_0>13$dB 2. I-Q REVERSAL SOME-WHAT PROBABLE FOR $E_b/N_0<8$dB	1. AGREEMENT WITH ANALYTICAL SIMULATION 2. TEST CONDITION NOT SPECIFICALLY TESTED	1. THEORY-PRACTICE AGREE 2. RECOMMEND ESTL TEST
SQUARE WAVE	PRESENT	1. I-Q REVERSAL WITH HIGH PROBABILITY IF SQUARE FRE-QUENCY EXCEEDS 1.7 MHz 2. NO I-Q REVERSAL (MOST PROBABLY) IF FREQUENCY IS <1.7 MHz	1. OBSERVED FOR FREQUENCIES GREATER THAN 2.1MHz 2. NOT OBSERVED FOR 192 kHz	1. THEORY-PRACTICE AGREE 2. THEORY-PRACTICE AGREE

LinCom

Recommendations for the future work are given in the following:

LinCom

RECOMMENDATIONS

- CONDUCT ESTL TESTS TO CONFIRM OTHER SUBLITIES NOT OBSERVED IN ESTL BUT PREDICTED BY ANALYTICAL SIMULATION

  (1) I-Q REVERSAL VANISHING DURING TAPE RECORDER START-UP IF $E_b/N_0 > 28$ to 30 dB AND NO FILTER INSERTED IN SIGNAL PATH

  (2) I-Q REVERSAL IF E_b/N_0 IS REDUCED FROM NOMINAL TO 8 dB WITH 2.5 MHz FILTER INSERTED IN SIGNAL PATH.

  (3) OPTIMIZE FILTER DESIGN BY ANALYTICAL SIMULATION (TBD) AND VERIFY IN ESTL.

- INSERT OPTIMIZED FILTER DESIGN IN SHUTTLE KU-BAND TRANSMITTER AND TEST IN ESTL (MINIMIZES PROBABILITY OF I-Q REVERSAL)
- IMPLEMENT MRD SOFTWARE FIX

LinCom

3. Analytical Simulation Model

The analytical simulation model used is given in Figure 2. The noise generator is used to control the additive channel noise power. The analysis of the loop S curve is given below with the help of Figure 3.

The input signal s(t) is in the form

$$s(t) = \sqrt{2P_1}\, d_1(t) \cos(\omega_0 t + \theta) + \sqrt{2P_2}\, d_2(t) \sin(\omega_0 t + \theta) \qquad (1)$$

where P_1 and P_2 are the power, and $d_1(t)$ and $d_2(t)$ are independent binary data, of the I and Q channels respectively. The phase θ is unknown and to be estimated by the carrier loop. The Gaussian channel noise can be expressed in the form of a narrow-band process about the carrier frequency

$$n(t) = \sqrt{2}[N_c(t)\cos(\omega_0 t+\theta) - N_s(t)\sin(\omega_0 t+\theta)] \qquad (2)$$

where $N_c(t)$ and $N_s(t)$ are approximately statistically independent, stationary, white Gaussian noise processes with single-sided noise spectral density N_0 W/Hz and single-sided bandwidth $B_H < \omega_0/2\pi$.

Let

$$\gamma_c(t) = \sqrt{2}\cos(\omega_0 t+\hat{\theta}) \qquad (3)$$

$$\gamma_s(t) = \sqrt{2}\sin(\omega_0 t+\hat{\theta}) \qquad (4)$$

Neglect the double frequency terms at the output of the multipliers, and let $\phi = \theta-\hat{\theta}$,

Fig. 2 ANALYTICAL-SIMULATION MODEL

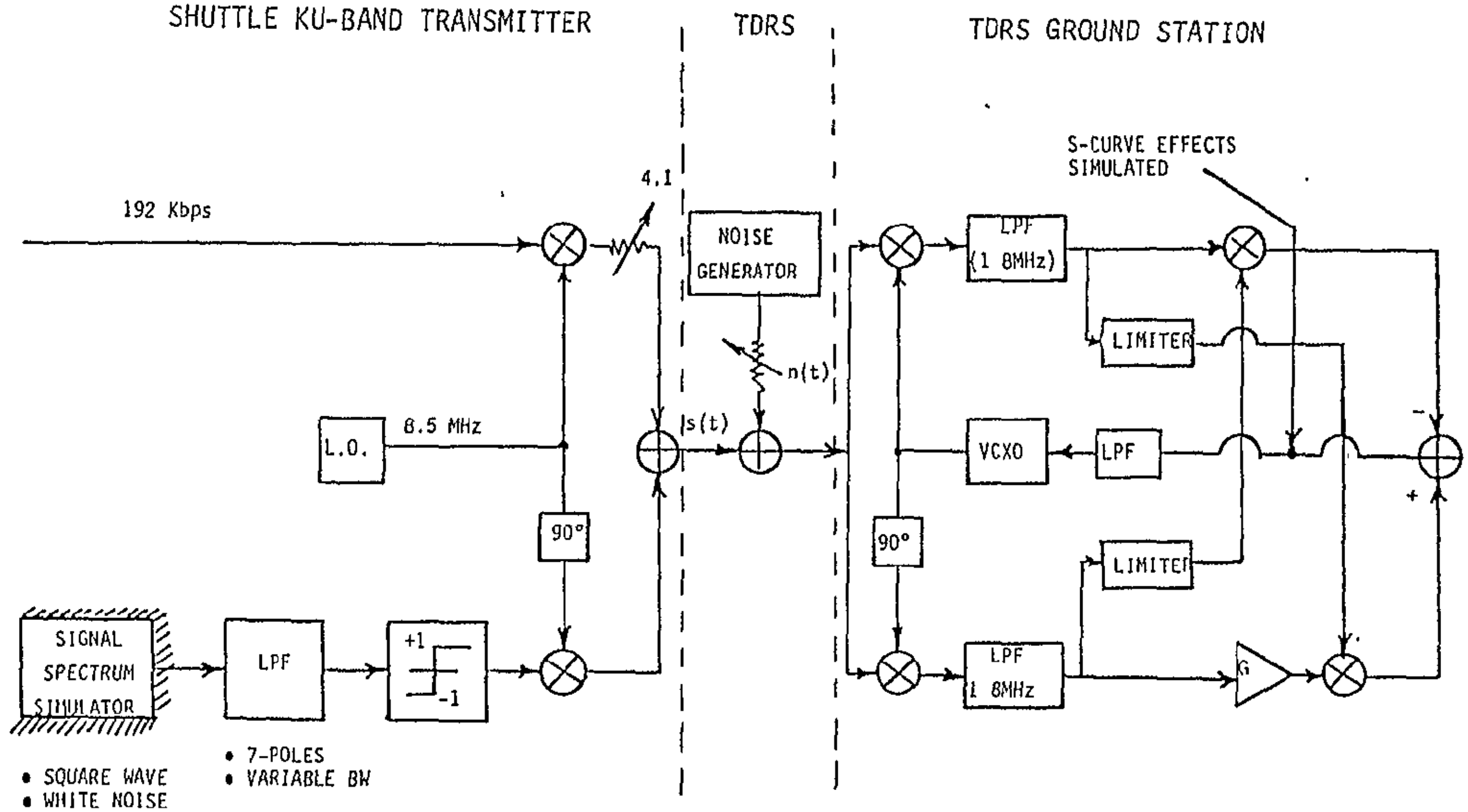
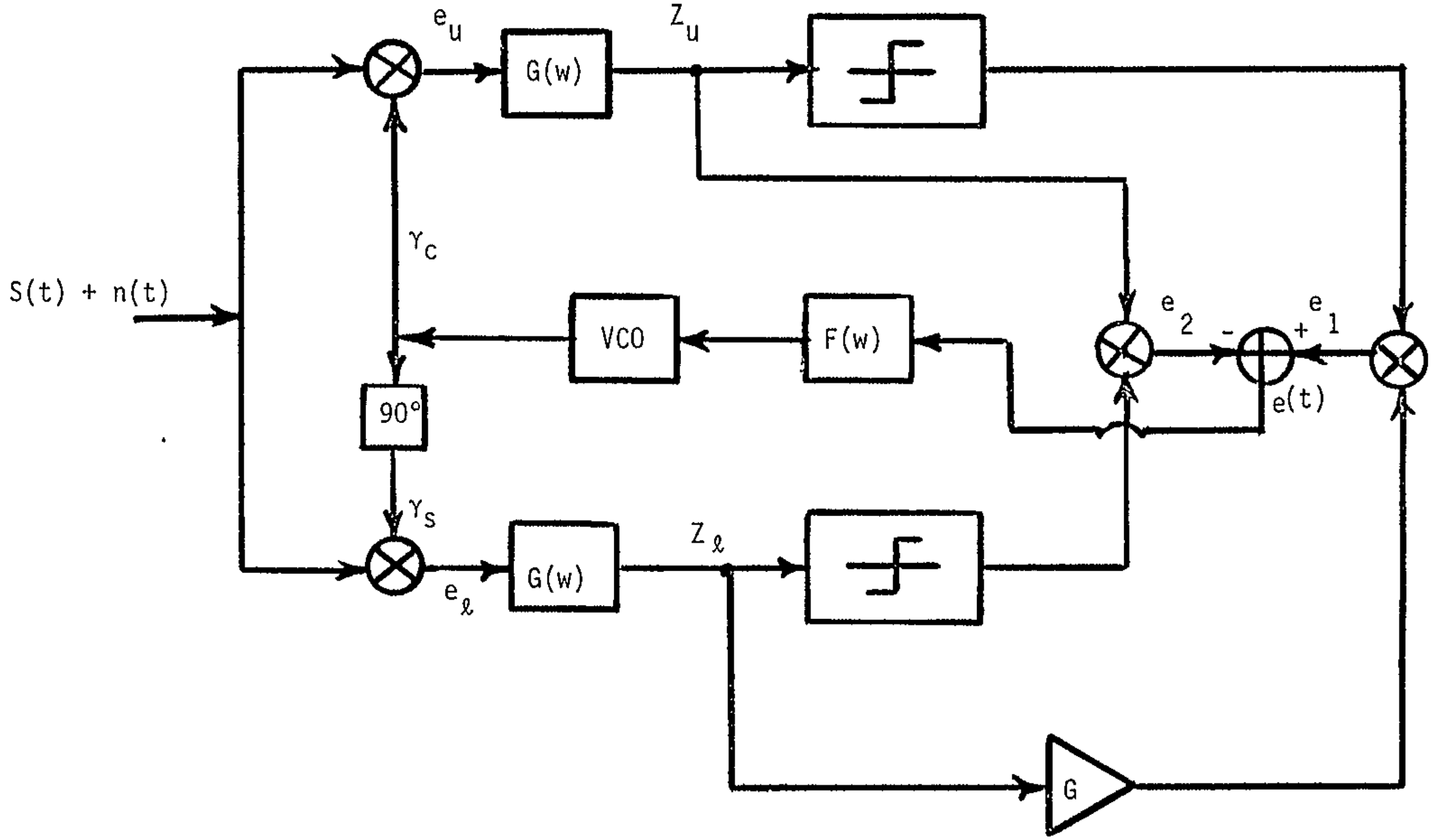


Fig. 3 MRD Carrier Tracking Loop



$$e_u(t) = \sqrt{P_1}d_1(t)\cos\phi+\sqrt{P_2}\, d_2(t)\sin\phi+N_c(t)\cos\phi-N_s(t)\sin\phi \quad (5)$$

$$e_\ell(t) = -\sqrt{P_1}d_1(t)\sin\phi+\sqrt{P_2}d_2(t)\cos\phi-N_c(t)\sin\phi-N_s(t)\cos\phi \quad (6)$$

To keep the problem manageable the effect of the filter on the signal will be considered in this section as a reduction in signal power only, i.e., the filter outputs are rectangular waveforms corrupted by Gaussian noise:

$$Z_u(t) = \sqrt{\ell P_1}\, d_1(t)\cos\phi+\sqrt{\ell P_2}d_2(t)\sin\phi+N_c(t)\cos\phi-N_s(t)\sin\phi \quad (7)$$

$$Z_\ell(t) = -\sqrt{\ell P_1}d_1(t)\sin\phi+\sqrt{\ell P_2}d_2(t)\cos\phi-N_c(t)\sin\phi-N_s(t)\cos\phi \quad (8)$$

where ℓ represents the filtering degradation in signal power.

The error signals are given by

$$e_1(t) = GZ_\ell(t)\mathrm{sgn}(Z_u(t)) \quad (9)$$

$$e_2(t) = Z_u(t)\mathrm{sgn}(Z_\ell(t)) \quad (10)$$

$$G = 2 \quad (11)$$

The loop S curve is then given by

$$g(\phi) = E[e(t)] = E[e_1(t)] - E[e_2(t)] \quad (12)$$

where the notation $E[\cdot]$ represents the average over the noise and independent data sequences d_1 and d_2. The gain G in this case is 2.

4. Numerical Results

In this section the numerical results of the analytic simulation are provided. The physical meanings of these results are discussed. Fig. 4 shows the normalized MRD S curves with tape recorder test signal modeled as a 0.96 MHz random data. The curves are plotted as a function of channel 2 E_b/N_0. As shown in this figure, there is a stable lock point at zero degree. I-Q reversal is improbable for $E_b/N_0 > 3$ dB.

In Fig. 5, normalized MRD S curves with tape recorder test signal modeled as white noise are shown with E_b/N_0 as a parameter. The 2.5 MHz LPF is absent for this figure. I-Q reversals will occur with high probability for $E_b/N_0 < 25$ dB. This situation was observed in ESTL tests. It is predicted that no I-Q reversal will occur wih high probability for $E_b/N_0 > 30$ dB. No ESTL test was made in the condition and future test is recommended for verification.

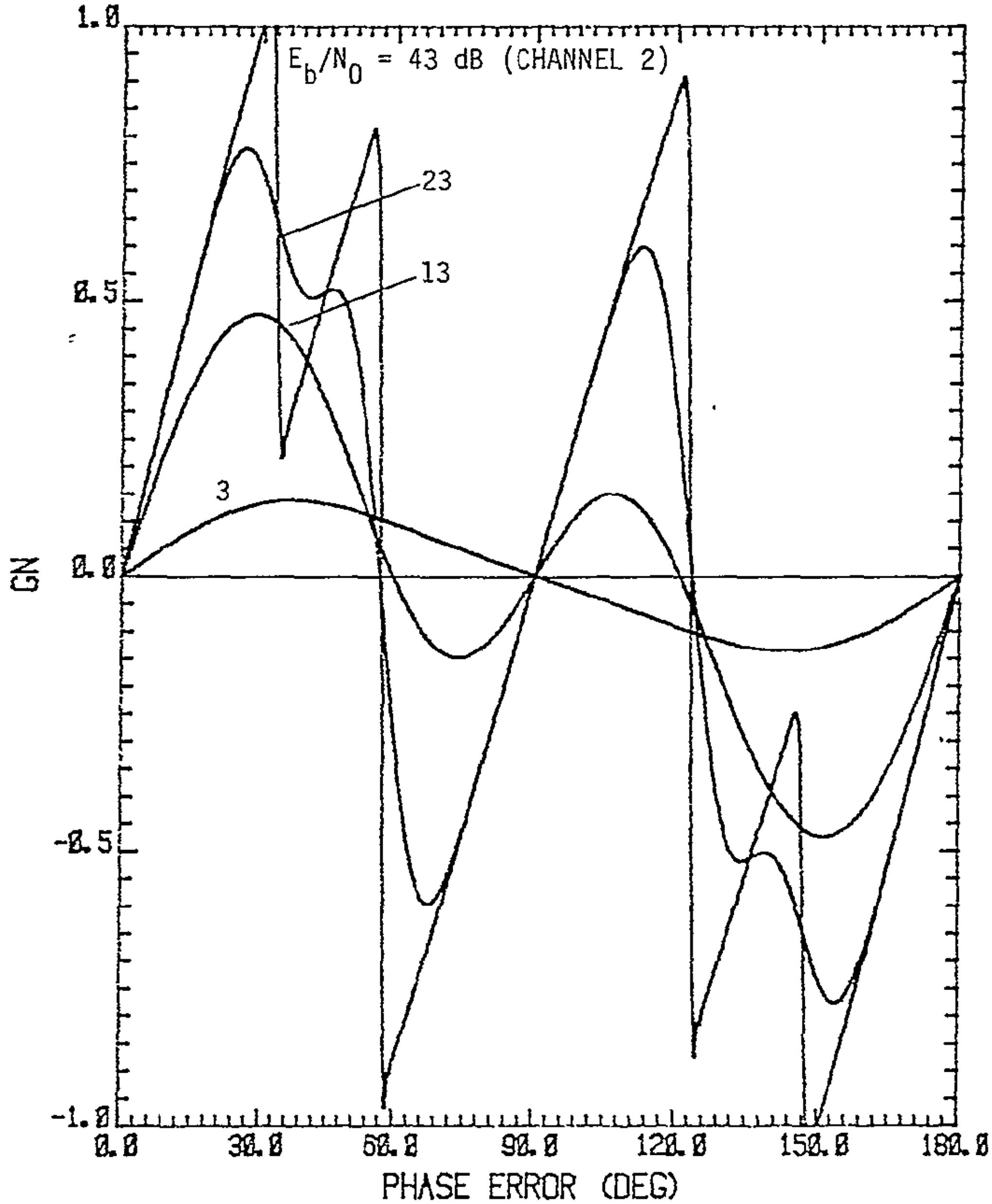
Figs. 6 and 7 are the same as Fig. 5 except the 2.5 MHz LPF is present. It can be seen that the stable lock point at zero degree is established for $E_b/N_0 > 13$ dB. The I-Q reversal will be most probably unobserved for $E_b/N_0 > 13$ dB. On the other hand, it will be somewhat probable for $E_b/N_0 < 8$ dB. ESTL tests are recommended to confirm these results.

ESTL tests were performed with a square wave test signal applied to channel 2. Fig. 8 illustrates the MRD tracking performance with square waves of different frequency applied. It is shown that with relatively high E_b/N_0, 20 dB or above for example, I-Q reversal will occur with high probability if square wave frequency is greater than 1.7 MHz. ESTL tests for 192 KHz and 2.1 MHz verified these results. A summary of the analytic-simulation is provided in Table 1.

LinCom

Fig. 4 NORMALIZED S-CURVES WITH TAPE RECORDER TEST SIGNAL

MODELED AS A 0.96 MHz FLAT RANDOM DATA

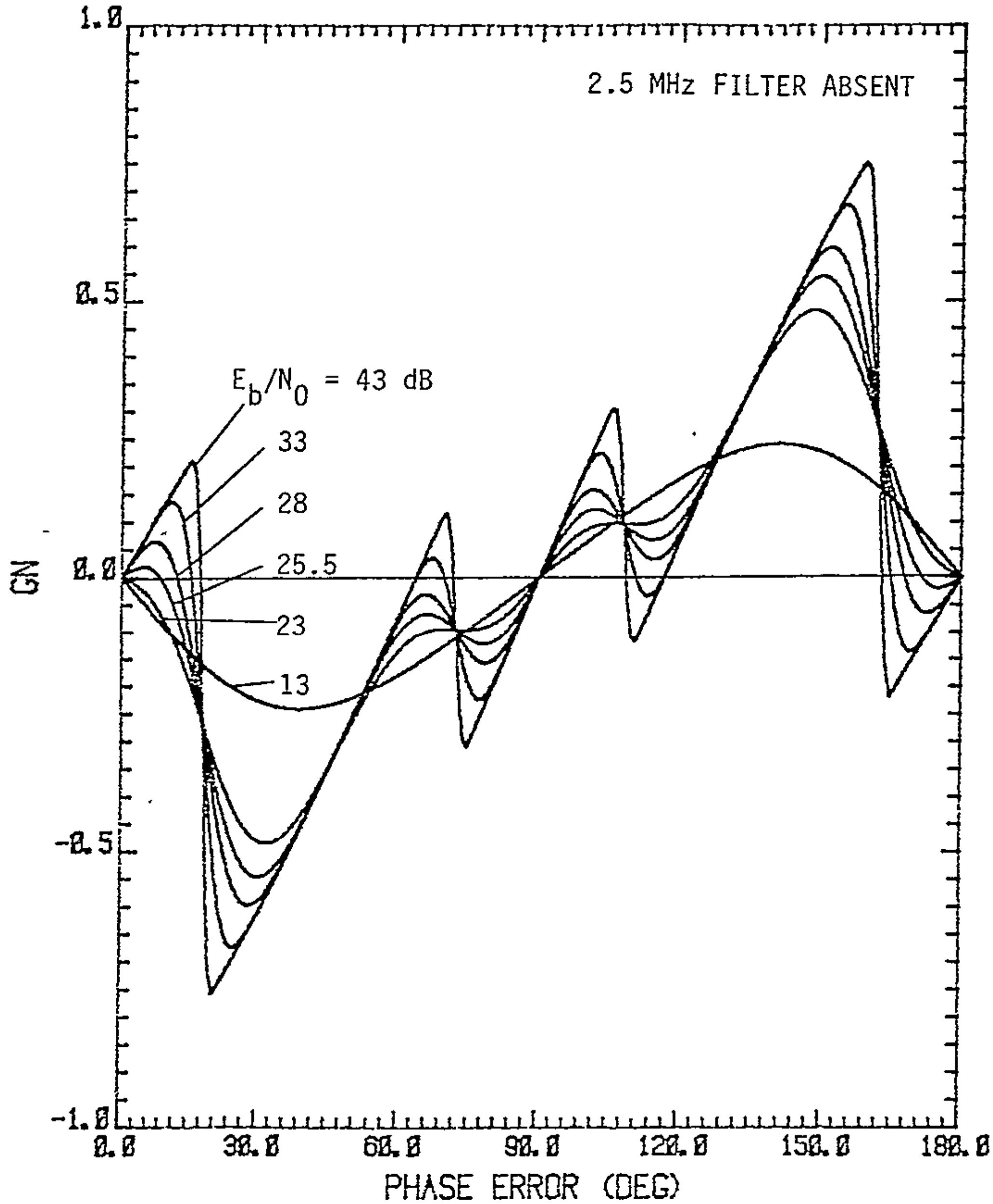


CONCLUSION: I-Q REVERSAL HIGHLY IMPROBABLE FOR E_b/N_0>3 dB

LinCom

LinCom

Fig. 5 NORMALIZED MRD S-CURVES WITH TAPE RECORDER TEST SIGNAL MODELED AS WHITE NOISE



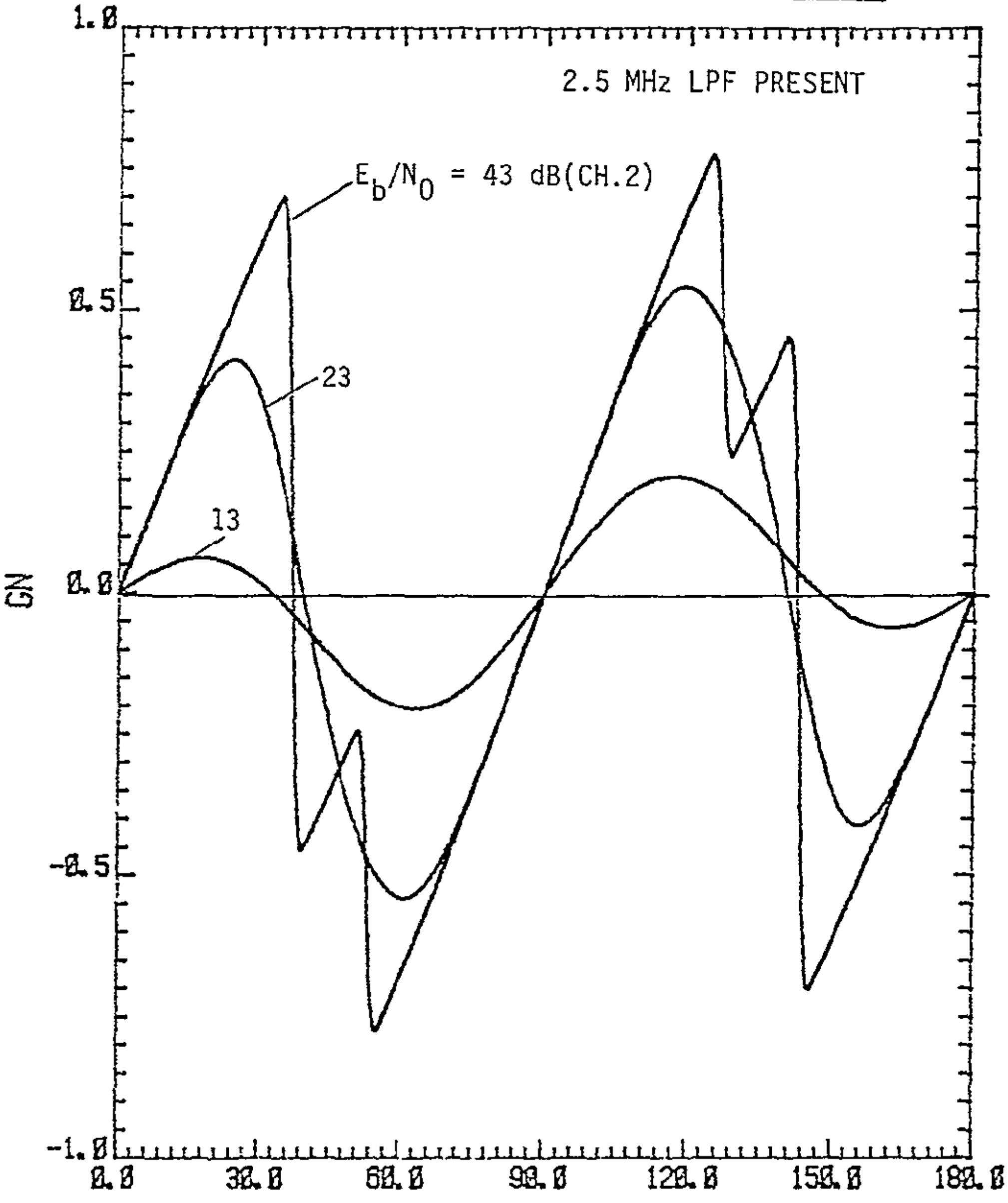
CONCLUSION: (a) I-Q REVERSAL WITH HIGH PROBABILITY FOR $E_b/N_0 < 25$ dB. (OBSERVED IN ESTL)

(b) NO I-Q REVERSAL WITH HIGH PROBABILITY FOR $E_b/N_0 > 30$ dB

LinCom

LinCom

Fig. 6 NORMALIZED MRD S-CURVES WITH TAPE RECORDER TEST SIGNAL MODELED AS FILTERED WHITE GAUSSIAN NOISE

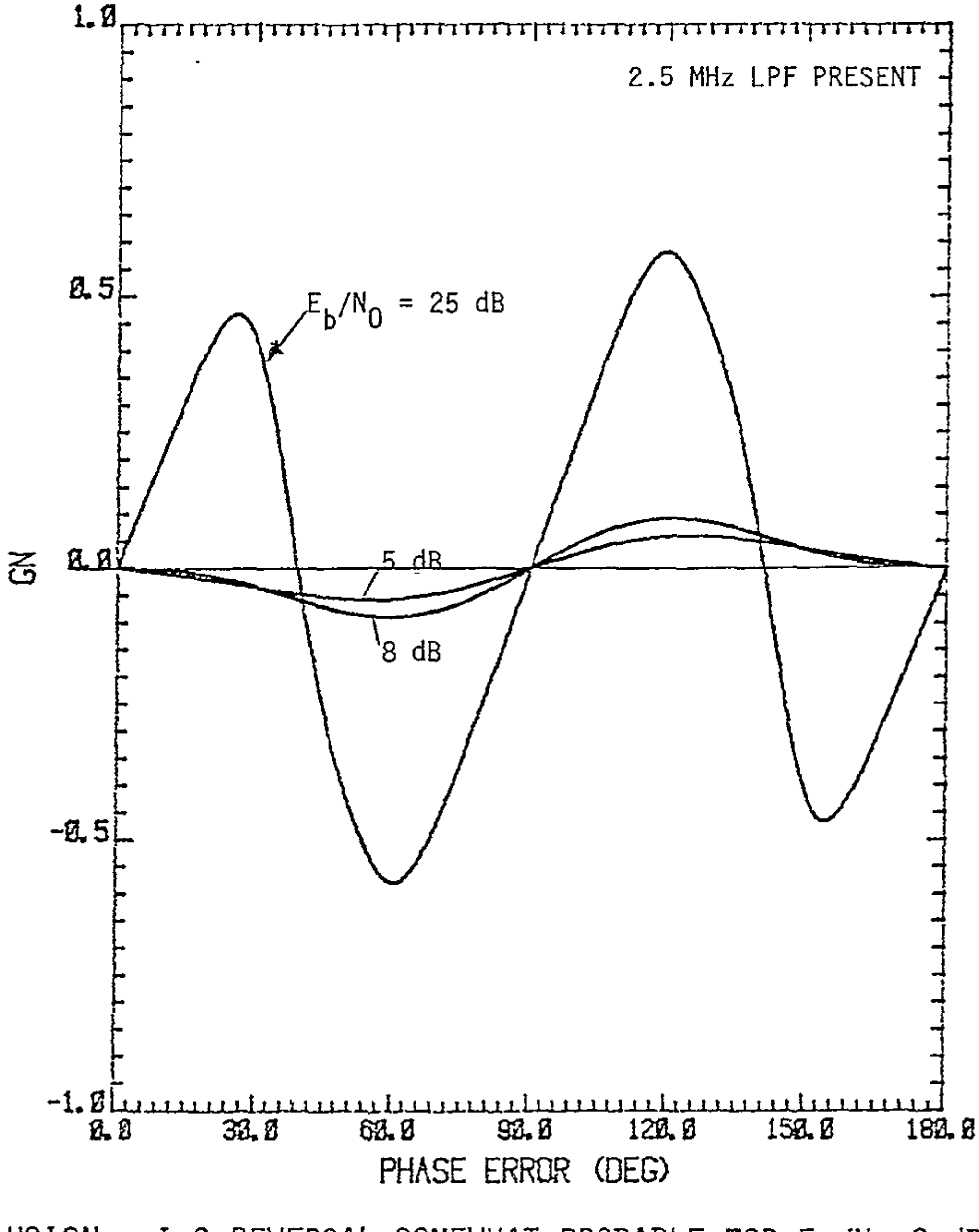


CONCLUSION: I-Q REVERSAL MOST PROBABLY UNOBSERVED FOR $E_b/N_0 > 13$ dB

LinCom

LinCom

Fig. 7 NORMALIZED MRD S-CURVES WITH TAPE RECORDER TEST SIGNAL MODELED AS WHITE GAUSSIAN NOISE

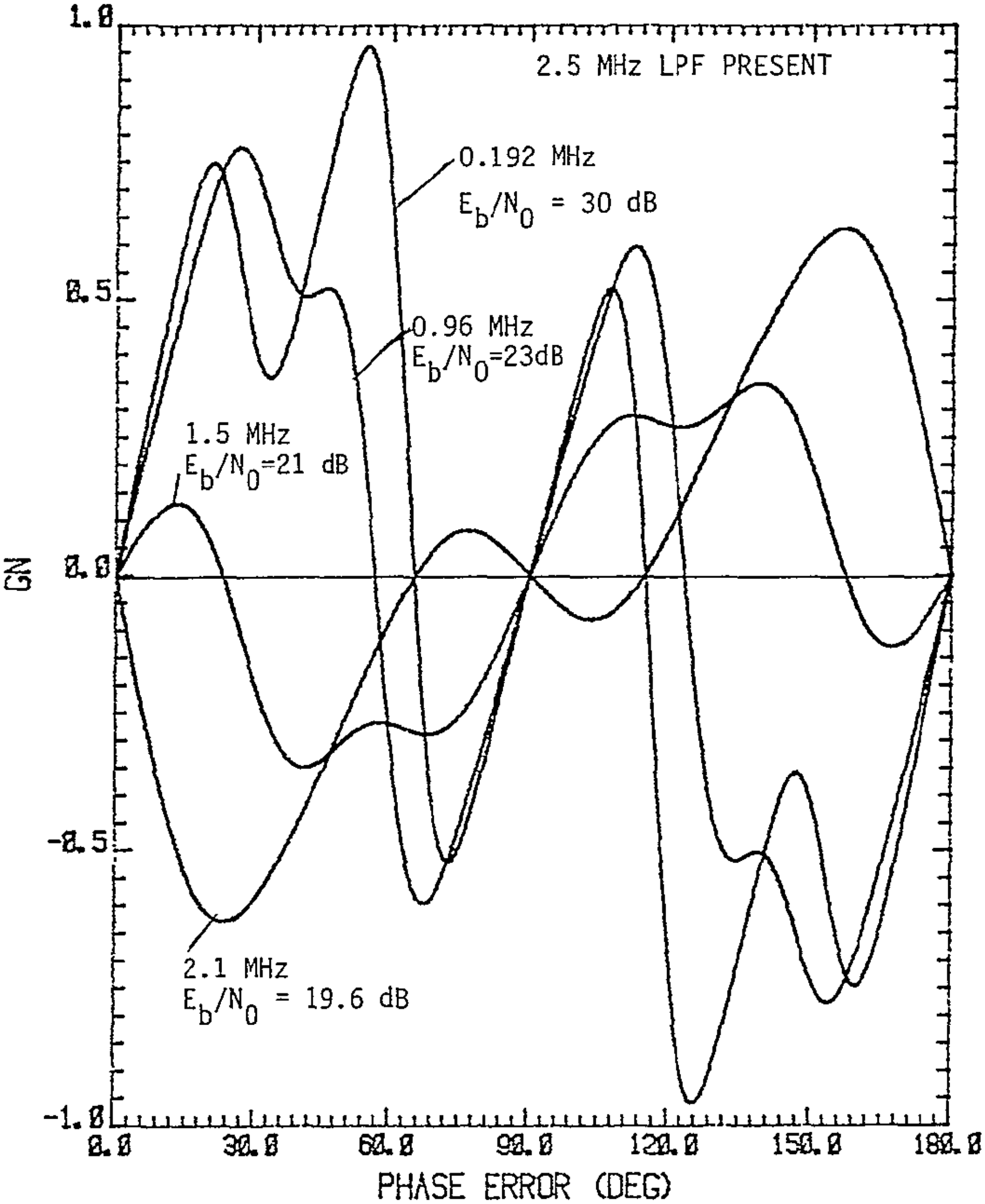


CONCLUSION: I-Q REVERSAL SOMEWHAT PROBABLE FOR E_b/N_0<8 dB

LinCom

LinCom

Fig. 8 NORMALIZED MRD S-CURVES WITH A SQUARE WAVE TEST SIGNAL FREQUENCY IN CHANNEL 2



CONCLUSION: I-Q REVERSAL WITH HIGH PROBABILITY IF SQUARE WAVE FREQUENCY IS GREATER THAN 1.7 MHz

LinCom

5. Matched NRZ Signal Power

It is helpful in estimating the portion of signal power at the output of the hard limiter matched to the NRZ waveform. Fig. 9 is the model of the signal power estimation. A random white noise passes through in cascade a RC LPF of 3 dB bandwidth W_c, a hard limiter and a NRZ waveform matched filter. The output signal power is estimated. This value would be approximately the power available in the carrier tracking loop if the arm filters were matched to the signal waveform. Fig. 10 shows this matched signal power as a function of the RC LPF 3 dB bandwidth W_c. As shown, a power of approximately 7.4 dB degradation arises if the RC LPF bandwidth increases from 2.5 MHz to 15 MHz.

6. Software Fix

Software fix is another step in fixing the MRD I-Q reversal problem. In the present Motorola MRD software, once the loop is in lock, it does not check again whether the loop later locks, whatever the causes are, into an undesirable lock point. This procedure has to be remedied, although it is not a complete fix of the MRD I-Q reversal problem.

Fig. 11 is the current Motorola MRD software flowchart. Fig. 12 is the proposed fix for the software. It forces the software to update itself to the current stable lock point. However, the 2.5 MHz LPF is still a necessary hardware fix; since in the absence of the LPF, the loop will create an undesirable ($\approx 90^0$ offset) stable lock point which causes the I-Q reversal.

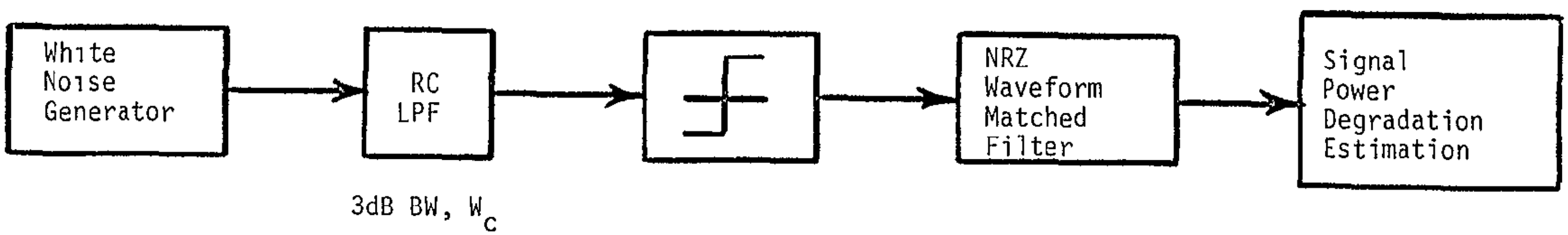


Fig. 9 Model in Estimating Signal Power Matched to the NRZ Waveform

LinCom

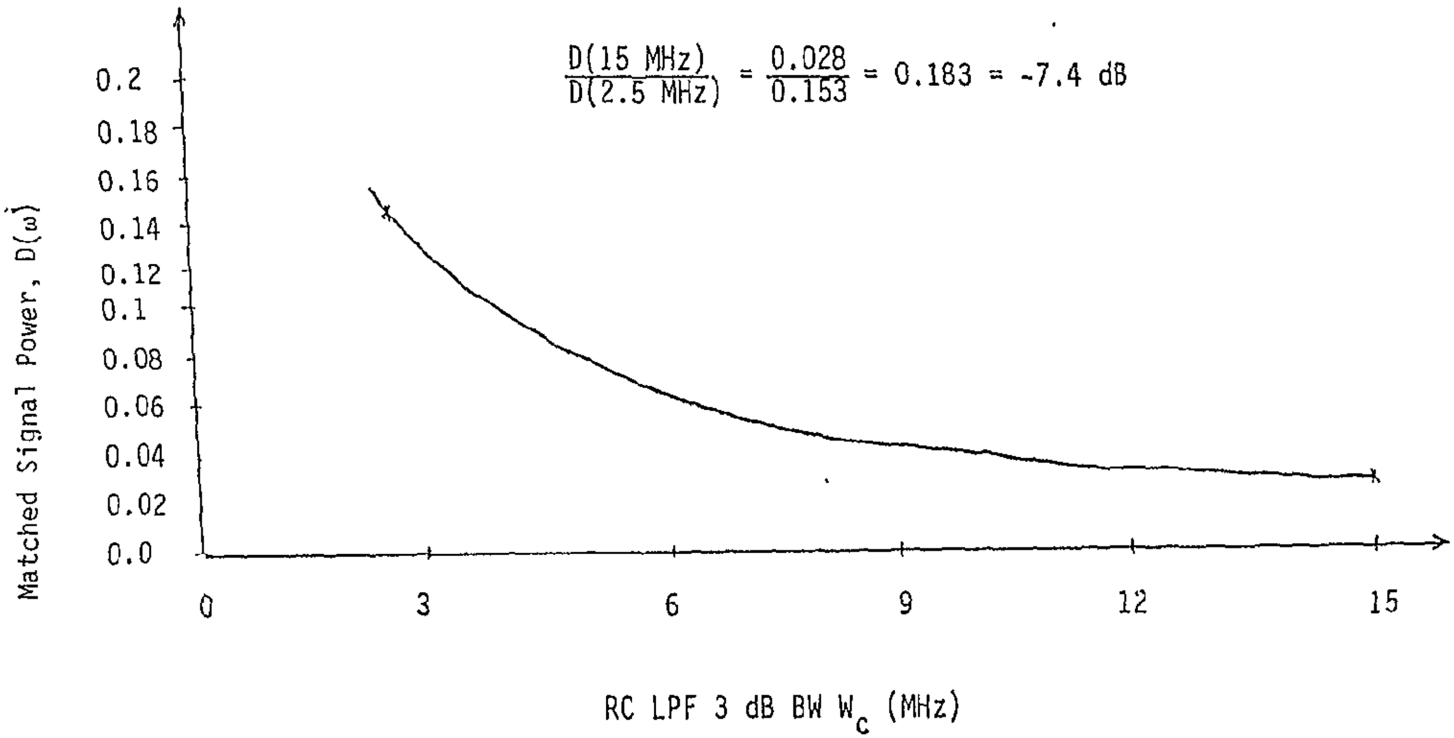
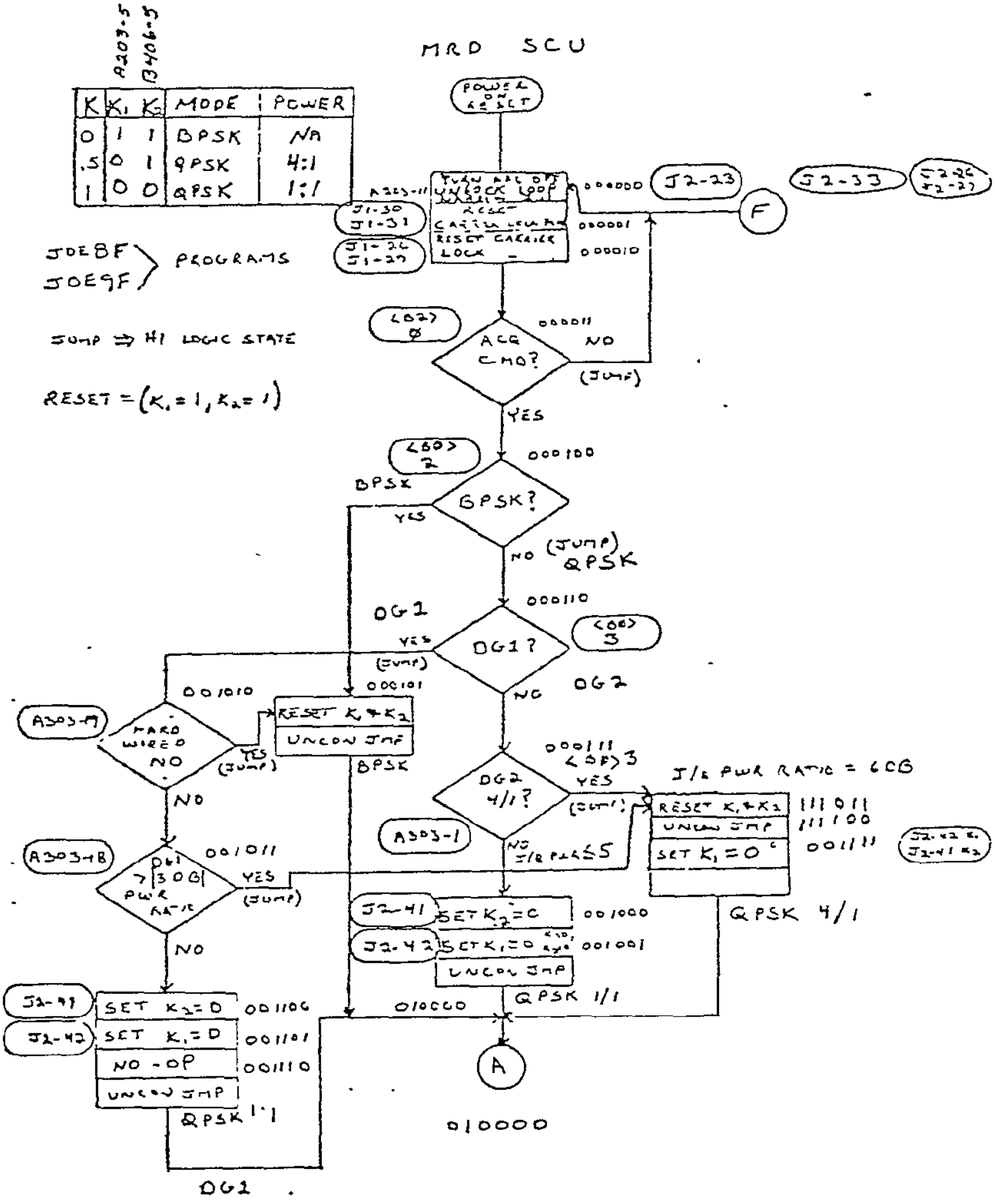


Fig. 10 Signal Power Degradation Versus W_c

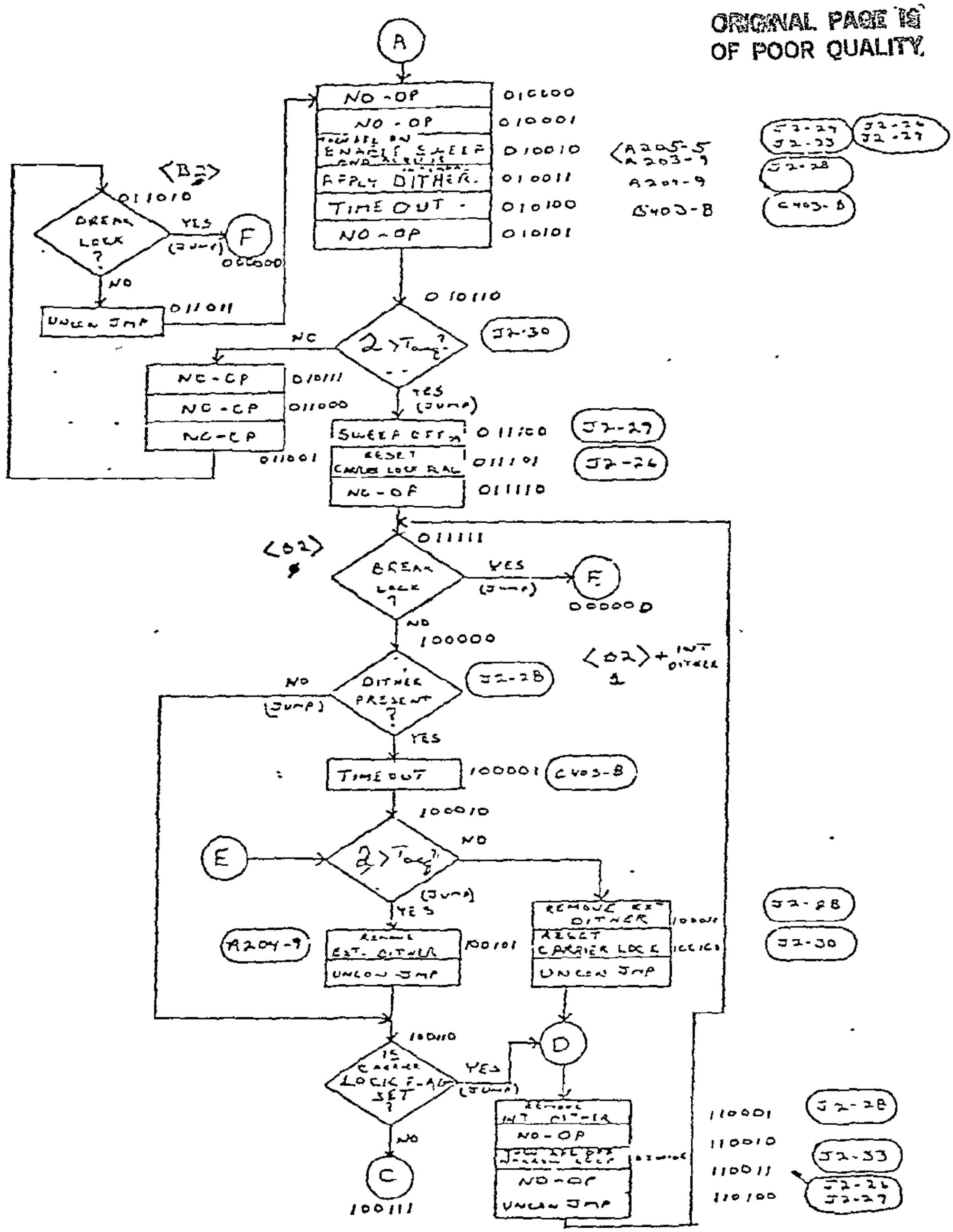
LinCom

Fig. 11a Motorola MRD Software Flowchart



LinCom

Fig. 11b Original Motorola MRD Software Flowchart



LinCom

Fig. 11c Loop to Correct the I,Q Channel Reversal

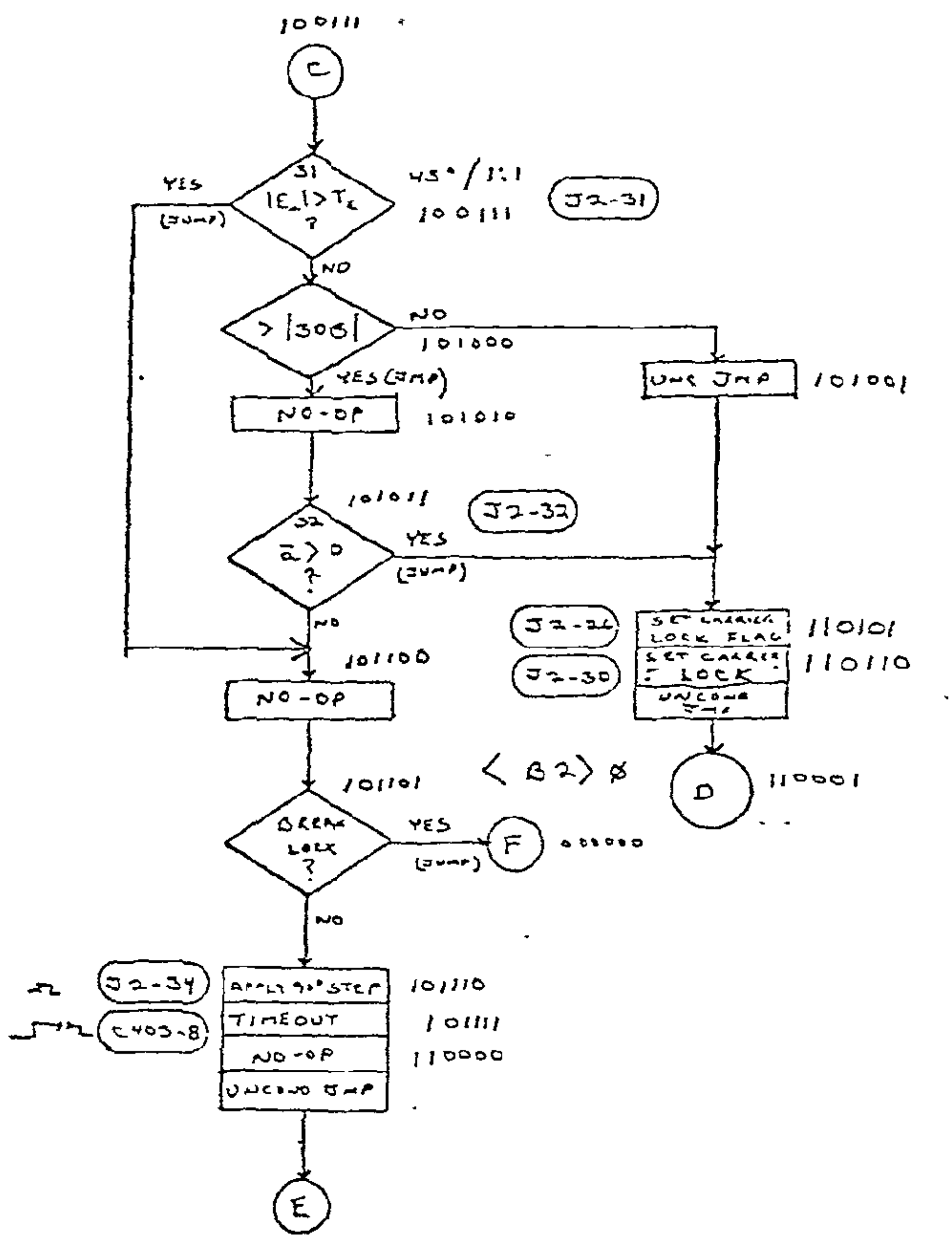
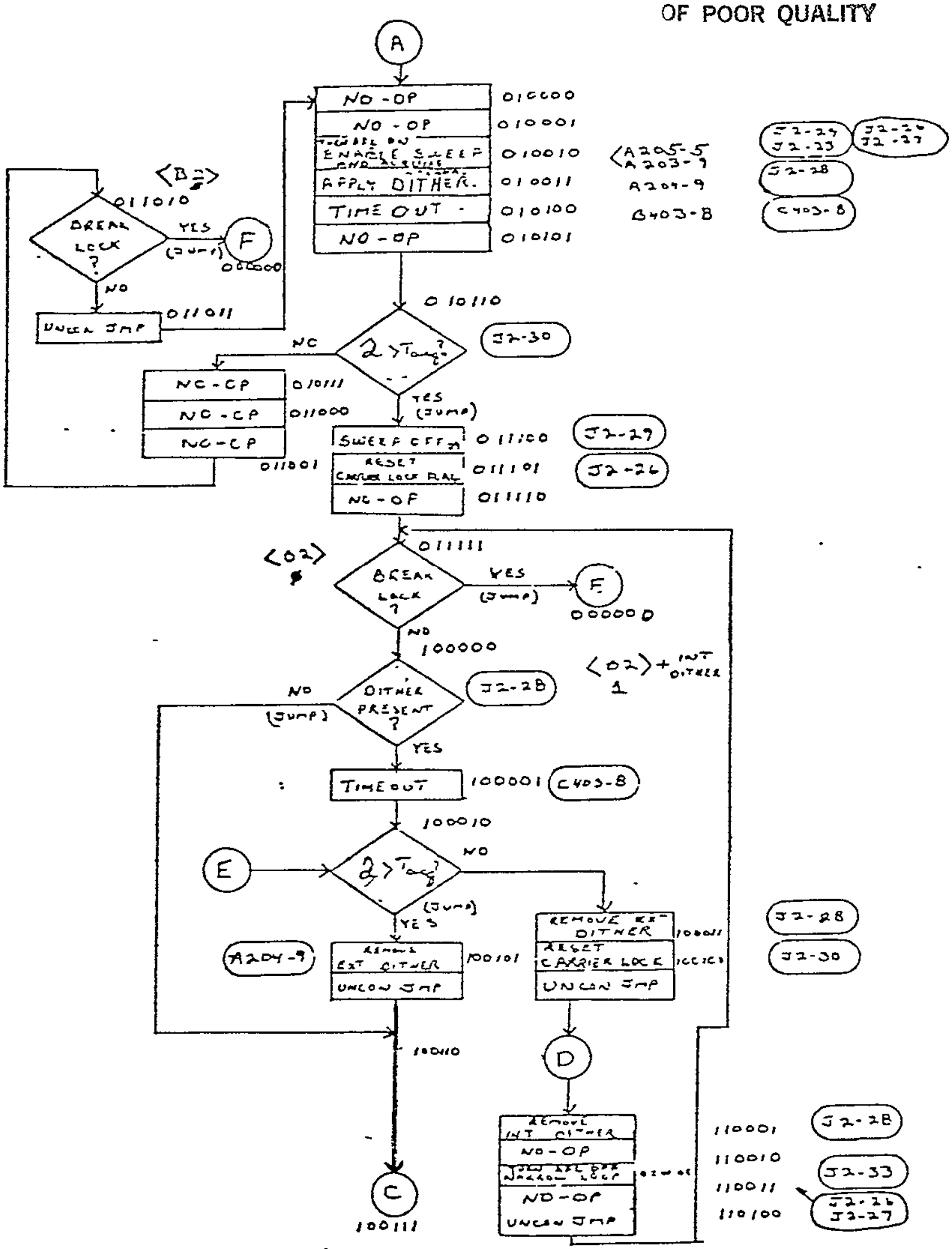


Fig. 12 Proposed Fix on MRD Software

ORIGINAL PAGE IS
OF POOR QUALITY



APPENDIX B

LASER COMMUNICATION SYSTEM TECHNOLOGY

The key components of the laser communication system and the advantages in using them were addressed. Typical performance bit error rate was evaluated.

KEY ISSUES FOR SS LINK PLANNING AND DESIGN

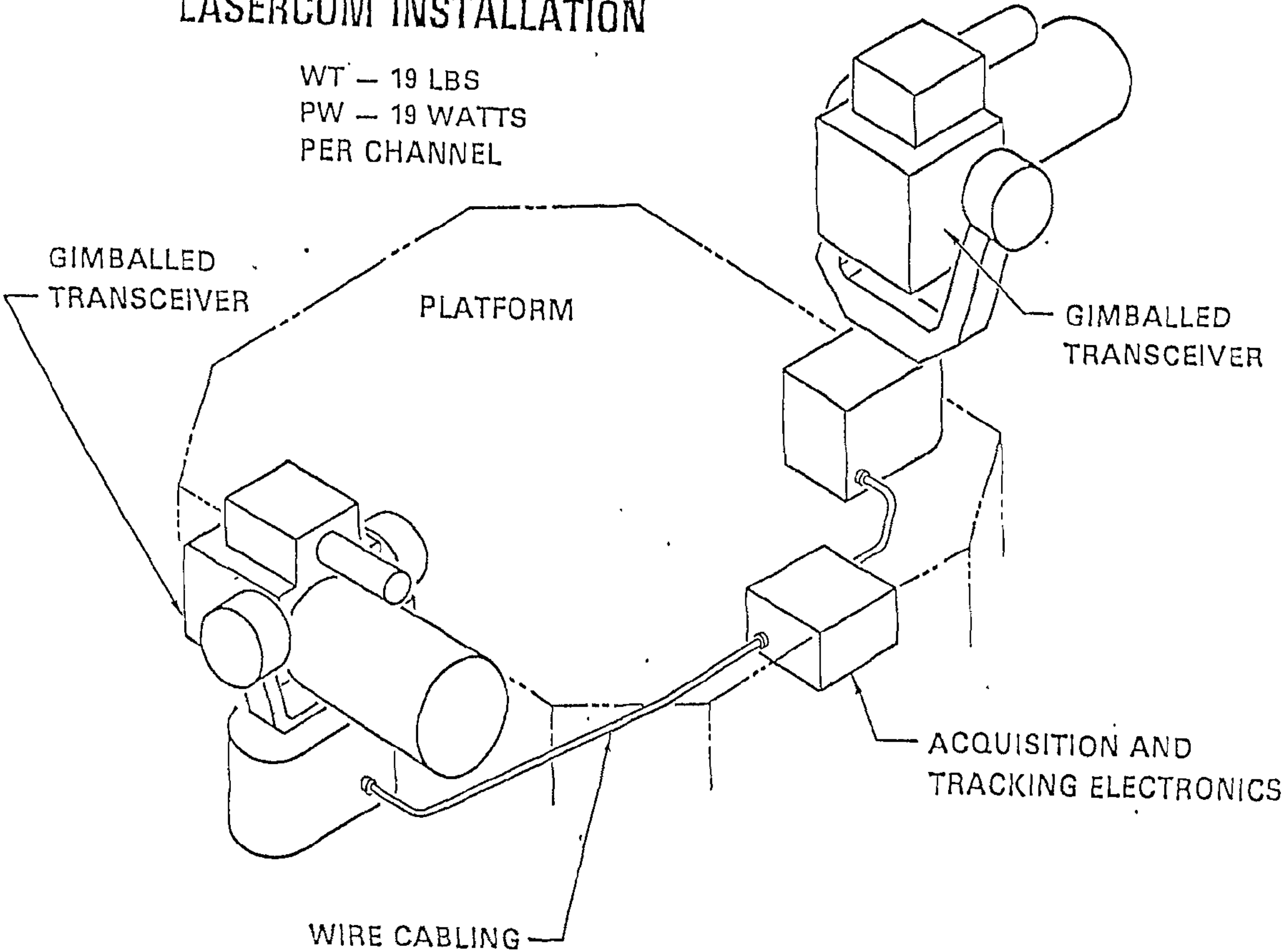
- EHF TECHNOLOGY VS OPTICAL TECHNOLOGY
- COMPATIBILITY OF WAVE FORM DESIGN WITH ANTENNA SYSTEM
- MULTIPLE ACCESS TECHNIQUE CHOICE
- USER HARDWARE COMPLEXITY AND COST
- INTERFERENCE
- NEAR FAR PROBLEM

LinCom Experience

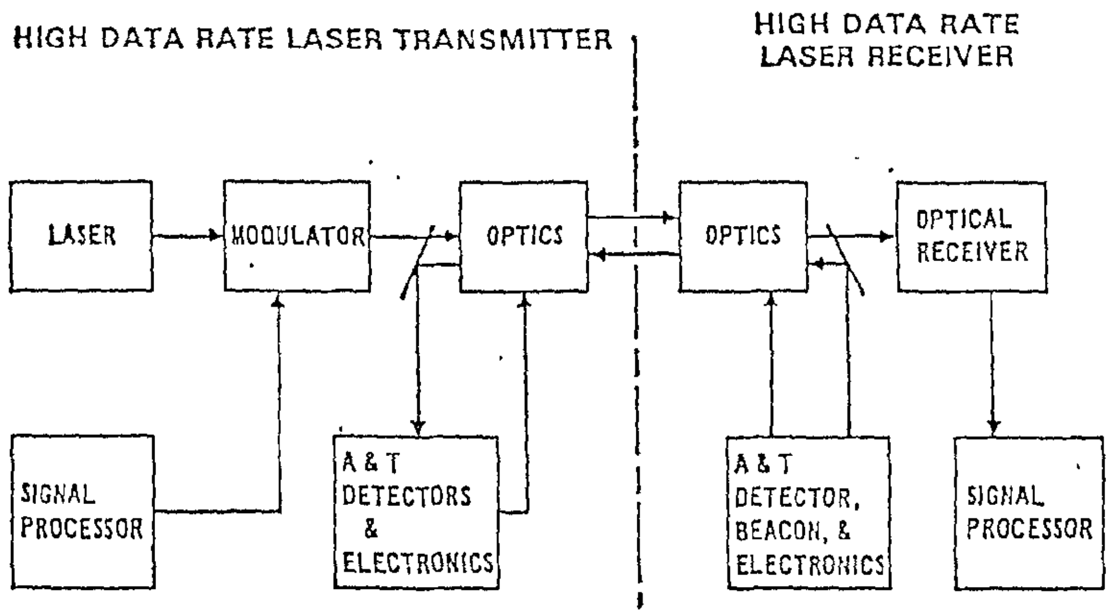
- Development of Specifications
- Clock Synchronization For Laser X-Links
- Time and Frequency Transfer System Simulation/Evaluation
- X-Link Ranging System Simulation/Evaluations
- Testing and Hardware Evaluation

LASERCOM INSTALLATION

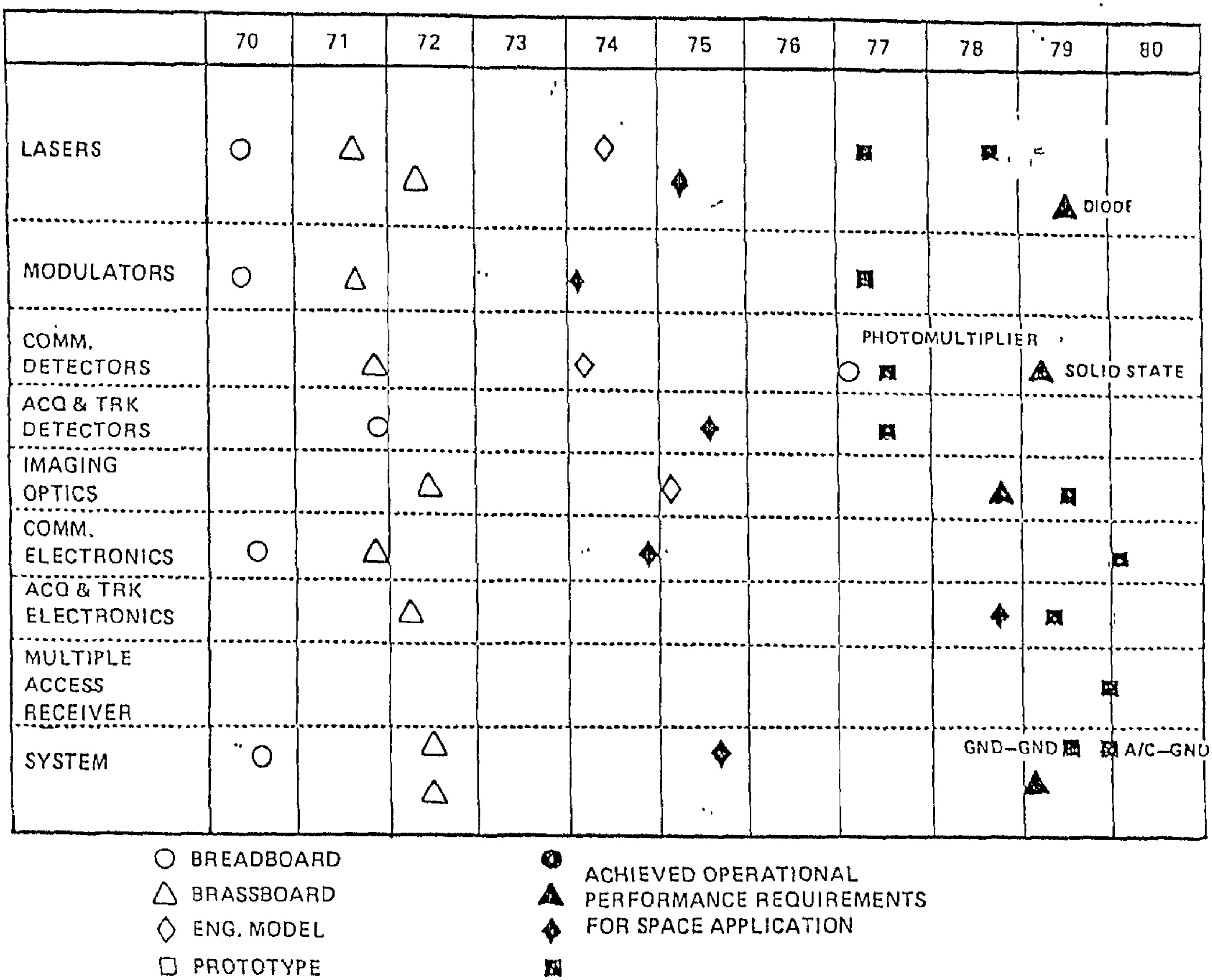
WT — 19 LBS
PW — 19 WATTS
PER CHANNEL



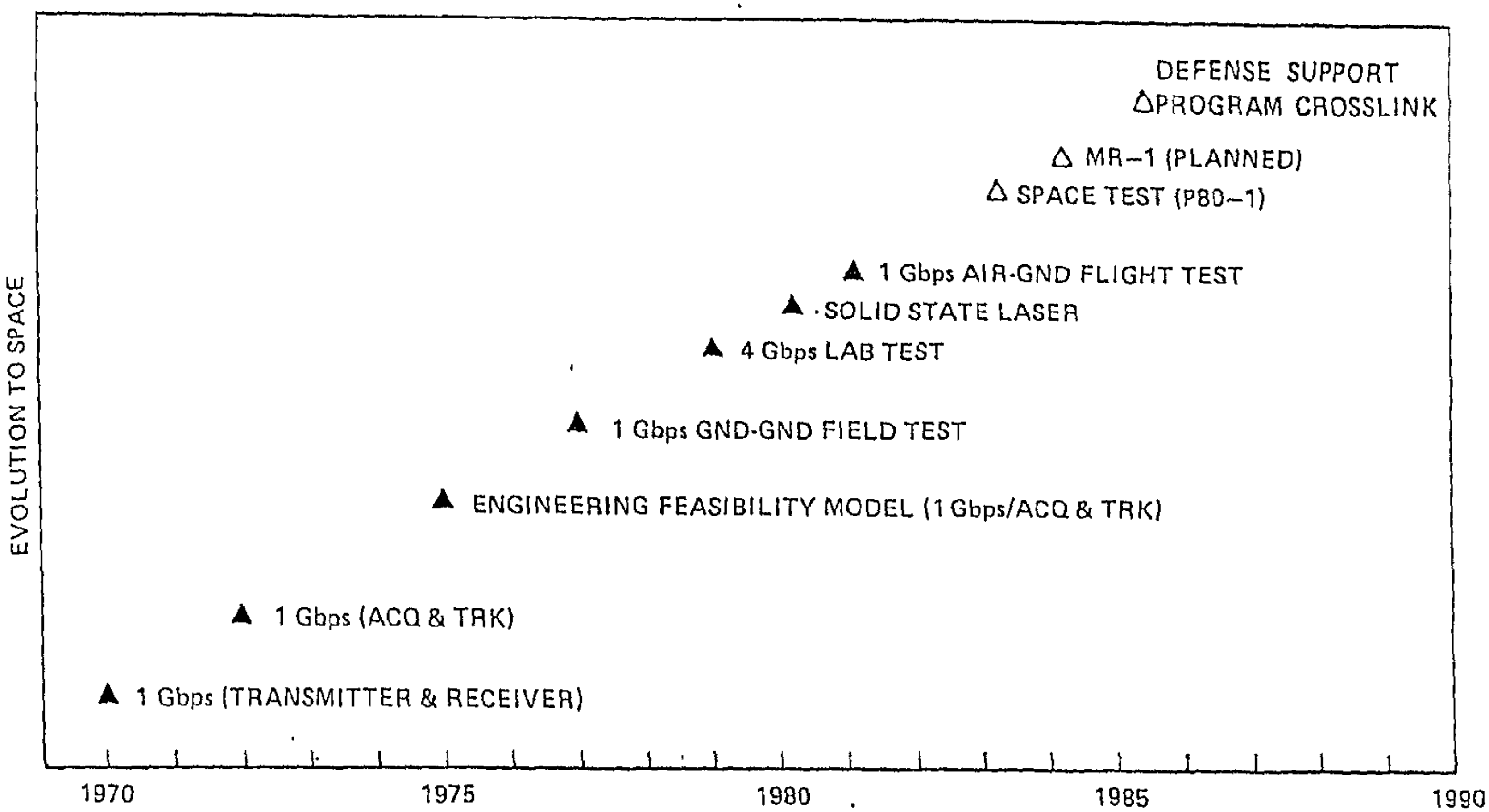
COMMUNICATION SYSTEM BLOCK DIAGRAM



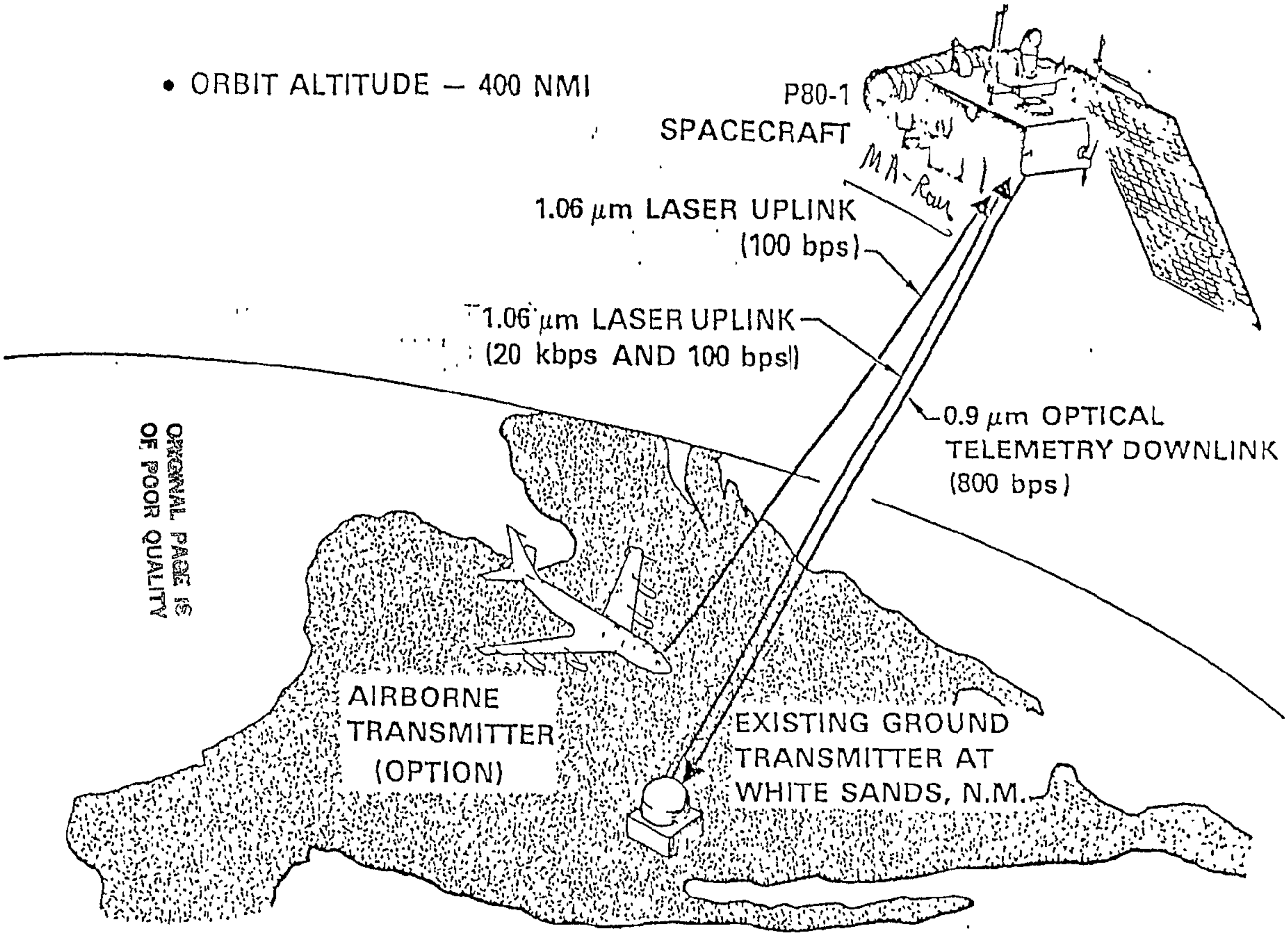
LASERCOM HARDWARE MATURITY



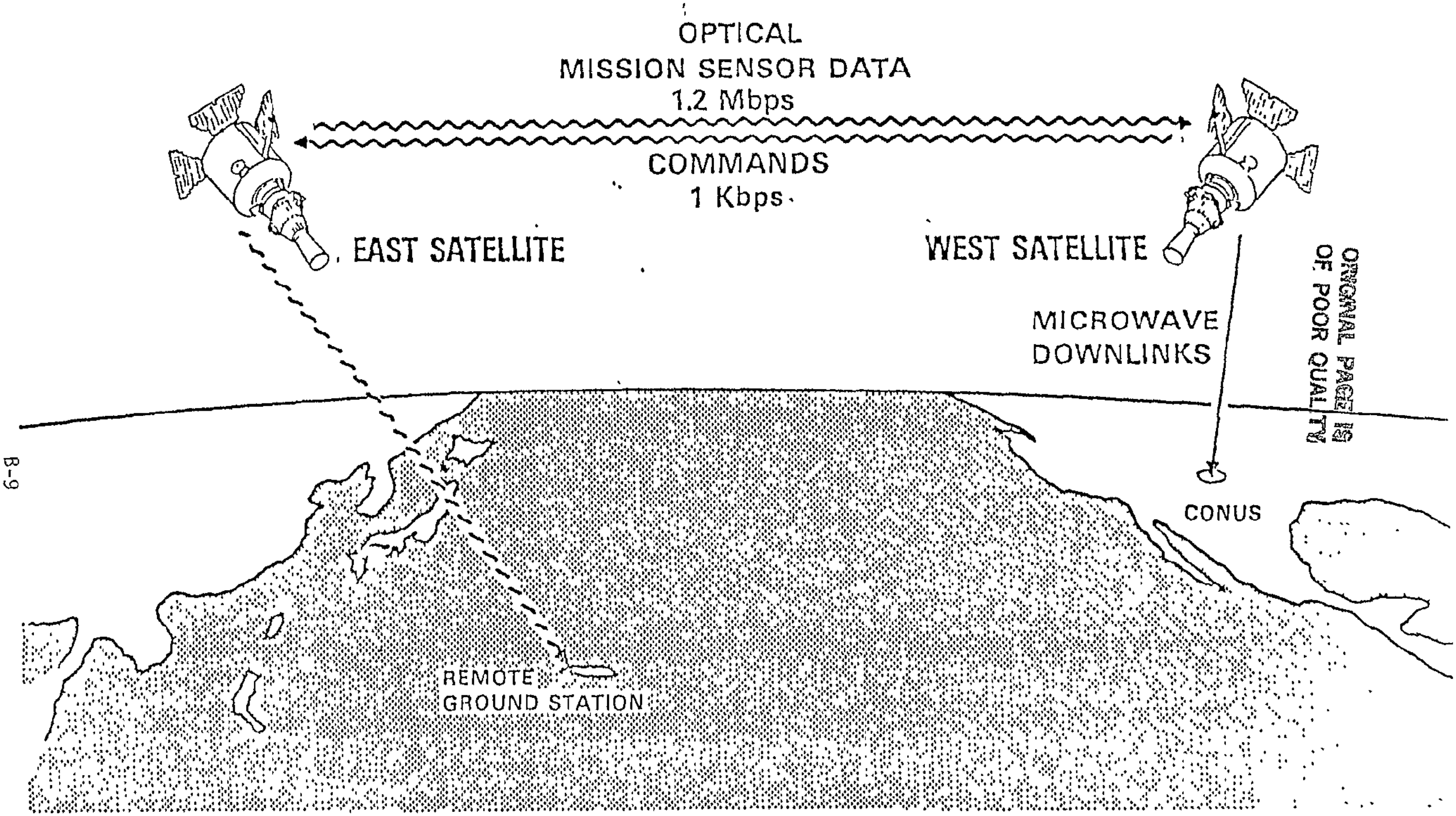
LASERCOM SYSTEM MILESTONES



LSMU EXPERIMENT CONCEPT

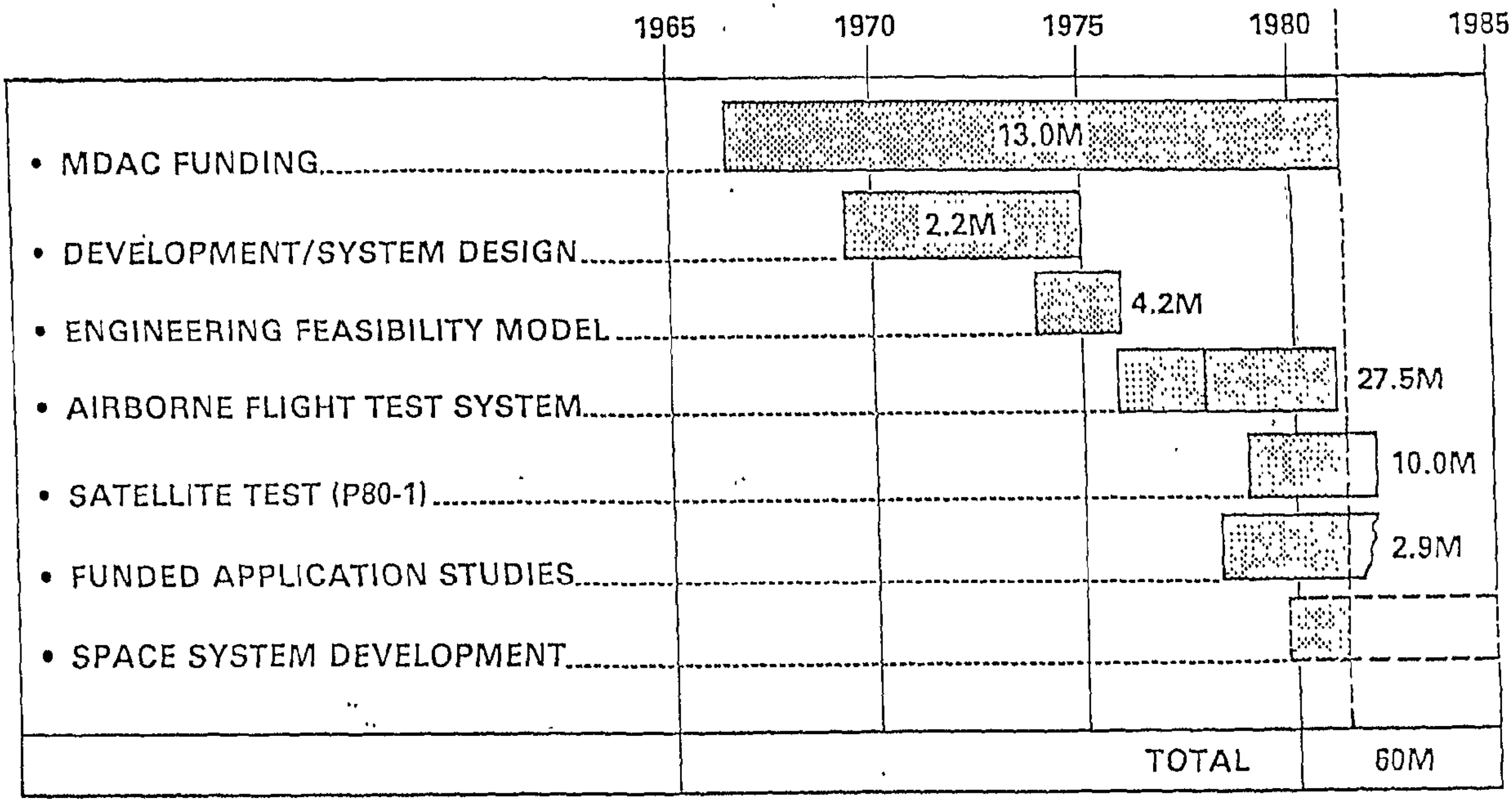


DEFENSE SUPPORT PROGRAM (DSP) LASER CROSSLINK REDUCES THE NEED FOR A REMOTE GROUND STATION



AIR FORCE LASER COMMUNICATIONS PROGRAMS

REAL TIME DOLLARS



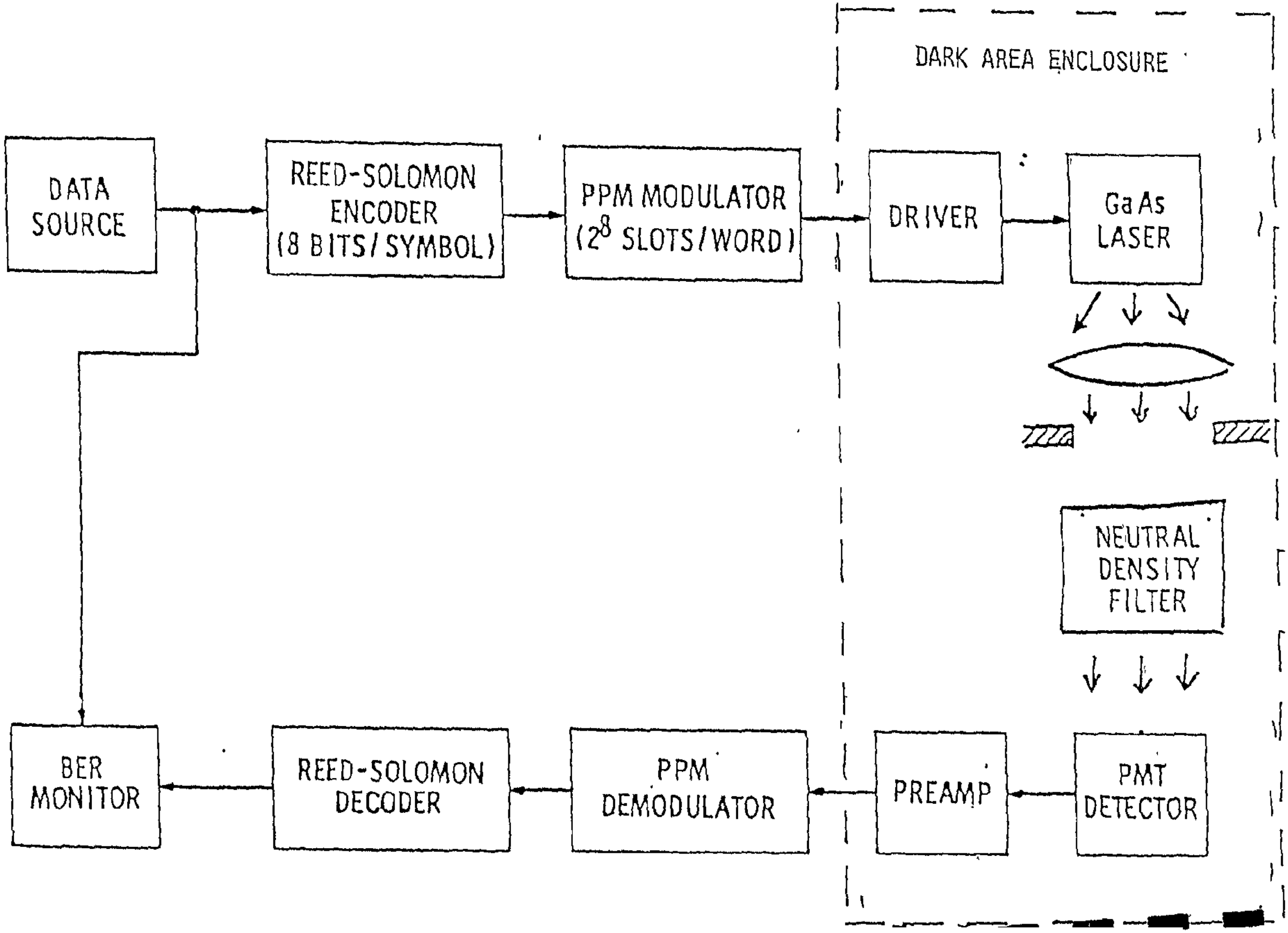
Nd:YAG RISK ASSESSMENT ACTIVITY

- MDAC-STL
  (1970—1979)

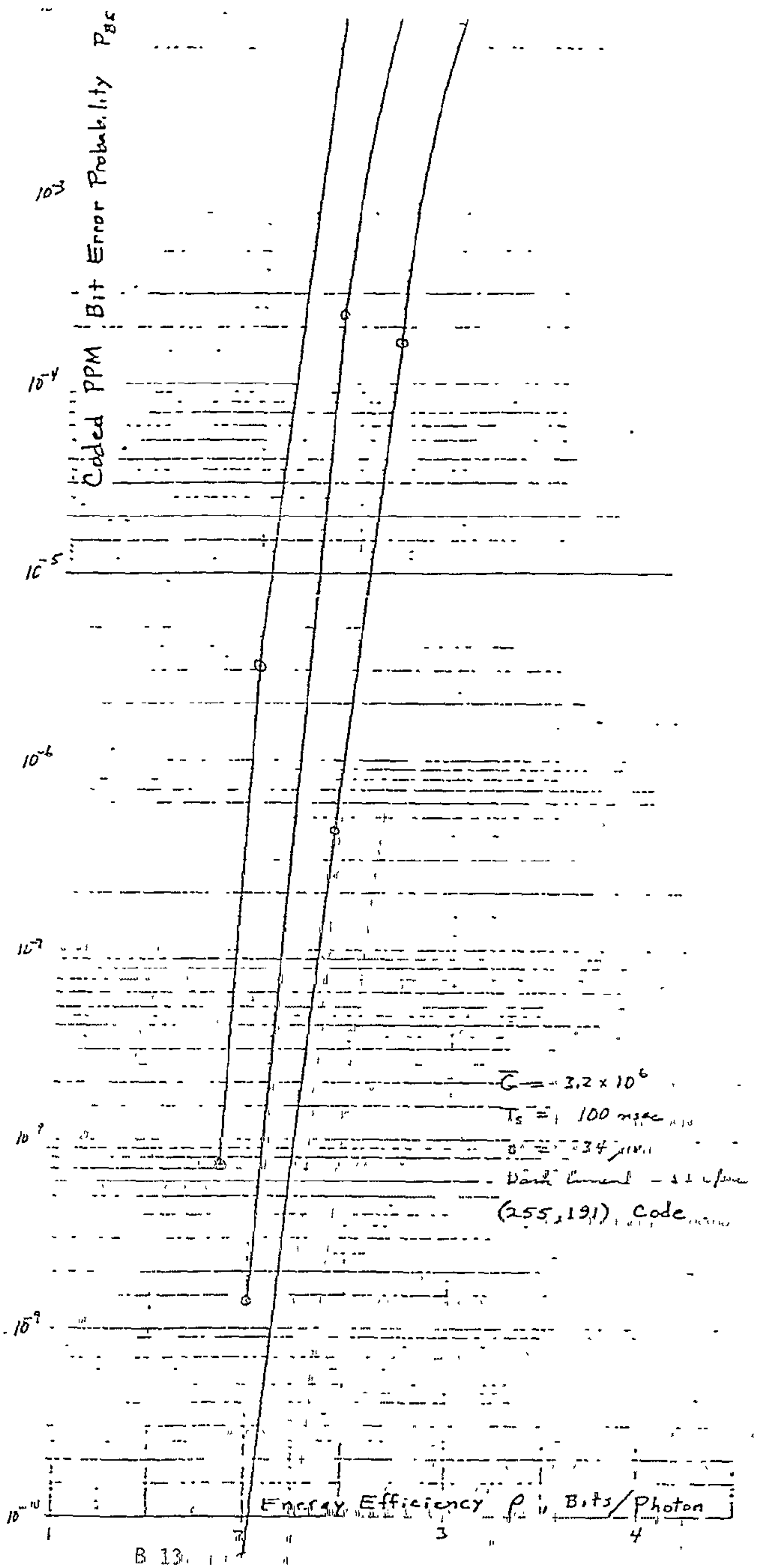
- TECHNOLOGY ASSESSMENT TEAM — JUNE 1980
  - USAF
  - AEROSPACE CORP.
  - LINCOLN LAB.

- ASSESSMENT BRIEFINGS
  (JUNE 1980)
  - GEN. WARD
  - HQ/AF/RDQ

- DSP ASSESSMENT MILESTONES
  - LASERCOM SELECTED OVER RF
  - Nd:YAG SELECTED OVER PULSED GaAs
  - MDAC-STL PROPOSED FIXED-PRICE


DARK AREA ENCLOSURE
DATA SOURCE
REED-SOLOMON ENCODER (8 BITS/SYMBOL)
PPM MODULATOR (2^8 SLOTS/WORD)
DRIVER
Ga As LASER
NEUTRAL DENSITY FILTER
PMT DETECTOR
PREAMP
PPM DEMODULATOR
REED-SOLOMON DECODER
BER MONITOR

ORIGINAL PAGE IS
OF POOR QUALITY



ADVANTAGES OF LASER SYSTEM

ADVANTAGES	REASON
LOW RISK	• SOLID STATE SIMPLE SYSTEM NO NEW TECHNOLOGY • FLY IN '83 (P80-1) AND '8X (DSP)
MINIMUM TIMING ERRORS	• >1 GHz BANDWIDTH
SYSTEM SIMPLICITY	• NO TUNED CIRCUITRY • DIRECT DETECTION RECEIVER
MINIMUM SPACE STATION IMPACT	• LOW SIZE, WEIGHT AND POWER • VERY SIMPLE INTERFACES
MAXIMUM ORBIT FLEXIBILITY	• FULL HEMISPHERE COVERAGE
HIGH COMMUNICATION CAPACITY POSSIBLE	• ACOMMODATES LOW TO HIGH DATA RATES

ADVANTAGES OF OPTICAL SYSTEM (Continued)

ADVANTAGES	REASON
PROVIDES MA TECHNIQUE WHICH SOLVES NEAR-FAR PROBLEM	• ONE TELESCOPE PER USER
SIMPLIFIES ANTENNA SYSTEM AND ALLEVIATES RFI INTERFERENCE	• NATURE OF LASER BEAM • SMALL GIMBALLED APERTURE (4" TO 6")
IMMUNE TO JAMMING AND RFI INTERFERENCE	• NO KNOWN THREAT • LOW SIDELOBES • REQUIRES CO-ORBITAL JAMMER

WHAT IS CATSS?

- CATSS IS A COMPUTER SIMULATION TOOL USED TO SUPPORT C & T SYSTEM ARCHITECTURE DEVELOPMENT AND IS USEFUL DURING ALL PHASES OF THE SS PROJECT FROM CONCEPTIONAL DESIGN TO AND DURING MISSION OPERATION

- THERE ARE TWO PARALLEL ON GOING COMPONENTS OF CATSS:
  - C & T WITHOUT GEOMETRY OF PROBLEM
  - GEOMETRIC EFFECTS ON C & T

THE TWO COMPONENTS WILL BE INTEGRATED EVENTUALLY FOR MISSION OPERATION SUPPORT

APPENDIX C

LINK BUDGET AND MINIMUM DETECTABLE SIGNAL FOR NONCOHERENT OPTICAL COMMUNICATION SYSTEMS

1. Introduction

Optical link has been proposed [1] as a possible alternative to microwave technology for:

(1) Intersatellite links,

(2) Space Station-Free-Flyer and/or EVA links, and

(3) On board communication and data transfer.

The motive of an optical C&T system is of course due to its multifold advantages over conventional methods: reasonable antenna size, immunity to RFI and jamming, large bandwidth, to mention a few. However, optical technology is not without its own drawbacks, for example the acquisition and pointing problems between moving platforms [6].

In this report, we shall concentrate our attention on direct detection (DD) systems. Heterodyne or coherent systems can, in principle, have a sensitivity much higher than that of DD systems. However,this latter technology is by no means mature at this moment. Some critical technology and system issues, notably frequency and phase locking, were discussed by Chan et al. [7] recently.

2. System Block Diagram and Link Budget

2.1 Digital Optical Link Model

A typical digital optical transmission system is illustrated in the block diagram of Fig. 1. The data source generates a sequence of binary symbols at rate 1/T. The encoded baseband waveforms are then used to modulate an optical source. The modulation can be either in the form of

FM, PM, AM like those implemented in the microwave range, of intensity modulation (IM), polarization modulation (PLM), hybrid (IM-PLM), which are more popular in the optical band [2]. The received light waveforms, weakened and distorted by propagation, are converted into electrical signals by photodetector, amplified to increase the power level of the detector output. The optical detectors are generally categorized into two basic types: direct (power) and heterodyne detection, depending on the modulation format of the incoming waveforms. The decision device, in general, consists of postamplifier, equalizer and filter, before a threshold comparison is made.

2.2 Typical Link Budget

The range equation for such a link is given by [9]

$$P_R = P_T L_T G_T \eta_T L_P \left(\frac{\lambda}{4\pi R}\right)^2 T_A G_R L_R \eta_R \qquad (1)$$

where

P_R: Average received light power

P_T: Average transmitting laser power

L_T: Transmitter optical transmission loss

G_T: Transmitter telescope gain

η_T: Transmitter antenna efficiency

L_P: Transmitter pointing loss

$(\lambda/4\pi R)^2$: Free space loss $\triangleq L_F$

R: Distance between transmitter and receiver

T_A: Atmospheric transmission loss

G_R: On-axis receiving antenna gain

L_R: Receiver optics transmission loss

η_R: The receiver antenna efficiency and

$$G_R = \left(\frac{\pi D}{\lambda}\right)^2 \quad (2)$$

D: Receiving telescope clear aperture diameter

$$G_T = 2\left(\frac{2\pi R_0}{\lambda}\right)^2 \left[\exp\left(\frac{-b^2}{R_0^2}\right) - \exp\left(\frac{-a^2}{R_0^2}\right)\right]^2 \quad (3)$$

for a Cassegrainian Telescope [4] with $R_0 \triangleq 1/e^2$ power point of the feed beam at the exit aperture of diameter D, where $D \geq 3R_0$, a = radius of the primary mirror and b = the radius of the secondary mirror.

Typical budget for the Space Station-Free Flyer and Space Station-Relay Satellite links are shown in Table 1. Itemized losses for L_T and L_R are tabulated in Table 2.

2.3 Minimum Detectable Signal

Also shown in Table 1 is the average converted power (I_{ph}^2) pre-amplifier thermal noise power (N_0). Note that even if there is no background noise, shot noise, dark-current noise, zero extinction ratio (perfect modulation) and high input impedance ($10^4\ \Omega$), the thermal noise alone can, in many cases, make the correct detection almost impossible.

Thus, it is clear that some mechanism of photomultiplication is needed in most applications. The effect of photomultiplication can be easily seen by the resulting signal-to-noise ratio (SNR) defined as [8]

$$SNR = \frac{\frac{1}{2}(mI_{ph})^2|G(\omega)|^2}{[2q(I_{ph}+I_B+I_D)|G(\omega)|^2F(G) + 4K\frac{T_{eff}}{R_{in}}]B}, \quad (4)$$

where

$G(\omega)$ = Mean current gain or multiplication factor of the photomultiplier

B = Bandwidth of the photodetector

I_B = Background induced photocurrent

I_D = Multiplied dark current

m = Modulation index of the light

R_{in} = Input impedance of the pre-amplify circuit

T_{eff} = Effective temperature

$2q(I_{pn}+I_B+I_D)B$ = Mean-square shot-noise current

and

$F(G)$ = the excess noise (see eq. (7))

More elaborate analysis of these noise components, taking intersymbol interference into consideration, can be found elsewhere [3,5]. For our present purpose, the above expression (4) is clear enough, since typically, $50 < G < 300$ for optimal avalanche photodiode (APD) gain and $G \approx 10^6$ for photomultiplier tubes (PMT) [10]. From (4), the minimum detectable signal for the quantum noise limited (QNL) case, i.e. $I_B = I_D = 0 = T_{eff}$, and the thermal noise limited (TNL) case, i.e., the first term in the denominator of (4) is 0, are derived as follows.

$$P_{QNL} = 4h\nu \frac{B}{m} (SNR) \frac{F(G)}{\eta} \tag{5}$$

$$P_{TNL} = \frac{h\nu}{q} [B(SNR)]^{1/2} \left(\frac{8kT_{eff}}{\eta^2 R}\right)^{1/2} \tag{6}$$

where

$$F(G) = G\left[1-(1-k)\left(\frac{G-1}{G}\right)^2\right] \quad \text{for APD}$$
$$\approx 1 \quad \text{for PMT} \tag{7}$$

and k is the ratio of the ionization coefficients of the holes and electrons in the APD junction.

2.4 Application of CATSS to Link Budget Analysis

Fig. 2 to Fig. 4 demonstrate the variations of the received power as a function of distance (R). The first curve use the following parameters for a low data rate (5 Kb/sec) link:

P_T (dBW) = 12.5

L_T (dBW) = -2.0

G_T (dBW) = 70.0

λ (μm) = 0.875

D (cm) = 6.5

L_R (dBW) = -4.10

The second set of link budget (Fig. 3) is a typical median data rate (1 Mb/sec) link:

P_T = -12.0

L_T = -5.1

G_T = 81.3

λ = 0.825

D = 20.3

L_R = -8.5

The high data rate (250 Mb/sec) link (Fig. 4) has the following parameters:

P_T = -10.0

L_T = -5.0

G_T = 100

λ = 0.825

D = 20.3

L_R = -8.5

REFERENCES

[1] NASA, "Space Operation Center, Communication and Tracking System: Preliminary Conceptual Design," EE8-4/82-47, JSC, Houston, Apr. 1982.

[2] Gagliardi, R. M., and Karp, K., Optical Communication, John Wiley and Sons, New York, 1975.

[3] Personick, S., Balaban, P., Bobsin, J., Kumar, P., "A Detailed Comparison of Four Approaches to the Calculation of the Sensitivity of Optical Fiber System Receivers," IEEE Trans. on Commun., Vol. 25, pp. 541-549, May 1977.

[4] Klein, B. J., and Degman, J. J., "Optical Antenna Gain I: Transmitting Antennas," Appl. Opt., Vol. 13, No. 9, Sept. 1974.

[5] Dogliotti, R., Luvison, A., and Pirani, G., "Error Probability in Optical Fiber Transmission Systems," IEEE Trans. on Info. Theory, Vol. 25, pp. 170-179, March 1979.

[6] Van Hove, P. and Chan, V. W. S., "Spatial Acquisition Algorithms and Systems for Optical ISL," ICC'83, Boston, MA, June 1983.

[7] Chan, V. W. S., Jeromin, L. L. and Kaufmann, J. E., "Heterodyne Lasercom Systems Using GaAs Lasers for ISL Applications," ICC 83, Boston, MA, June 1983.

[8] Melchior, H., Fisher, M. B., and Arams, F. R., "Photodetectors for Optical Communication Systems," Proc. IEEE, pp. 1466-1486, Oct. 1970.

[9] Ross, M., Freedman, P., Abernathy, J., Matassov, G., Wolf, J. and Barry, J. D., "Space Optical Communications with the Nd: Yag Laser," Proc. IEEE, Vol. 66, pp. 319-344, March 1978.

[10] Leverenz, D. J. and Gaddy, O. L., "Subnanosecond Gating Properties of the Dynamic Cross-Field Photomultiplier," Proc. IEEE, Vol. 58, pp. 1487-1490, Oct. 1970.

Table 1. Typical Link Budget.

LINK PARAMETER	SS-FREE FLYER*1	SS-RELAY SATELLITE*2
TRANSMITTER		
LASER POWER (AVG.)	-12.22 (dBW)(60 mW)	-12.22
OPTICAL TRANSMISSION	-6.56(dBW)	-6.56
TELESCOPE GAIN	72.93 (80) (dB)	72.93 (80)
TELESCOPE EFFICIENCY	0 (dB)	0
POINTING LOSS	0 (dB)	0
FREE SPACE LOSS	-241.67 (dB)	-261.95
ATMOSPHERIC TRANSMISSION	0 (dB)	0
RECEIVER		
ANTENNA GAIN (dB)	113.43*3 (127.47)*4	113.47 (127.47)
OPTICS TRANSMISSION	-8.45 (dB)	-8.45
ANTENNA EFFICIENCY	0 (dB)	0
RECEIVED SIGNAL (dBW)	-82.54 (-61.42)	-102.84 (-81.7)
CONVERTED RECEIVED POWER (AVE.)*5 (dBW)	-168.34 (126.12)	-208.94 (-166.66)
PRE-AMP. THERMAL NOISE POWER *6	-134.80 (-157.81)	-134.80 (-157.81)

*1 $R = 2.0 \times 10^3$ Km

*2 $R = 3.8 \times 10^4$ Km

*3 $\lambda = 851$ nm

$D = 127$nm

*4 $\lambda = 532$ nm

$D = 40$ cm

*5

$$(I_{ph})^2 = \eta \frac{qP_R}{h\nu}^2$$

*6

$$N_0 = (4KT_{eff}/R_{in})B$$

Here we assume $T_{eff} = 300°K$,

$B = 100$ MHz

$R_{in} = 50\ \Omega\ (10^4 \Omega)$

η: quantum efficiency for the detector (Assume = 1 here)

h: Planck's constant

q: 1.6×10^{-19}

ν: optical carrier frequency

Table 2. Itemized Optics Transmission.

FIBER COLL EFF	0.40
FRESNEL LOSS	0.85
COUPLING LOSS	0.80
INTEGRATOR LOSS	0.95
LENS COLL EFF	0.88
LENS TRANSMISSION	0.97
TRANSMITTER OPT. TRANS.	0.221 (-6.56 dB)
CONDUCTIVE COAT	0.86
NARROW BAND FILTER	0.86
PRIMARY	0.80
SECONDARY	--
TRIPLET	--
POWER SPLIT	0.25
PYRAMID	--
TRIPLET	--
WINDOW	0.97
RECEIVER OPT. TRANS.	0.143 (-8.45 dB)

Fig. 1. A Typical Digital Optical Link.

LinCom

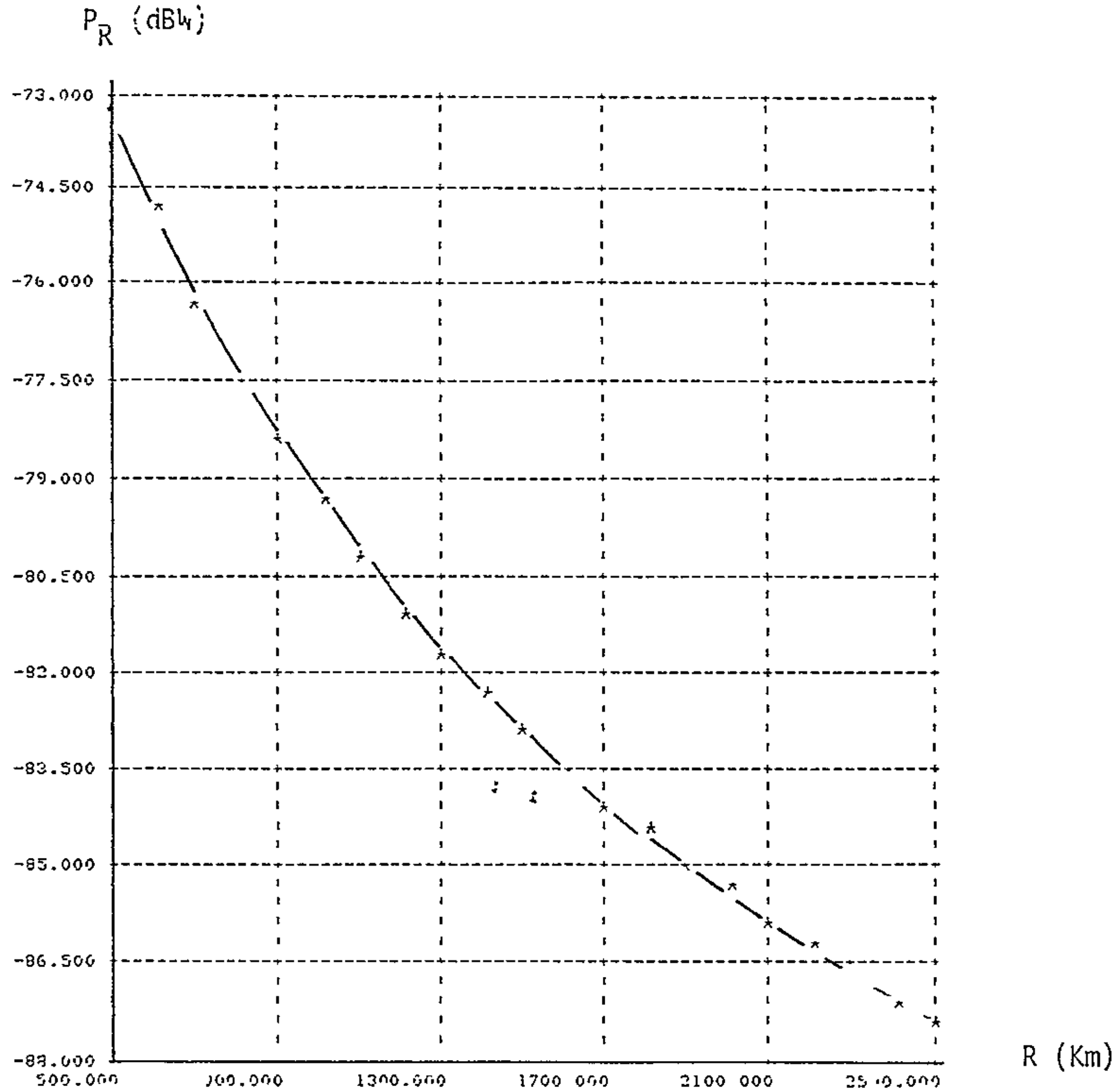


Fig. 2. Received Optical Power for Low Data Rate Link.

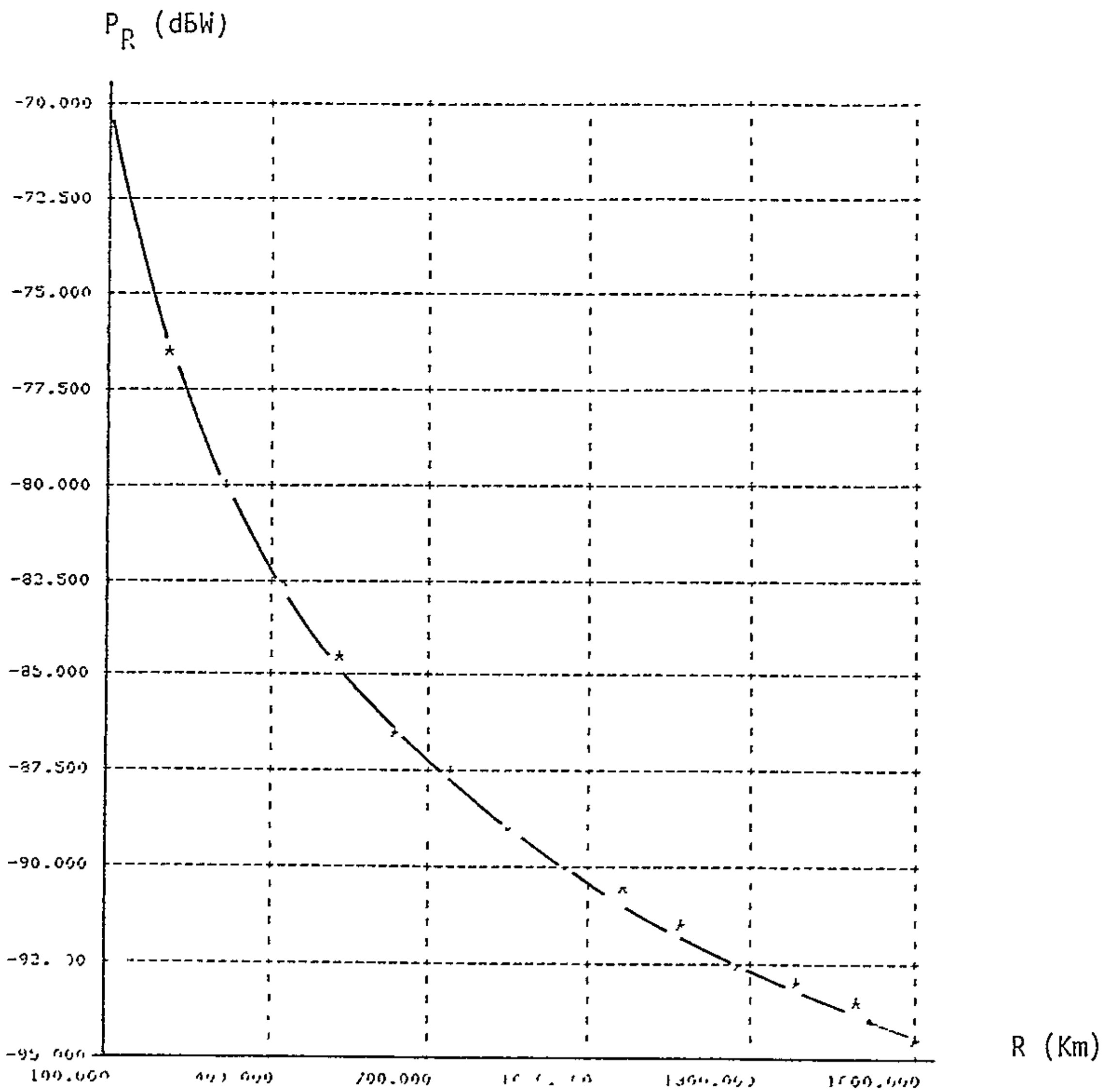


Fig. 3. Received Optical Power for a Median Data Rate Link.

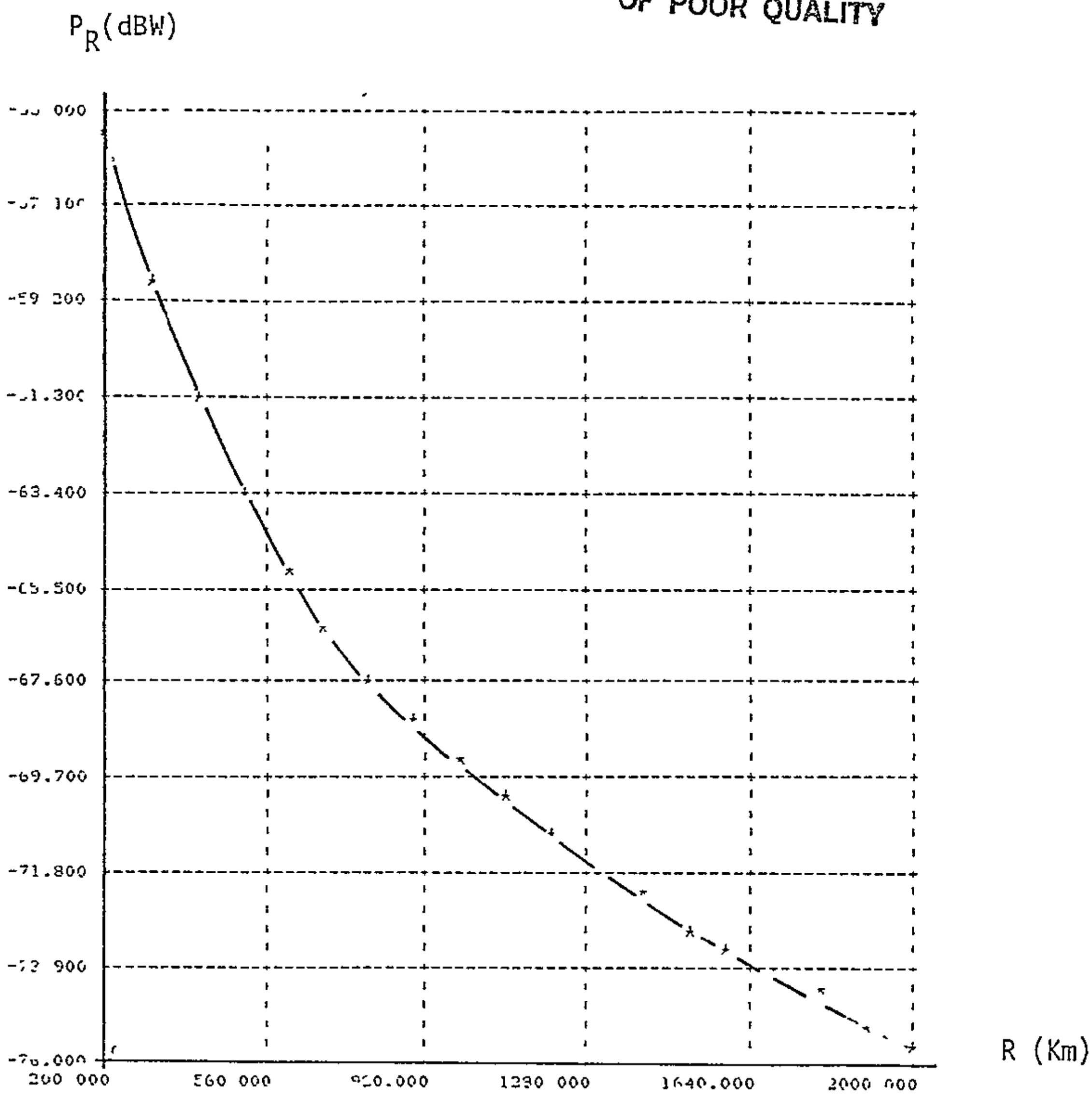


Fig. 4. Received Optical Power for a High Data Rate Link.

APPENDIX D

PERFORMANCE CONSIDERATION FOR UNCODED NONCOHERENT OPTICAL LINK

INTRODUCTION

In general, exact performance evaluations for direct detection optical links are difficult and very time-consuming. A variety of techniques have been proposed to deal with many different situations in photodetections [1,2]. Gaussian approximation (GA) is used to give quick and, in many cases, accurate enough estimations when APD is used. However, because of the skewness of avalanche gain distribution, it is not always satisfactory [1]. If computing time is not very restrictive, the Chernoff bound method [1] would be more preferable.

2. PERFORMANCE OF APD RECEIVERS

2.1 NRZ Signaling

Using Gaussian approximation, Personick [4] obtained the following equation for the required average power $\overline{P}$, where the designed error rate is P_e and NRZ format is used.

$$\eta\overline{P} = \left(\frac{h\nu}{q}\right) Q\left[\frac{\langle i\rangle_c^{1/2}}{G} + qBI_1QF(G)\right] \tag{1}$$

where

$Q \quad = \mathrm{erfc}^{-1}(P_e)$

$\mathrm{erfc}^{-1}(\cdot)$ = inverse complement error function

$\langle i^2\rangle_c$ = the mean square noise current in the absence of signal

$I_1 \quad \sim 0.5$ for rectangular pulse shape

Since G and F(G) are related by [6, Eq.7], it is obvious from (1) that there exist an optimal G which minimize $\overline{P}$ for a given P_e. This optimal G is found to be

$$G_{opt} = \frac{1}{k^{1/2}} \left[\frac{\langle i^2\rangle_c^{1/2}}{qBI_1Q} + k - 1\right]^{1/2} \qquad (2)$$

This equation is valid only if the terms inside the bracket is positive, i.e., $\langle i^2\rangle_c^{1/2}$ cannot be too small.

Substituting (2) into (1) one gets

$$\eta\overline{P} = 2h\nu BI_1Q^2(kG_{opt}+1-k) \qquad (3)$$

Note that (3) also yields an optimal $\hat{G}_{opt}$, the G that maximizes SNR. This $\hat{G}_{opt}$ is, however, not in agreement with G_{opt} in most cases, due to the fact that P_e is not a monotonic function of SNR even under the Gaussian approximation. The comparison between $\hat{G}_{opt}$ and the optimal gain which minimizes P_e, using more accurate computation (other than GA) was made by Gagliardi and Prati [7].

Applying [6, Eq. 1] and (3), we can then find the required transmitting antenna gain as a function of error rate.

In the above analysis, zero extinction ratio, $\gamma = 0$, is assumed. When $\gamma \neq 0$ (3) becomes

$$\eta\overline{P} = \left(\frac{h\nu}{q}\right)\left(\frac{1+\gamma}{1-\gamma}\right)\Big\{(1+r)\frac{Q^2qBI_1F(G)}{1} + \left[4\gamma\left(\frac{Q^2qBI_1F(M)}{1-\gamma}\right)^2 + \frac{Q^2\langle i^2\rangle_c}{G^2}\right]^{1/2}\Big\} \qquad (4)$$

2.2 Bi-phase Signaling

One disadvantage of the NRZ code is that its performance depends on the threshold and gain which are in turn functions of the signal and noise levels.

One way to get rid of threshold setting problem is to use Manchester format, or binary PPM. For this case the required detected optical power is given by Muoi [3]

$$\eta\overline{P} = Q\,\frac{h\nu}{q}\,\frac{1+r}{1-r}\,\{qQBI_1'\,\frac{1+r}{1-r}\,F(G) + \sqrt{\{QqBI_1'\,\frac{1+r}{1-r}\,F(G)\}^2 + \frac{\langle i^2\rangle_c}{G^2}}\}$$

where

$$I_1' \sim 0.5888. \tag{5}$$

On the other hand, the error probabilities for both NRZ and bi-phase modulation can be expressed in terms of received power P_r:

$$P_e = \mathrm{erfc}(Q) \tag{6}$$

where

$$Q = \frac{2\eta^2 qG(1-r)P_r}{h\nu(1+r)}\,[\sqrt{\langle i^2\rangle_c + \frac{4BI_1(G\eta q)^2F(G)P_r}{h\nu(1+r)}}$$

$$+ \sqrt{\langle i^2\rangle_c + \frac{4BI_1(G\eta q)^2F(G)P_r}{h\nu(1+1/r)}}\;]^{-1} \quad \text{for NRZ} \tag{7a}$$

and

$$Q = \frac{\eta^2 q(1-r)GP_r}{h\nu(1+r)}\,[\langle i^2\rangle_c + 2(\eta qG)^2F(G)BI_1'P_r/h\nu]^{-1/2} \quad \text{for Bi-phase} \tag{7b}$$

Numerical results using the Gaussian approximation show that the PPM is less sensitive than the NRZ coding by about 1 dB at G_{opt}, 3 dB at high G and 0 dB at low G. However, experiment as well as more accurate computations [3] indicate that the PPM receiver is more sensitive than the NRZ receiver by about 1.8 dB at high G and 1.5 dB at G_{opt} (r=0.1). The reasons for these discrepancies are discussed in [1,3]. In summary, the Gaussian approximation is very accurate in predicting the receiver sensitivity when APD and NRZ are applied. On the other hand GA underestimate the BPPM (Manchester) receiver sensitivity by about 1.8 dB ($P_e = 10^{-5}$), overestimate the G_{opt} and underestimate the threshold (specific numerical results shown for $P_e = 10^{-9}$ can be found in [3]).

Muoi also compared the tradeoffs between Manchester and NRZ systems. We summarize these* pro and con, besides the sensitivity issue mentioned before, as follows.

1. Laser pattern dependency which can degrade receiver sensitivity is virtually nonexistent in the Manchester format. This also simplifies the laser power output stabilization circuitry.
2. Baseline wander is not a problem since the Manchester data have zero dc content. This should help to alleviate the dynamic range problem** at the receiver (particularly when a high impedance amplifier design is used**).
3. It avoids the use of scrambling and descrambling as often used in the NRZ* case thus making timing extraction easier.
4. Since the threshold offset variation is not a problem a large

*Some of them can be applied to the PMT case also.

**See [9] for discussions on these problems.

variation in the peak signal output can be tolerated and the requirements on the AGC design are less severe.

5. The Manchester data have very small low frequency energy. Thus, any possible pole-zero mismatch** in equalizing the high impedance receiver does not degrade the receiver sensitivity.
6. The Manchester format effectively doubles the transmitting pulse rate.

3. Performance of PMT Receiver

The photocurrent distribution at the output of Photomultiplier Tube (PMT) for a space-space link is usually assumed to be Poisson or conditional Poisson. Under Poisson counting assumption, the NRZ system has an error rate [5]

$$P_e = \frac{1}{2}\left[1 + \frac{\Gamma(K_T,K_s) - \Gamma(K_T,K_s+K_b)}{\Gamma(K_T,\infty)}\right] \tag{8}$$

where

$$\Gamma(n,x) = \int_0^x e^{-t}t^{n-1}dt \,, \tag{9}$$

K_T is the threshold used,

K_s an K_b are the average signal and noise counts over a bit time.

If bi-phase modulation is employed, the bit error rate becomes

$$P_e = Q(\sqrt{2m_0},\sqrt{2m_1}) - \frac{1}{2}\exp[-(m_1+m_0)]I_0(2\sqrt{m_1m_0})] \tag{10}$$

where

$m_1 = K_s + K_b$

$m_0 = K_b$

$$Q(a,b) = \int_b^\infty x \exp\left[-\frac{(a^2+x^2)}{2}\right] I_0(ax)dx \qquad (11)$$

A useful bound for (10) can be obtained by expanding the Q function in terms of Bessel function series and upper-bounding $I_n(x)$ by $I_0(x)$ to yield

$$P_e < \left(\frac{\sqrt{m_0}}{\sqrt{m_1}-\sqrt{m_0}} + \frac{1}{2}\right) \exp[-(m_1+m_0)] I_0(2\sqrt{m_0 m_1}) \qquad (12)$$

Note that P_e of NRZ is a function of the threshold K_T. Though the optimal K_T is not easy to come by as in the AWGN channel case, a pair of bounds can be derived as follows. Rewriting (8) as [5]

$$P_e(K_T) = \frac{1}{2}\sum_{k=0}^{K_T} \gamma_{kK_T} \frac{(K_0+K_s)^k}{k!} e^{-(K_b+K_s)} + \frac{1}{2}\sum_{k=K_T}^{\infty} \gamma_{kK_T} \left(\frac{K_b^k}{k!}\right) e^{-K_b} \qquad (13)$$

where

$$\gamma_{kK_T} = \begin{cases} 1/2 & \text{if } k = K_T \\ 1 & \text{otherwise} \end{cases} \qquad (14)$$

Then the optimal K_T must satisfy the following two inequalities, K_T being restricted to integers,

$$P_e(K_T) < P_e(K_T+1)$$

$$P_e(K_T) < P_e(K_T-1) \,.$$

It can be shown that the above two inequalities are equivalent to

$$e^{-K_s}\left(1 + \frac{K_s}{K_b}\right)^{K_T} \left(\frac{1+K_b+K_s+K_T}{1+K_b+K_T}\right) > 1 \qquad (15a)$$

and

$$e^{-K_s}\left(1+\frac{K_s}{K_b}\right)^{K_T-1}\left(\frac{K_b+K_s+K_T}{K_b+K_T}\right) < 1. \qquad (15b)$$

A closed form solution of (15a) and (15b) eludes all out exertions. However an efficient algorithm in locating the optimal K_T can be found as follows. First, we observe that (15b) can be rewritten as

$$e^{-K_s}\left(1+\frac{K_s}{K_b}\right)^{K_T-1} < \frac{K_b+K_T}{K_b+K_T+K_s} \qquad (16)$$

A reasonable range of the optimal K_T is

$$K_b < K_T < K_s + K_b.$$

By inserting the above constraint into (16), we obtain

$$\left(1+\frac{K_s}{K_b}\right)^{K_T-1} < e^{K_s}\left(1-\frac{K_s}{2(K_s+K_b)}\right)$$

or,

$$K_T < T\left\{\frac{K_s+\ln\left(1-\frac{K_s}{2(K_s+K_b)}\right)}{\ln(1+K_s/K_b)}+1\right\} \triangleq K'_{u_1} \qquad (17)$$

where $\quad T\{x\} \triangleq$ the integer part of x.

Similarly, from (15a) we obtain

$$K_T > T\left\{\frac{K_s+\ln\left(1-\frac{K_s}{1+K_s+2K_b}\right)}{\ln(1+K_s/K_b)}+1\right\} \triangleq K'_L \qquad (18)$$

Letting

$$K_{u_1} = \min(K'_{u_1}, K_s+K_b) \tag{19a}$$

$$K_{L_1} = \max(K_b, K'_{L_1}) \tag{19b}$$

then

$$K_{L_1} < K_T < K_{u_1} .$$

Hence, the algorithm follows the procedure below.

(1) Compute K_{L_1}, K_{u_1} by (17), (18), (19a) and (19b).

(2) If $|K_{L_1}-K_{u_1}| \geq 1$ then substituting K_T by K_{u_1} (K_{u_n}) and K_{L_1} (K_{L_n}) respectively in (15), obtaining two new bounds by (19a) and (19b).

(3) Repeat (2) until $|K_{L_n}-K_{u_n}| < 1$.

It turns out that in most cases the convergence rate is very fast, usually $n \leq 3$.

REFERENCES

[1] Personick, S. et al., "A Detailed Comparison of Four Approaches to the Calculation of the Sensitivity of Optical Fiber System Receivers," IEEE Trans. Comm., Vol. COM-25, pp. 541-549, May 1977.

[2] Dogliotti, R. et al., "Error Probability in Optical Fiber Transmission Systems," IEEE Trans. Inform. Theory, Vol. IT-25, pp. 170-179, March 1979.

[3] Muoi, T. V., "Receiver Design for Digital Fiber Optic Communication Systems Using Manchester (Biphase) Coding," IEEE Trans. Commun., Vol. COM-31, pp. 608-619, May 1983.

[4] Personic, S. D., Optical Fiber Transmission Systems, Plenum Press, New York, 1981.

[5] Gagliardi, R. M., and Karp, S., Optical Communication, John Wiley and Sons, New York, 1975.

[6] LinCom Corp., "Shuttle Communications and Tracking Systems Modeling and TDRSS Link Simulations," Monthly Progress Report, June 1984.

[7] Gagliardi, R. M., and Prati, G., "On Gaussian Error Probabilities in Optical Receivers," IEEE Trans. Commun., Vol. COM-28, pp. 1742-1747, Sept., 1980.

APPENDIX E

A SIMPLE COARSE RANGING SYSTEM FOR SPACE STATION APPLICATIONS

Summary

A simple ranging system which can provide a raw range accuracy of about 500 m can be designed using the information provided by the frame sync in the data structure. With additional smoothing, the accuracy should improve to about 200 m. This technique does not impact on the modulation and hardware design and is implemented through the data channel. If the accuracy is tolerable, this technique is much simpler than using PN ranging.

Frame Sync Ranging System

Figure 1 summarizes the principle of the frame sync ranging technique. The range difference can be determined from $\tau_D = \frac{1}{2}(\tau_s - \tau_u)$. The user measured time difference τ_u must be relayed back to the SS for the required computation for range determination.

For the MA system, assume a lowest data rate of 48 Kbps. Assuming that the bit sync can track to an accuracy of 5% of the bit rate, the time accuracy of the τ_u and τ_s measurement is on the order of $0.05/(48\times10^3) = 1\ \mu s \sim 313$ m. This is the raw accuracy of the system. If additional filtering is provided, this accuracy can be improved 2 to 3 times, so that an accuracy on the order of 100 m can be achieved.

In reality, one must account for the differential delays of the user and SS equipment, which must be calibrated out. The other source of error is due to the relative motion between the SS and the user over a time interval comparable to τ_U. If the motion is slow (< 10 Km/sec) this is negligible. If the motion is not slow but linear, it can be filtered out.

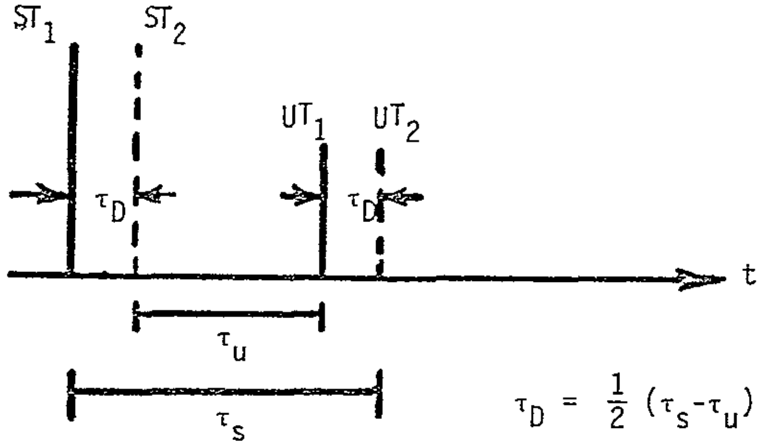
Design Consideration and Accuracy Improvement

The frame period must be selected so that it can resolve an ambiguity of 2000 Km, or 7 msec. If a 1024 frame is selected, then at 48 Kbps, the frame period is 21 msec. At 100 Kbps, the frame period is 10 msec. In either case, the 1024 frame period is adequate.

The measurement accuracy can be improved with coding so that the bit syncs work on the symbol stream rather than the bit stream. If rate 1/2 convolutional codes are used, then the accuracy improves by a factor of two. However, this must be traded against the bandwidth expansion required by coding.

Hardware Implementation

Figure 2 shows a simple way to implement this Frame Sync ranging system. The 10 MHz counter takes the two frame syncs and counts the number of pulses generated by a 10 MHz quartz crystal. The quartz crystal does not have to be very accurate (10^{-7} stability should be adequate).



ST_1 = SS Frame Sync Epoch transmitted at SS

ST_2 = SS Frame Sync Epoch received at User

UT_1 = User Frame Sync Epoch transmitted at User

UT_2 = User Frame Sync Epoch received at SS

τ_D = Propagation delay due to SS to User range

τ_s = SS measured time difference of transmitted SS Frame Sync Epoch and received User Frame Synch Epoch

τ_u = User measured time difference of received SS Frame Sync Epoch and transmitted User Frame Synch Epoch

Figure 1. Principle of Operation.

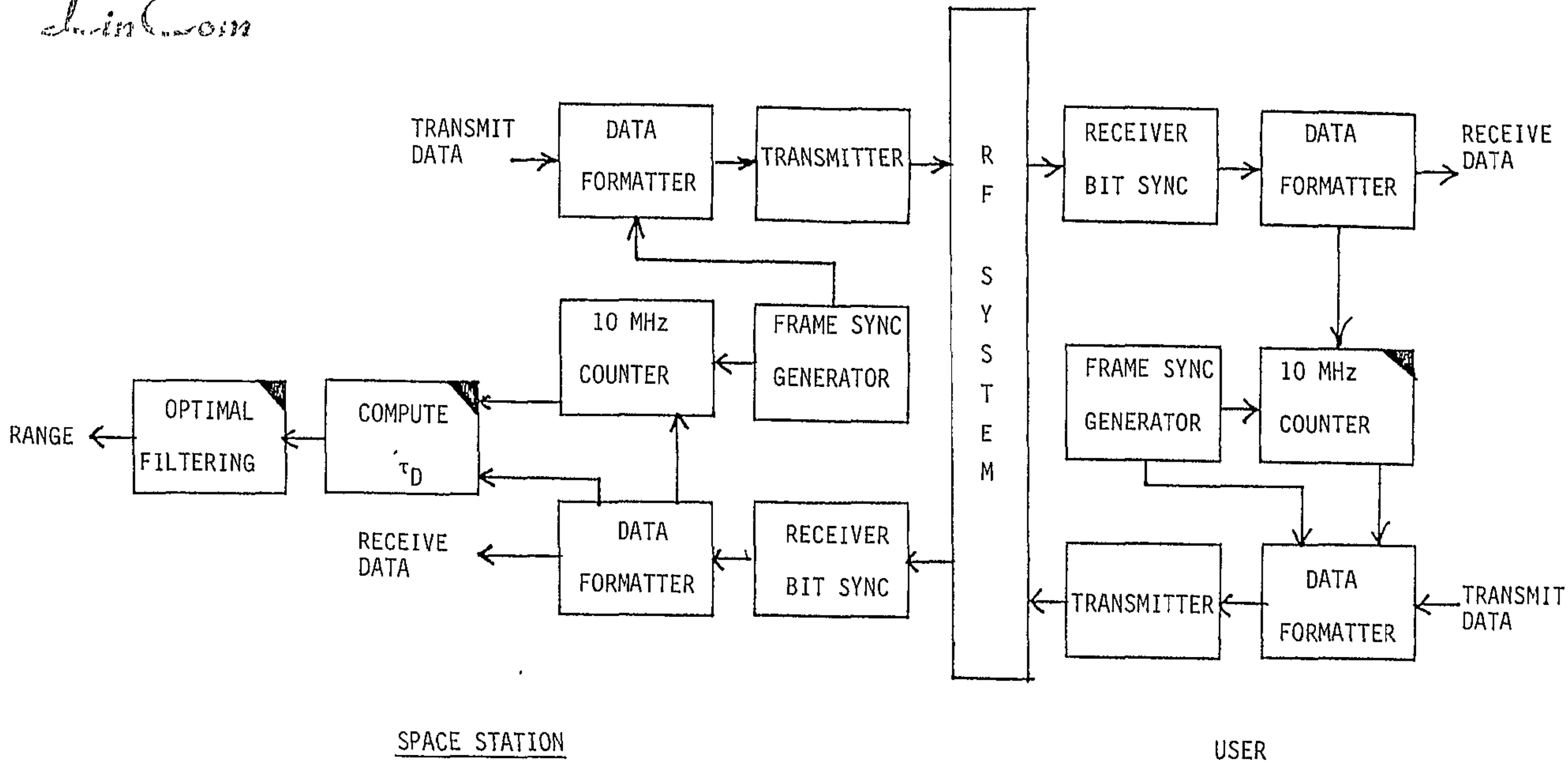


Figure 2. Hardware Implementation. [ ] Denotes Additional Circuitry.

APPENDIX F

A FIRST CUT TO SIZE THE VARIOUS MA LINKS

SUMMARY

The purpose of this memo is to point out the critical links for the Space Station MA links in terms of link margins. A simple reference K-band (15 GHz) system using 0 db antenna gains and 0 dbw transmit power is assumed. It is also assumed that the system is uncoded, the system RF loss is 3.5 db, and the receiver noise figure is 5 db.

It appears that the EVA links are in good shape - a margin of 7 db is available in the worst case. Therefore, one can trade hardware complexity with performance.

The low rate command and telemetry links for proximity operation have a healthy 11 db margin. However, the return TV links have negative margins; viz., 8 and 12 db respectively. It appears that the user must provide additional transmit power over 0 dbw as well as some directivity with its antenna to use the TV link.
Some form of coding should be employed to provide additional margin.

The free flyer links have negative margins from 2 to 23 db depending on the range separation. If a policy of providing a link at the closest range (185 Km) is adopted for 0 db user antenna gain and 0 db transmit power, then some form of coding must be employed and the data rate must be reduced to approximately 50 kbps.

REFERENCE LINK BUDGET

For the purpose of sizing the links, it is useful to define a reference link and determine the associated system margin. Any deviations from this link can be tallied and the resulting loss (or gain) can be subtracted from the reference margin to compute the particular link budget. As a reference configuration, we consider a hypothetical link between the Space Station and a user at 1 Km away, both employing isotropic gain antennas. The desired BER performance of the link is 10^{-5} so that a E_b/N_o of 9.6 db is required for an ideal additive white gaussian noise channel, in the absence of any coding gains (which is presently assumed) . For this reference link a bit rate of 1 Mbps is used. The rest of the communication system parameter are summarized by the following link budget:

	LINK PARAMETERS	VALUE	REMARKS
1	TRANSMIT POWER, DBW	0.0	1.0 WATT
2	TRANSMIT LOSS, DB	-2.0	ESTIMATE
3.	TRANSMIT ANTENNA GAIN, DB	0.0	OMNI
4	EIRP, DBW	-2.0	SUM 1 THRU 3
5	SPACE LOSS, DB	-116.0	15.0 GHZ 1 KM
6	POLARIZATION LOSS, DB	-0.5	
7	RECEIVE ANTENNA GAIN, DB	0.0	OMNI
8	RECEIVE CABLES, ETC., LOSS, DB	-1.0	ESTIMATE
9	TOTAL RECEIVE POWER, C, DBW	-119.5	SUM 4 THRU 8
10	RECEIVER NOISE FIGURE, DB	5.0	ESTIMATE
11	RECEIVE SYSTEM TEMP, DBK	29.6	NF * 290K
12	BOLTZMAN'S CONSTANT, DBW/K/HZ	-228.6	
13	NOISE SPEC. DENS., NO, DBW/HZ	-199.0	SUM 11 THRU 12
14	RECEIVED C/NO, DB-HZ	79.5	9 - 14
15	DATA RATE, DB-HZ	60.0	1.0 MBPS
16	AVAILABLE EB/NO, DB	19.5	14 - 15
17	REQUIRED EB/NO, DB	9.6	FOR 1E-5 BER
	SYSTEM MARGIN, DB	9.9	16 - 17

The system margin can be improved in a linear fashion by increasing the antenna gains, by increasing the transmit power, by reducing the implementation losses, or by reducing the data rate. The system margin can also be improved in a square law fashion,by reducing the range between the Station and the user.

SIZING THE MA LINKS

The reference link margin will now be used to determine the MA links. Unless otherwise specified the link parameters are the same as the reference link. The parameters to be varied are the Space Station antenna gain, the range and the data rate.

EVA LINKS

The EVA links have two modes of operation involving different Station antennas; namely, 0 db gain up to 300 m and 23 db gain from 300 m to 1 Km:

Antenna Gain(db)	Data Rate(Mbps)	Margin(db)
0 (300 m)	0.1	30.4
	0.275	26.0
	8	11.3
	22	6.9
23 (1 km)	0.1	42.9
	0.275	38.5
	8	23.9
	22	19.5

To arrive at these margins from the reference system margin (9.9 db), one simply add the antenna gain, subtract 20 * log (range in km) and subtract the data rate in db relative to 1 Mbps. For example, the first line in the table is computed by 9.9 + 0 + 10.5 + 10 = 30.4. The above tabulation shows that the EVA links post no major design issues since there is enough link margin.

PROXIMITY OPERATIONS

Proximity operations involve a maximum range of approximately 38.1 km. The antenna used on the Station has a gain of 23 db:

Antenna Gain(db)	Data Rate(Mbps)	Margin(db)
23	0.1	11.3
	8	-7.7
	22	-12.1

The above tabulation shows that the low rate link for proximity operations has enough system margin. However, the TV links require additional gains. These gains can be provided in the form of higher user antenna gains, higher transmit power, better receivers (lower noise figure), and coding gains.

FREE FLYER LINKS

The free flyer links operate between 185 to 2000 km. The antenna used on the Station has a gain of 23 db:

Antenna Gain(db)	Data Rate(Mbps)	Margin(db)
23 (2000Km)	0.1	-23.1
23 (185Km)	0.1	-2.4

It appears that the near range operation at 185 km is marginal whereas additional gains must be provided for the edge operation near 2000 km.

APPENDIX G

GEOMETRY OF THE MA SYSTEM FOR FREE-FLYER LINKS.

ANTENNA GAIN AND BEAMWIDTH

SUMMARY

If the SS uses a 23db gain antenna, it can be shown that the free-flyers and platforms will be within its 3 db beamwidth if the antenna is pointed parallel to the SS velocity vector. This simplifies the acquisition scenario considerable. This also gives an option for the SS to use the antenna for the order-wire channel for these users.

ANTENNA GAIN VS BEAM WIDTH

For a horn-fed paraboloidal reflector, the antenna beamwidth in degrees is approximately given by:

$$\text{beam width} = (27000/\text{gain})^{1/2}$$

For a 23db gain antenna, the beamwidth is approximately 12 degrees. The beamwidth will be wider for other types of antenna such as conformal arrays. The following is a table of beamwidth as a function of the antenna gain from 20 to 40 db:

Gain(db)	Beam width(deg)
20	16
22	13
24	10
26	8
28	7
30	5
32	4
34	3
36	3
38	2
40	2

GEOMETRY OF THE MA LINKS

The FF links extend from 185 to 2000 KM. The width of the zone is 37 KM. The angle substended by the SS to cover the zone ranges roughly from 2*ATAN(37/2/185) to 2*ATAN(37/2/2000), or 11.4 to 1.1 degree. If a 23 db gain antenna is used on the SS, the user will be within the 3 db beamwith as long as the SS antenna is pointed parallel to its velocity vector. This simplifies the acquisition procedure considerably. Unfortunately, adjacent channel interferences between users become unavoidable.

The prox ops zone translates to a large angle and the above observation does not apply.

APPENDIX H

SHORT RANGE LASER LINKS FOR SPACE STATION

This appendix provides a critique of the TRW Laser Communication Analysis results that were presented at a recent TRW/JSC space station review held at TRW. Two practical approaches to closing a 400 Mbps link for short range Space Station links are presented and compared with TRW results.

Review of TRW Analysis

TRW's analysis, Appendix A, for an open space transfer system operating at 400 Mbps was checked using the LinCom CATSS program for analytical validity as well as parametric validity. After wading through the terminology differences it is apparent that with the same inputs, LinCom's analysis provided results within a few decibels of that given here, see Figures 1 and 2 and compare with Appendix A results. A few parametric inputs were of concern, (specifically the non-multiplied and multiplied current numbers), however, this link is basically background noise limited in the worst case, and pre-amplifier noise limited in the best case, it is not of any importance to the results.

LinCom agrees with the TRW results which stated that an acquisition and tracking system will be needed to compensate for the structural movements of the space station boom. Beam steerers with swings of $\sim\pm 1°$ could easily follow this slow fluctuation in the line-of-sight, (± 5 feet at 492 feet).

LinCom Analysis and Recommendation

The basic concern between LinCom analysis and TRW's lies in the parametric values selected. With beam steerers controlling the line-of-sight uncertainties, narrower beams and small view fields could be employed. Figure 3 illustrates a recommended approach to solving the sun-in field-of-view problem. The view field is reduced by a factor of two (reducing the received background by a factor of four) and the transmit beam divergence is reduced by a factor of six, thereby increasing the transmit gain by a factor of 2.5 from the given analysis. Thus, a recommended approach to beat the sun-in field-of-view condition requires only ~0.5 mW as opposed to the ~40 mW shown in the TRW analysis. A factor of ~80 difference! Table 1 shows how the margin increases by changing the parameters. Link analysis for these recommended parameters is given in the link budget provided in Figure 4. It would therefore appear that the sun-in field-of-view condition can be beaten by selecting appropriate parameters that are reasonable and allow a lower transmit power.

Afterthought

A final and important thought remains. Why bother with the sun in the field-of-view requirement? A 1 mrad FOV (as given in the provided analysis) yields a footprint at 500 ft of 0.5 ft (or 6 inches). A sun block of greater than 6 inches diameter placed behind the transmitter totally eliminates the sun in the field-of-view phenomenon. Only the off-axis sun-scatter will be of concern and it is typically 20 dB down from the on-axis condition. The limiting link factor may very well be the pre-amp noise power.

Table 1. Recommended Parameters.

	Parameter	Change	Margin Increases By
1.	Transmit Gain	1/6 X	~15.5 dB*
2.	FOV**	1/2 X	~3.0 dB*
3.	Filter BW**	1/2.5 X	~2.0 dB*
4.	Receive Aperture**	2 X	~3.0 dB*

* All values approximate. Other factors reduce these improvements slightly.

**Background limited conditions.

Figure 1. Open Data Transfer APD Detector.

		TRW ANALYSIS	LinCom ANALYSIS USING TRW PARAMETERS
TRANSMIT POWER	43.8 mV	-13.59 dBW	-13.59 dBW
OPTICS TRANSMISSION	0.5	-3.01 dB*	-3.01 dB
POINTING LOSS	2 mrad	-3.73 dB*	-3.73 dB
TRANSMIT GAIN	6 mrad	58.69 dB*	58.69 dB
WAVEFRONT LOSS	σ = .15	-3.86 dB*	-3.86 dB
RANGE LOSS	150 M	-187.18 dB*	-187.18 DB
RECEIVER GAIN	1"	99.71 dB*	99.71 dB
RECEIVER OPTICS / FILTER TRANSMISSION	0.5	-3.01dB*	-3.01 dB
STRUT LOSS		-0.3 dB*	-0.3 dB
MARGIN		-3.00 dB	-3.00 dB
RECEIVED SIGNAL		-59.28 dBW	-59.28 dBW
REQUIRED SIGNAL		-59.43 dBW	-59.21 dBW
EXCESS SIGNAL		.15 dB	.07 dB

TRANSFER LOSS = ADDITION OF ASTERISKED ITEMS = -45.84

ORIGINAL PAGE IS
OF POOR QUALITY

LinCom

Figure 2. LinCom Analysis of 100 Mbps Link.

```
FILENAME:LINKCY1 DATE: 84/11/16

SPACE STATION FREE SPACE TRANSFER LINK
LINK PARAMETERS
   1. LASER TYPE                                           GAALAS
   2. WAVELENGTH, METERS                                   .8250E-06
   3. LINK TYPE                                            COMMUNICATION
   4. SYMBOL ERROR PROBABILITY                             .1000E-06
   5. DATA RATE, BPS                                       .4000E+09
   6. MODULATION FORMAT                                    MANCHESTER
   7. EXTINCTION RATIO                                     10.:1
                                                            FACTOR       DB
TRANSMITTER PARAMETERS
   1. PEAK POWER, WATTS                                    .4360E-01    -13.59
        AVERAGE POWER, WATTS                                .2409E-01
        PULSEWIDTH, SECONDS                                 .1250E-08
        PULSE RATE, PPS                                     .4000E+09
   2. TRANSMISSION EFFICIENCY                              .2055E+00    -6.87
        OPTICS EFFICIENCY                                   .5000E+00
        RMS WAVEFRONT ERROR LOSS                            .4110E+00
        CIRCULARIZING LOSS                                  .1000E+01
   3. POINTING LOSS                                        .4239E+00    -3.73
        POINTING BIAS ERROR, RADIANS                        .2000E-05
        POINTING JITTER ERROR, RADIANS RMS                 0.
        TOTAL POINTING ERROR BIAS+JITTER, RADIANS           .2000E-05
   4. TRANSMITTER GAIN (HAND INPUT GTYPE=6)                .8582E+06     59.34
        APERTURE DIAMETER, METERS                           .1000E-01
        OBSCURATION DIAMETER, METERS                       0.
        BEAM DIVERGENCE, RADIANS                            .6106E-05
CHANNEL PARAMETERS
   1. FREE SPACE LOSS                                      .1916E-16   -167.18
        RANGE SEPARATION, METERS                            .1500E+03
   2. BACKGROUND RADIANCE, W/M**2/SR/A                     .1709E+04
   3. BACKGROUND IRRADIANCE, W/M**2/A                     0.
   4. ATMOSPHERIC LOSS                                     .1000E+01     0.00
RECEIVER PARAMETERS
   1. RECEIVER ANTENNA GAIN                                .9355E+10     99.71
        APERTURE DIAMETER, METERS                           .2540E-01
        OBSCURATION DIAMETER, METERS                       0.
   2. TRANSMISSION EFFICIENCY                              .5164E+00    -2.87
        OPTICS EFFICIENCY                                   .5164E+00
   3. OPTICAL FILTER TRANSMISSION                          .9000E+00     -.46
   4. OPTICAL FILTER BANDWIDTH, ANGSTROMS                  .1000E+03
   5. RECEIVER DIAMETRICAL FIELD OF VIEW, RAD              .1000E-02
   6. AC COUPLING LOSS                                     .1000E+01     0.00
DETECTOR PARAMETERS
   1. DETECTOR TYPE                                      APD
        DETECTOR NOISE FACTOR                               .1959E+01
        AVALANCHE GAIN                                      .9000E+01
        IONIZATION COEFFICIENT                              .7000E-02
   2. DETECTOR QUANTUM EFFICIENCY                          .8000E+00
   3. DETECTOR NOISE EQUIVALENT POWER, W/HZ^.5             .1025E-13
   4. PREAMP NOISE CURRENT DENSITY, A/HZ^.5                .6000E-11
        EFFECTIVE RECEIVER TEMP., DEG K                     .3000E+03
        EFFECTIVE RECEIVER RESISTOR, OHMS                   .4602E+03
   5. NOISE EQUIVALENT BANDWIDTH, HERTZ                    .4000E+09
   6. PROCESSING LOSS                                      .5623E+00
RECEIVED SIGNAL AND NOISE PARAMETERS
   1. RECEIVED SIGNAL, WATTS PEAK                          .2738E-05    -55.63
        RECEIVED PHOTOELECTRONS/PULSE                       .1180E+05
   2. NOISE POWER SPECTRAL DENSITY, A^2/HZ                 .1132E-22
                            IN (PE/PULSE)^2                 .2759E+06
        OPTICAL BACKGROUND POWER DENSITY, A^2/HZ            .1088E-22
                            IN (PE/PULSE)^2                  .2650E+06
        DETECTOR NOISE POWER DENSITY, A^2/HZ                .3204E-28
                            IN (PE/PULSE)^2                  .7808E+00
        PREAMP NOISE POWER DENSITY, A^2/HZ                  .4445E-24
                            IN (PE/PULSE)^2                  .1082E+05
REQUIRED SIGNAL PARAMETERS
   1. REQUIRED SIGNAL, WATTS PEAK                          .1343E-05    -58.72
        REQUIRED SIGNAL, PE/PULSE                           .5786E+04
   2. REQUIRED SIGNAL TO NOISE RATIO                       .2404E+02
=======================================================================
     SYSTEM LINK MARGIN                          -         .2039E+01     3.09
```

LinCom

Figure 3.

	TRW APPROACH		RECOMMENDED APPROACH	
TRANSMIT POWER	43.8 mV	-13.59 dBW	0.53 mW	-32.76 dBW
OPTICS TRANSMISSION	0.5	-3.01 dB*	0.5	-3.01 dB
POINTING LOSS	2 mrad	-3.73 dB*	354 μrad	-4.34 dB
TRANSMIT GAIN	6 mrad	58.69 dB*	1 mrad	75.05 dB
WAVEFRONT LOSS	σ = .15	-3.86 dB*	σ = .15	-3.86 dB
RANGE LOSS	150 M	-187.18 dB*	150 M	-187.18 dB
RECEIVER GAIN	1"	99.71 dB*		99.71 dB
RECEIVER OPTICS	0.5	-3.01 dB*	0.55	-2.60 dB
FILTER TRANSMISSION	--	--	0.80	-.97 dB
STRUT LOSS	.93	-0.3 dB*	--	--
ADDITIONAL MARGIN	2	-3.00 dB*	2	-3.00 dB
RECEIVED SIGNAL		-59.28 dBW		-69.96
REQUIRED SIGNAL		-59.43 dBW*		-62.97 dBW***
EXCESS SIGNAL		.15 dB		.01

The given approach lists a transfer loss of 45.81 dB, adding the asterisked items gives -45.69 which may account for the discrepancy

**The required signal is based upon 800 MHz BW, 1.5 dB proc. loss, 1:1 extinction, and -45 dBW background (actually sun in field yields -46.2 dBW) 1 mrad FOV and 100 Å filter.

***The required signal here is based upon 400 MHz BW and 2.5 dB or proc. loss, a 5:1 extinction ratio, a 500 μrad FOV and a 40 Å filter BW. (Sun in field gives - 55 dBW background).

LinCom

Figure 4. Link Analysis for Recommended Parameters.

```
FILENAME:LINDSY2 DATE: 84/11/16

SPACE STATION FREE SPACE TRANSFER LINK
LINK PARAMETERS
   1. LASER TYPE                                         GAALAS
   2. WAVELENGTH, METERS                                 .8250E-06
   3. LINK TYPE                                          COMMUNICATION
   4. SYMBOL ERROR PROBABILITY                           .1000E-06
   5. DATA RATE, BPS                                     .4000E+09
   6. MODULATION FORMAT                                  MANCHESTER
   7. EXTINCTION RATIO                                    5.:1
                                                          FACTOR         DB
TRANSMITTER PARAMETERS
   1. PEAK POWER, WATTS                                  .5200E-03     -32.76
        AVERAGE POWER, WATTS                              .2180E+00
        PULSEWIDTH, SECONDS                               .1250E-08
        PULSE RATE, PPS                                   .4000E+09
   2. TRANSMISSION EFFICIENCY                            .2055E+00      -6.87
        OPTICS EFFICIENCY                                 .5000E+00
        RMS WAVEFRONT ERROR LOSS                          .4110E+00
        CIRCULARIZING LOSS                                .1000E+01
   3. POINTING LOSS                                      .3670E+00      -4.35
        POINTING BIAS ERROR, RADIANS                      .3540E-03
        POINTING JITTER ERROR, RADIANS RMS               0.
        TOTAL POINTING ERROR BIAS+JITTER, RADIANS         .3540E-02
   4. TRANSMITTER GAIN (HAND INPUT GTYPE=6)              .3200E+08      75.05
        APERTURE DIAMETER, METERS                         .1000E-01
        OBSCURATION DIAMETER, METERS                     0.
        BEAM DIVERGENCE, RADIANS                          .1000E-02
CHANNEL PARAMETERS
   1. FREE SPACE LOSS                                    .1916E-18    -187.18
        RANGE SEPARATION, METERS                          .1500E+05
   2. BACKGROUND RADIANCE, W/M**2/SR/A                   .1709E+04
   3. BACKGROUND IRRADIANCE, W/M**2/A                   0.
   4. ATMOSPHERIC LOSS                                   .1000E+01       0.00
RECEIVER PARAMETERS
   1. RECEIVER ANTENNA GAIN                              .9355E+10      99.71
        APERTURE DIAMETER, METERS                         .2540E-01
        OBSCURATION DIAMETER, METERS                     0.
   2. TRANSMISSION EFFICIENCY                            .5500E+00      -2.60
        OPTICS EFFICIENCY                                 .5500E+00
   3. OPTICAL FILTER TRANSMISSION                        .8000E+00       -.97
   4. OPTICAL FILTER BANDWIDTH, ANGSTROMS                .4000E+02
   5. RECEIVER DIAMETRICAL FIELD OF VIEW, RAD            .5000E-03
   6. AC COUPLING LOSS                                   .1000E+01       0.00
DETECTOR PARAMETERS
   1. DETECTOR TYPE                                     APD
        DETECTOR NOISE FACTOR                             .2121E+01
        AVALANCHE GAIN                                    .2500E+02
        IONIZATION COEFFICIENT                            .7000E-02
   2. DETECTOR QUANTUM EFFICIENCY                        .6300E+00
   3. DETECTOR NOISE EQUIVALENT POWER, W/HZ^.5           .1025E-13
   4. PREAMP NOISE CURRENT DENSITY, A/HZ^.5              .6000E-11
        EFFECTIVE RECEIVER TEMP., DEG K                   .2000E+03
        EFFECTIVE RECEIVER RESISTOR, OHMS                 .4602E+03
   5. NOISE EQUIVALENT BANDWIDTH, HERTZ                  .4000E+09
   6. PROCESSING LOSS                                    .5623E+00
RECEIVED SIGNAL AND NOISE PARAMETERS
   1. RECEIVED SIGNAL, WATTS PEAK                        .1008E-05     -59.96
        RECEIVED PHOTOELECTRONS/PULSE                     .4235E+04
   2. NOISE POWER SPECTRAL DENSITY, A^2/HZ               .1181E-23
                              IN (PE/PULSE)^2             .2875E+05
        OPTICAL BACKGROUND POWER DENSITY, A^2/HZ          .1123E-23
                              IN (PE/PULSE)^2   -          .2735E+05
        DETECTOR NOISE POWER DENSITY, A^2/HZ              .3204E-26
                              IN (PE/PULSE)^2              .7802E+00
        PREAMP NOISE POWER DENSITY, A^2/HZ                .5760E-25
                              IN (PE/PULSE)^2              .1402E+04
REQUIRED SIGNAL PARAMETERS .
   1. REQUIRED SIGNAL, WATTS PEAK                        .5051E-06     -62.97
        REQUIRED SIGNAL, PE/PULSE                         .2176E+04
   2. REQUIRED SIGNAL TO NOISE RATIO   -                 .2404E+02
=============================================================================
     SYSTEM LINK MARGIN                        -         .1997E+01       3.00
```

TRW Electronic Systems Group
One Space Park
Redondo Beach CA 90278
213 535 4321

9 November 1984

Dr. William C. Lindsey
Chairman of the Board
LinCom Corporation
3200 Wilshire Blvd., Suite 300
Los Angeles, California 90010

Subject: Short Range Laser Links for Space Station

Dear Dr. Lindsey:

I am enclosing a few computer runs for laser link budgets for 400 and 10 Mbps and for PIN diode and APD detectors.

Most of the parameters are obvious, but a few need explanation. Consider:

a) Pre-amp noise: the indicated value is the value of current noise/Hz expressed as amps^2/Hz. The specific value is a function of video bandwidth, hence is different for the 400 and 10 Mbps cases.

b) Manchester data format is assumed.

c) Extinction ratio is the off/on optical power ratio.

d) Strut loss - ignore this parameter; I have assumed unobstructed apertures.

e) $\sigma\lambda$ is an allowance for RMS wave front error at the transmitting aperture.

f) P_I is the background power, in dBw that reaches the detector.

g) Preq is the signal power required at the detector, expressed in dBw.

h) P_{TX} is the required source laser power in milliwatts.

i) M is the optimum value of APD avalanche gain, for the indicated value of P_I.

The last page is a solar interference analysis, based on the receiver FOV looking directly at the sun, and intercepting a fraction of the solar energy in a bandwidth, $\Delta\lambda$. As you can see, the solar interference puts the required power near the top of each of the three computer runs where the required laser power for 400 Mbps is quite large.

Note that the transmitter beamwidth (center to e^{-2} level) is approximately 3 mR. Pointing errors are assumed to be 2 mR. This is equivalent to a pointing error of .984 feet at 492 feet (0.15Km). I understand that Space Station keel distortion can be as much as ± 5 feet. It follows that it is necessary to put the laser transmitter and receiver on a gimbal, or to use controllable tilt mirrors with auto track to maintain the link. If one goes to this expense, the transmit beamwidth might as well be narrowed to improve link performance and to permit reducing laser power.

If you would like to discuss this further, please call me at 536-2698.

Sincerely,

Bud

S. B. Franklin
TRW Electronic Systems Group

SBF:dh

20 OCT 198

OPTICAL LINK BUDGET FOR: SPACE STATION.

INTERNAL LINKS -

$R = 0.15 Km$
$= 492$ feet

$\theta_{B_{TX}} = 3.05$ mR

$D_{REC} \approx 1$ inch

$R_D = \left.\begin{array}{l} 400 \text{ MBPS} \\ 10 \text{ MBPS} \end{array}\right\}$ APD, DE

$= 400$ MBPS – PIN DIODE

ORIGINAL PAGE IS
OF POOR QUALITY

RUN DATE 29 OCT 1984
1-PIN DIODE DET 0
POWER REQUIRED FOR DATA

400 MBPS

PIN DIODE

DATA RATE	400000
KBPS	
BER	0000001
? Q	5.19969
BANDWIDTH	800000 KHZ
? POWER RATIO	1
EXTINCTION RATIO	.1
NOISE FACTOR,K2	05
POST DETECT DEGRAD	1.5 DB
PREAMP NOISE	3.6E-23
? I-M(PICOAMPS)	0
? I-NM(NANOAMPS)	.1
LAMBDA	8250
ANGSTROMS	
MARGIN	3 DB
RANGE	.15 KM
TNX APERTURE DIA=	.01
TNX OBSCURATION	0
? TNX APERTURE AREA	0 M^2 what kind of antenna
TNX BEAMWIDTH	3053 15 μR
RCV APERTURE DIA	1 INCHES
RCV OBSCURATION	0
RCV APERTURE AREA	.0005 M^2
TNX τ	.5
RCV τ	.5
POINTING ERROR	2000 μR
POINTING LOSS	-3.727 DB
σ-λ	.15
WFE LOSS	-3.858 DB
STRUT LOSS	3 DB
TNX GAIN	58.692 DB
? TRANSFER LOSS	45.81 DB

PI	M	P-REQ	P-TX
-40	1	-56 236	90.618
-45	1	-56 773	80 069
-50	1	-56.975	76 431
-55	1	-57.043	75.243
-60	1	-57 065	74.864
-65	1	-57.072	74.743
-70	1	-57 074	74.705
-75	1	-57.075	74 693
-80	1	-57.075	74 689
-85	1	-57 075	74.688
-90	1	-57 075	74 688
-95	1	-57.075	74 688
-100	1	-57.075	74.688
-105	1	-57.075	74 688
-110	1	-57.075	74.688
-115	1	-57.075	74 688
-120	1	-57.075	74 688
-125	1	-57.075	74.688

P_I = -46.2 dB
FOR
$\overline{\Delta\lambda}$ = 100 Å
FOV = 1 mR

background dBW | signal power required at detector dBW | required source laser power dBW

RUN DATE 20 OCT 1984
3-APD DET 0
POWER REQUIRED FOR DATA

BER	.0000001
Q	5.19969
DATA RATE	400000
KBPS	
BANDWIDTH	800000 KHZ
POWER RATIO	1
EXTINCTION RATIO	.1
NOISE FACTOR,K2	05
POST\DETECT DEGRAD	1.5 DB
PREAMP NOISE	3.6E-23
I-M(PICOAMPS	1
I-NM(NANA-AMPS	15
LAMBDA	8250 Å
ANGSTROMS	
MARGIN	3 DB
RANGE	.15 KM
TNX APERTURE DIA=	.01
TNX OBSCURATION	0
TNX APERTURE AREA	0 M^2
TNX BEAMWIDTH	3053.15 μR
RCV APERTURE DIA	1 INCHES
RCV OBSCURATION	0
RCV APERTURE AREA	0005 M^2
TNX τ	.5
RCV τ	5
POINTING ERROR	2000 μR
POINTING LOSS	-3.727 DB
σ-λ	.15
WFE LOSS	-3.858 DB
STRUT LOSS	.3 DB
TNX GAIN	58.692 DB
TRANSFER LOSS	45.81 DB

400 mbrs

APD DET.

PI	M	P-REQ	P-TX
-40	2	-57 220	73 736
-45	4	-59.434	43 879
-50	8	-61.644	26 104
-55	12	-63.787	15.947
-60	18	-65.818	9 986
-65	26	-67.679	6 504
-70	36	-69 282	4 494
-75	46	-70.512	3.386
-80	54	-71 283	2 835
-85	58	-71 656	2.602
-90	60	-71.800	2.517
-95	60	-71.849	2 489
-100	61	-71 865	2 479
-105	61	-71.870	2 476
-110	61	-71.872	2.475
-115	61	-71.872	2 475
-120	61	-71 872	2.475
-125	61	-71.872	2 475

ORIGINAL PAGE IS
OF POOR QUALITY

RUN DATE 20 OCT 1984
3-APD DET 0
POWER REQUIRED FOR DATA

BER	0000001	10 MBPS
Q	5.19969	
DATA RATE	10000 KBPS	APD - DET
BANDWIDTH	20000 KHZ	
POWER RATIO	1	
EXTINCTION RATIO	1	
NOISE FACTOR,K2	05	
POST\DETECT DEGRAD	1 5 DB	
PREAMP NOISE	4.9E-25	
I-M(PICOAMPS	1	
I-NM(NANA-AMPS	15	
LAMBDA	8250	
ANGSTROMS		
MARGIN	3 DB	
RANGE	.15 KM	
TNX APERTURE DIA=	.01	
TNX OBSCURATION	0	
TNX APERTURE AREA	0 M^2	
TNX BEAMWIDTH	3053.15 µR	
RCV APERTURE DIA	1 INCHES	
RCV OBSCURATION·	0	
RCV APERTURE AREA	0005 M^2	
TNX τ	.5	
RCV τ	.5	
POINTING ERROR	2000 µR	
POINTING LOSS	-3.727 DB	
$\sigma-\lambda$	.15	
WFE LOSS	-3.858 DB	
STRUT LOSS	.3 DB	
TNX GAIN	58 692 DB	
TRANSFER LOSS	45.81 DB	

PI	M	P-REQ	P-TX	
-40	1	-66.656	8.226	
-45	1	-69.025	4 767	P = -46.2 dBW
-50	1	-71.160	2.916	
-55	1	-72.983	2.001	FOR
-60	3	-75 135	1 170	FOV = 1 mR (5m.!
-65	5	-77.354	705	$\Delta\lambda$ = 10 nm
-70	9	-79.535	424	
-75	13	-81.626	262	
-80	20	-83.571	167	
-85	28	-85 287	113	
-90	36	-86.663	092	
-95	44	-87 589	.066	
-100	48	-88.074	059	
-105	50	-88 273	057	
-110	51	-88.343	056	
-115	51	-88.366	056	
-120	51	-88.373	055	
-125	51	-88.375	.055	

Solar Interference

$$E_D = E_T \, \overline{\Delta\lambda} \; A_{Receiver} \left(\frac{FOV}{D_{sun}} \right)^2 \tau$$

=

$$E_T = 1050.7 \;\; w/m^2 \;\; at \;\; 8250 A^\circ$$

Let $\overline{\Delta\lambda} = 100 \, A^\circ = 10 nm$

$$A_{Receiver} = \frac{\pi D^2}{4}$$

$$= \frac{\pi}{4} (.0254)^2 \quad For \; D = 1 \; inch$$

$$= 5.067 \times 10^{-4} \; m^2$$

$$FOV = 10^{-3} \; Radian$$

$$D_{sun} = 0.603848 \; DEG$$

$$= 10,539 \; \mu R$$

$$\tau = 0.5$$

Then $E_D = E_T \, \overline{\Delta\lambda} \; A_R \left(\frac{FOV}{D_{sun}} \right)^2 \tau$

$$= (1050.7)(.010)(5.067 \times 10^{-4}) \left(\frac{10^{-3}}{10.539} \right)^2 (.5)$$

$$= 23.96 \times 10^{-6} \; Watts$$

$$= -46.20 \; dBW$$

FOV = 1 mR
$\overline{\Delta\lambda}$ = 10 nm
Rec. D. = 1 inch.
τ = .5

APPENDIX I

A REVIEW OF MULTIPLE-ACCESS SCHEMES

Introduction

Four different multiple-access schemes are discussed. The systems are addressed in the context of the envisioned Space Station multiple access scenarios and requirements. Each scheme is described and its advantages and drawbacks are identified. Most of the features of importance in the Space Station application are addressed. These include bandwidth utilization, power requirements, hardware complexity, interference rejection, potential for growth and ranging capabilities.

C-2

1. Frequency-Division Multiple Access

In frequency-division multiple access (FDMA) the available communication bandwidth is divided into a number of nonoverlapping frequency bands which constitute access channels. Each link is allocated an access channel.

The selection of the frequency channels may be fixed or assigned. In fixed frequency operation each user carrier is assigned a dedicated frequency band. In demand-assigned operation the frequency bands are shared by different user carriers, with a particular band assigned at the time of need, depending on availability. A demand-assigned (DA) FDMA system is usually more efficient than a fixed system when the usage times of the different carriers are relatively low. It may require, however, more complex routing control and link setup hardware. For the space station application since the different users have significantly varying communication requirements, with each user possibly needing different kinds of service at the same or at different times, the need is for a flexible and adaptable multiple-access scheme.

Although FDMA is generally the simplest form of multiple accessing, and the required technology is proven, relatively simple and available at reasonable cost in today's market, there are three main factors that limit its permanance and may conceivably cause it to be unsuitable for application in the space station. The three limiting factors are crosstalk due to spectral overlap of adjacent access channels, intermodulation distortion due to nonlinear amplification and weak signal suppression if carriers of different strengths share the same amplifying devices. The first factor, viz. crosstalk, is particularly important when the channels handle high data rates (like 22 Mbps TV data) over the limited available bandwidth. The second factor concerning intermodulation distortion is important when several users have to be addressed through the same amplifying devices simultaneously creating the need for high power amplification, which is frequently obtained at the price of nonlinearity. The last factor regarding weak signal suppression may arise when the signals from users at vastly different ranges are amplified at the space station using the same amplifier. This may lead to amplifier saturation due to the strong signal, which in turn causes power robbing from the weak signals. In what follows the three limiting factors are discussed briefly.

Crosstalk power due to overlap of adjacent spectra is a function of separation between the adjacent channels. The spacing between adjacent carriers is usually denoted by $qB/2$ where $B/2$ is half the main lobe bandwidth for NRZ. This is illustrated in Fig. 1. The power of the i-th carrier is simply given by

$$P_i = \int_0^{\infty} S_i(f)df \tag{1}$$

where $S_i(f)$ is the one-sided power spectral density (PSD) of the i-th carrier. The fractional crosstalk power of the i-th carrier falling into the adjacent carrier bandwidth is given by

$$C_i = \frac{1}{P_i}\int_{(q-1)B/2}^{(q+1)B/2} S_i(f_i+f)df \tag{2}$$

Thus the crosstalk power is a direct function of the spectral shape and spacing. For illustrative purposes we consider Butterworth filtered BPSK and QPSK modulated carriers. The power to crosstalk ratio P_i/C_i is

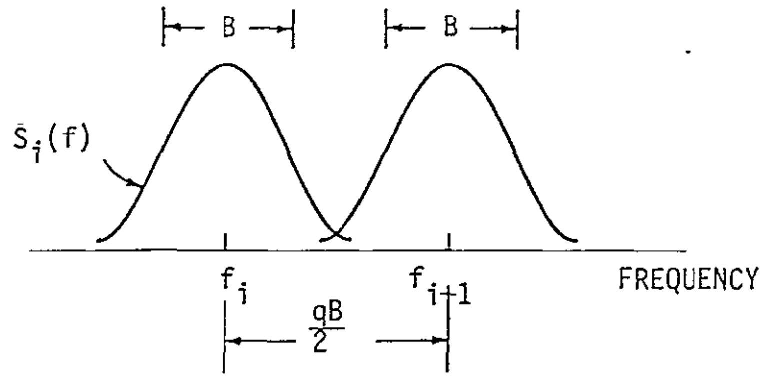


Fig. 1. Carrier Spacing of Adjacent FDMA Spectra.

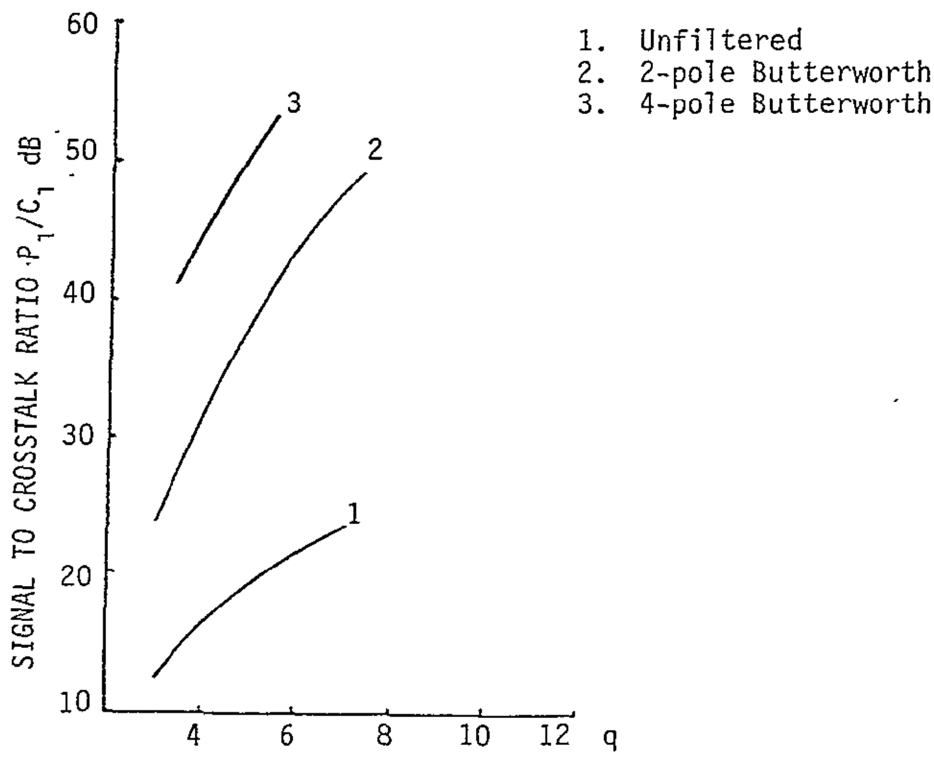


Fig. 2. Signal to Crosstalk Ratio as a Function of Separation of Adjacent Carriers for BPSK or QPSK.

plotted in Fig. 2 versus carrier separation for different degrees of filtering, while assuming equal carrier powers for simplicity. It is seen that with no filtering a rather large separation q=9 is needed to keep P_i/C_i at about 25 dB. It is also seen that a moderate amount of filtering brings the separation down significantly for example, q=3 for $P_i/C_i \approx 25$ dB with two-pole Butterworth filtering. Filtering obviously suppresses the tails or outer sidelobes of the modulation spectra. This has the undesirable effects of introducing carrier distortion and decoding degradation. Moreover, the suppressed sidelobes are susceptible to the tail regeneration problem associated with nonlinear amplification. Hence, if FDMA is adopted for the space station, particularly for the high data rate services like the 22 Mpbs TV transmission, a delicate balance has to be found among channel separation, extent of filtering and tolerable nonlinear amplification.

In an earlier study of the application of FDMA to the space station [1] it was assumed that it was possible to operate the high power amplifiers (HPA) significantly backed-off from saturation, avoiding the introduction of intermodulation interference. The link budgets developed in [1], and upon which such an assumption was made, oversee the possibility of transmitting simultaneously to many users through the same amplifying device. As mentioned earlier this would proportionally increase the power requirements of the HPA and may very well lead to operation close to saturation.

The intermodulation products due to the nonlinear characteristics depend on the nonlinearity model used. A hard-limiter model is appropriate if full output is desired. If some backoff is adopted then a soft limiter model becomes more appropriate. Several nonlinear analyses have been performed for the different models. They adopt different approaches depending on the number of carriers used in the given bandwidth. The analyses are generally involved and can be found in several references, e.g., [2,3]. Since the carrier to intermodulation noise ratio can be the limiting factor when crosstalk is maintained at a low level we give the following illustrative example. K equal amplitude carriers with identical spectral occupancy are employed with a soft limiting amplifier having the characteristics

$$g(V_{in}) = \begin{cases} V_L \operatorname{erf}\left(\dfrac{V_{in}}{V_{in_{sat}}}\right) & V_{in} > 0 \\ -V_L \operatorname{erf}\left(\dfrac{|V_{in}|}{V_{in_{sat}}}\right) & V_{in} < 0 \end{cases} \quad (3)$$

The input amplifier backoff is defined as

$$\beta_i = \left(\frac{V_{in_{sat}}}{V_{in}}\right)^2 \quad (4)$$

If to facilitate the computation of the needed convolutions, the carrier spectra are taken as Gaussian

$$S_i(f) = \frac{\exp[-(2\pi f - 2\pi f_i)^2/2B_i]}{\sqrt{2\pi B_i}} \quad (5)$$

the resulting carrier to intermodulation noise ratio is plotted in Fig. 3 versus the input backoff for different numbers of carriers. The figure shows that the contribution of intermodulation noise can be

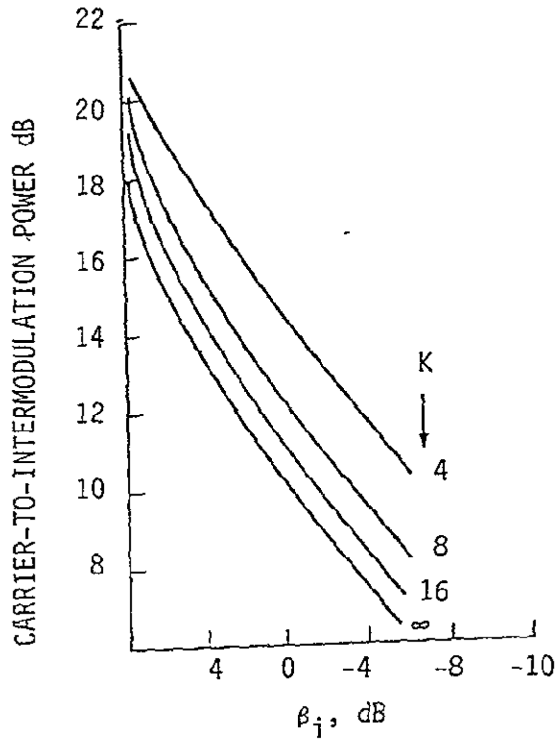


Fig. 3. Carrier-to-Intermodulation Power Ratio vs. Input Backoff for Different Number of Channels; Center FDMA Channel Assumed (Worst Case).

significant compared to crosstalk if insufficient backoff is adopted. A more careful analysis of intermodulation and the involved tradeoffs is therefore recommended if FDMA is seriously considered for the space station.

Weak signal suppression can appear if in the absence of power control the signals from distant free-flyers share the same low noise amplifier (LNA) with other users which are much closer in range to the space station. If the LNA saturates and the ratio of the received powers is in excess of 4:1 further suppression of the weak signal occurs. This can be avoided if the dynamic range of the amplifier is wide enough to encompass the variation in signal powers over the permitted range variation. This requires segmenting the 0 to 2000 Km range into different regions and using different receiving equipment for each of those ranges. Power control among the different users may also achieve the same result. In both cases the solution needs extra hardware. The cost of such hardware should enter in the overall tradeoff study of the different multiple access schemes envisioned for the space station.

Other factors that can influence the choice of a multiple access scheme are interference rejection and the capability to provide ranging. FDMA is a simple multiple access scheme that does not provide interference rejection or ranging capabilities. These factors, however, are by no means sufficient reasons to discount FDMA from consideration; ranging information may be available from other sources (like frame sync) and the total bandwidth available, as we shall see later, does not provide for meaningful interference rejection particularly against an intentional jammer.

Finally there is the multiple-access system potential for growth. Naturally, each additional user requires transmit and receive equipment; moreover, the antenna system onboard the space station will need modification to accomodate the additional users. If the data links accomodated through FDMA have low rates or if the available RF bandwidth is wide, then adding users will have little overall impact since crosstalk and intermodulation effects would still be small. If, however, the data links are of high rate (e.g. TV links at 22 Mbps) and the available bandwidth is limited, the possibility of adding more users would be quite limited and hard to remedy.

2. Time Division Multiple Access

In time division multiple access (TDMA) the different links are separated in time rather than in frequency. Instead of allocating a frequency band for each user link, each user is assigned a specific time interval and it transmits only during that alloted interval. The different users transmit their bursts of carrier signals such that they arrive and are processed by the space station without overlapping in time. Each user is therefore able to utilize the entire system bandwidth during the short burst time alloted for it. This approach avoids the crosstalk and intermodulation problems associated with simultaneous processing of the different carriers in the FDMA scheme. As a result TDMA can generally achieve signficantly better efficiency in bandwidth utilization. This is achieved, as we shall see in what follows, at the expense of higher power requirements, elaborate network timing control, and hardware that is capable of handling high speed data in terms of acquisition, tracking and decoding.

The basic unit used in describing the transmission in TDMA is the frame. Each frame contains N slots. Each slot contains the burst transmitted from each user. The frame contains also other burst(s) for network synchronization and control. The different bursts are separated by guard times to prevent time overlap of the different bursts.

A typical frame layout is shown in Fig. 4. As seen, each message burst, which is basically a modulated carrier burst, consists of a preamble, a message segment and a postamble. The preamble is further subdivided into a portion for carrier recovery (since carrier acquisition has to be performed at the beginning of each burst), a portion for symbol timing recovery, a burst code word sometimes referred to as a unique word which is used to establish word sync, a station or user identification code, and finally other miscellaneous data bits which are used mainly for timing control. The preamble is followed by the data segment which contains the actual information to be communicated between the user and the space station or vice-versa. The last segment of the carrier burst is called the postamble and may contain data bits to aid in fine synchronization.

In the space station application the space station performs the task of timing control and distribution to the different users. The station provides the reference timing and transmits the burst timing code words to the different users. These words are then returned by the users to the station which computes the propagation delays and then provides the users with the correct timing information. The space station is also responsible for controlling slot assignments and user initial access to the network. A network accessing protocol for TDMA has been suggested for the space station [1].

The hardware required for the performance of the timing control in

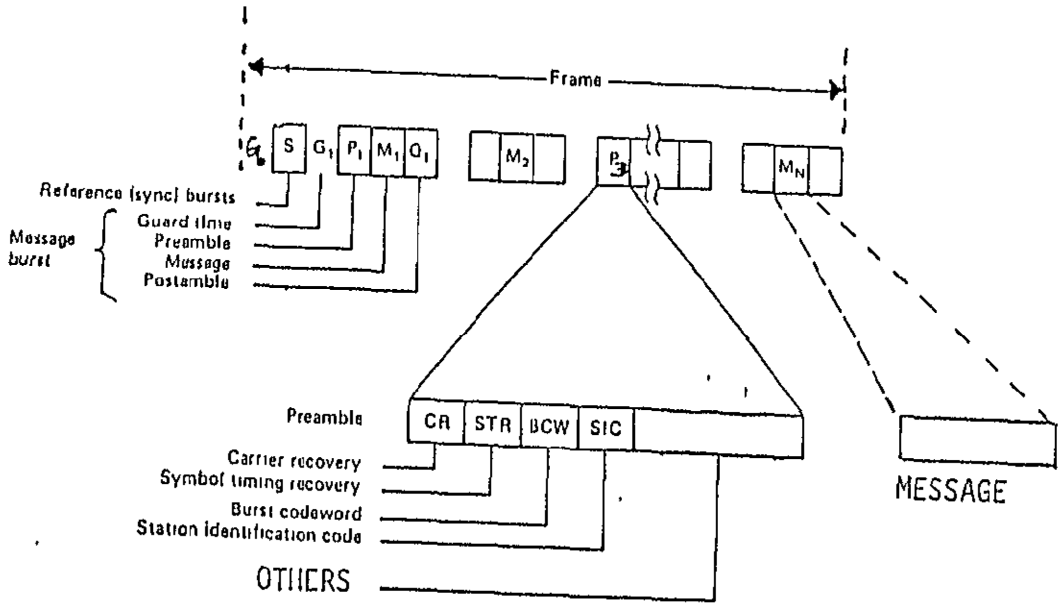


Fig. 4. A Typical TDMA Frame Format.

the network is more complicated than the type of hardware needed with FDMA. The space station and each user is supplied with a "burst modem" which contains buffers, a preamble processor and a burst codeword detector in addition to high speed encoder/decoder. The hardware for TDMA may not in itself be the limiting factor. The technology has matured and burst rates up to 100 Mbps are achievable. The physical size of the equipment yet may not be suitable for deployment in space.

The loss of efficiency due to the overhead functions of timing and network control, as well as guard times between bursts, which collectively consume about 15% of the overall system capacity, is also not a limiting factor for the application of TDMA.

TDMA has been used in satellite relay systems, where bandwidth is at a great premium, and where large antennas can be used in the earth stations at one or more ends of the communication links. The object being to provide the high gain necessary for the link budgets encountered in TDMA. The large power requirement stems directly from the fact that several users are communicating, virtually in real time, simultaneously. So the effective bit rate for a single user is multiplied by the number of users and further divided by the frame efficiency. This increase in needed power to preserve an acceptable E_b/N_0 (say to provide 10^{-5} BER) is sometimes referred to as TDMA loss. For 16 users it is about 13 dB. There might be also higher hardware losses due to the high data rate operation. Part of the increase in power can be achieved by driving the high power amplifiers well into saturation to obtain maximum amplifier gain. The remainder of the needed gain has to be obtained through higher antenna gains. When the high gain antennas are mounted in space on the space station, antenna size and weight, the so-called real estate problem, becomes of prime importance. The necessary cost of developing and deploying such antennas can become the ultimate factor that determines whether TDMA is or is not suitable for the space station application.

Among the other factors that play a role in the choice of the multiple access scheme are ranging and interference rejection capabilities. TDMA has a built-in ranging capability that is essential for its proper operation. In addition, there is no need for user power control as there is no near-far problem in TDMA, different users simply do not share amplifying or receiving equipment simultaneously. TDMA, however, is like FDMA in that it does not possess inherent interference rejection.

Finally, a growth in the number of users would impact the TDMA system in two ways. First, the basic frame construction has to be restructured to allow for a larger number of slots. The logic of the burst modems of the users will have to be changed accordingly. It is therefore greatly advantageous to design the system for the anticipated larger number of users, even though this may require a fair amount of guessing and perhaps waste if no significant growth materializes. The second way in which growth affects the TDMA system is the requirement for higher antenna gain to accomodate the even higher bit rate with a fixed acceptable E_b/N_0. If, however, the antenna is designed with sufficient gain margin to start with, then user growth would have no effect on the antenna system. It has to be kept in mind, though, that antenna gain is one of the most expensive commodities in the TDMA system.

3. Direct Sequence Multiple Access

In direct sequence multiple access (DSMA) the different users are identified by uniquely separable address codes embedded within their carrier waveforms. The addresses take the form of unique pseudo-random noise (PN) code sequences. All users transmit basically whenever they desire and the different waveforms are superimposed. As such, no frequency or time separation is needed and user separation is accomplished solely on the basis of recognizing the unique PN code of each user. Like TDMA, CDMA carriers utilize the entire communication bandwidth available for the multiple access system. CDMA, however, has the advantage over TDMA that there is no requirement for timely transmission bursts. Alternately, CDMA systems performance depends heavily on the ability to recognize the individual PN codes, which often becomes difficult if: (a) the number of users is large, or (b) available bandwidth is not much wider than the data bandwidth, or (c) carriers of unequal powers share the same receiving equipment. The salient features of CDMA as well as the needed processing, which can be substantial, are discussed briefly in what follows.

A block diagram of a typical DSMA link from the space station to a user is shown in Fig. 5. After the carrier is modulated (BPSK, QPSK, etc.) and convolutionally encoded, a PN sequence derived from a PN generator is used to directly multiply and spread the modulated carrier. A number of such transmitter chains equaling the number of users is needed. On each link the modulated carrier is spread by means of the PN sequence used for that particular user. All the spread carriers are then combined in the antenna system and transmitted to the users. At a user, the first fun^ction to be performed after down-

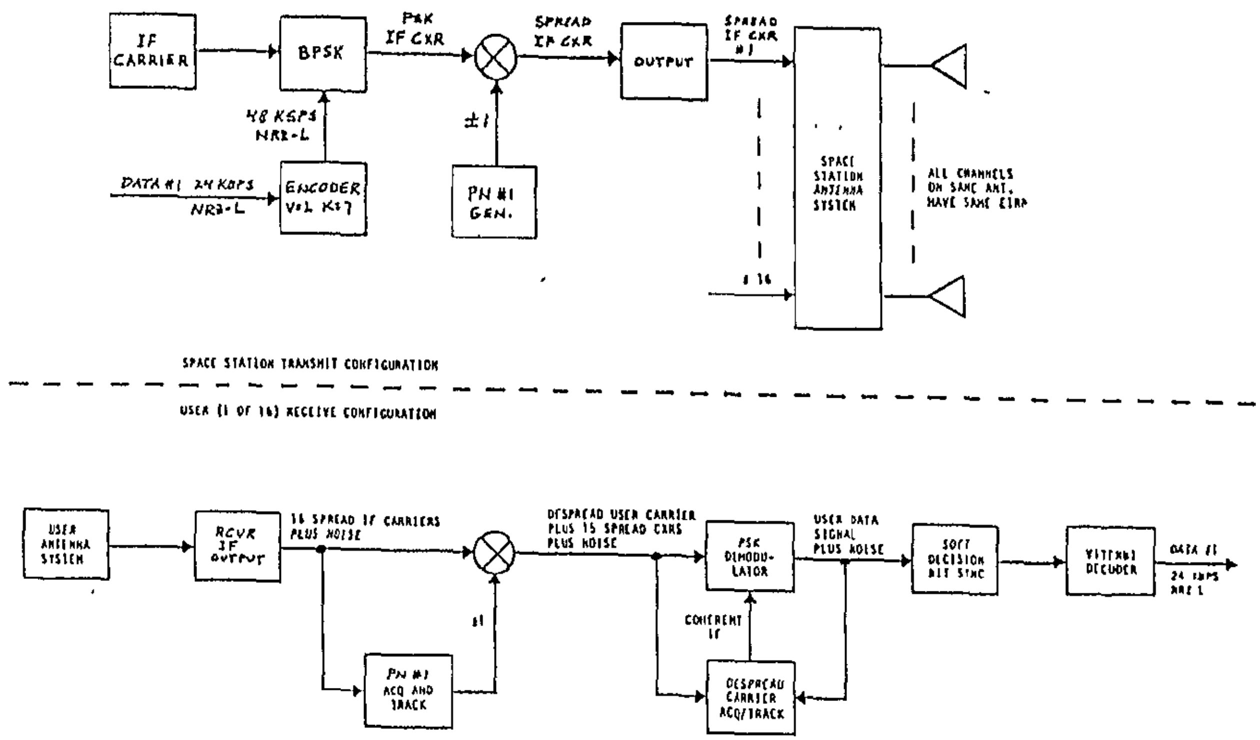


Fig. 5. Block Diagram of a Typical Space Station-to-User DSMA Link (Adapted From [1]).

conversion to IF is PN acquisition. This is then followed by carrier despreading, acquisition, tracking, coherent demodulation and finally Viterbi decoding.

In the preliminary DSMA system proposed for the space station [1] the PN acquisition function is performed by a tau-dither loop. This is coupled to a passive correlator for alignment (or misalignment) detection. The correlator threshold which determines alignment is set according to the expected signal to noise ratio E_b/N_0 of the desired carrier to be despread. Digital logic circuitry then switches the loop parameters to effect either acquisition or tracking as needed. A high-level block diagram of the PN acquisition and tracking proposed in [1] is given in Fig. 6 for illustration.

It is evident that the use of PN spreading and despreading to achieve DSMA adds to the hardware complexity of the multiple access system. This disadvantage of DSMA, however, has to be weighed against the desirable features DSMA provides which are not obtainable by TDMA or FDMA. The importance of these features for the particular application of the space station is, of course, what is of key significance.

DSMA has been used in military communication systems due to its ability to mitigate jamming. The key parameter that determines the resistence of a spread spectrum communication link to interference, intentional or otherwise, is the so-called processing gain. The processing gain is basically the ratio of the bandwidth of the modulated carrier after spreading to its bandwidth before spreading. It is also the ratio of the PN chip rate to the modulation data rate. The restriction on the increase in processing gain is usually the maximum available RF bandwidth which places an upper limit on the chip rate.

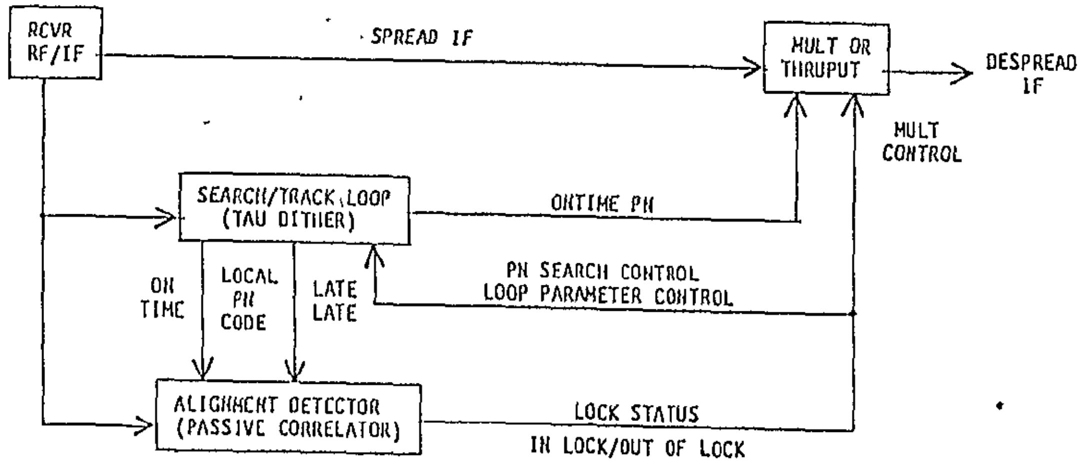


Fig. 6. A High-Level Block Diagram of the PN Acquisition and Tracking Proposed in [1] for the DSMA Scheme.

LinCom

The maximum RF bandwidths available to the forward (space station to user) and return (user to space station) links have been estimated at 50 and 100 MHz respectively. TV data links of 22 Mbps rates obviously cannot be accomodated by DSMA. Command and telemetry links of 100 Kbps can be accomodated with processing gains of about 24 and 27 dB respectively. This offers a moderate degree of interference rejection as will be discussed in what follows.

Interference can either be intentional or unintentional. We first consider unintentional interference. The receiver chain in the Space station corresponding to a particular user views the sprad signals emitted by the other users as interference. This interference effectively raises the level of background noise [4]. It makes it more difficult to acquire and track the PN code of the desired user, causes more bit errors because of an effective reduction in E_b/N_0, or alternately reduces the processing gain. If the interfering users all have equal power which is equal to the power of the desired user then the performance degaradation is not large. This is illustrated in Fig. 7 for 16 users and different processing gains. In the ordinate of the figure I_0 stands for the total interfering noise spectral density and N_0 is the customary thermal noise spectral density. It is seen that for R_c/R_b = 256 (i.e., 24 dB processing gain) the loss in effective E_b/N_0 is only about 2 dB at the value of E_b/N_0 that leads to 10^{-5} BER in a coded system. The situation is dramatically different if user power control is not employed. In the space station/free-flyer scenario a severe near-far problem exists which can completely disrupt communication relying on DSMA. The effect of the interfering users can be modeled in the case of interest (sufficiently long PN codes) by additional

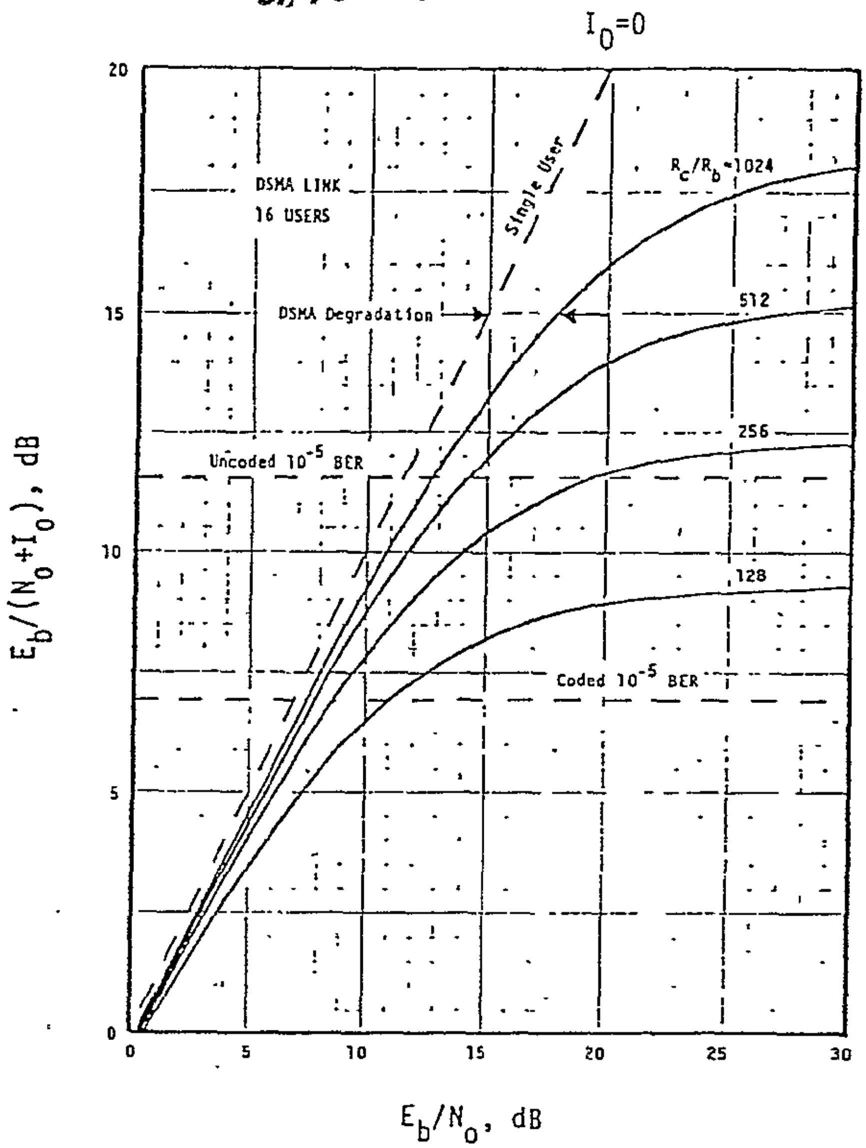


Fig. 7. DSMA Link BER Performance Degradation Due to Other DSMA System Users--16 Users of Equal Power.

I_0: Spectral Density of Total Interfering Power.

background noise [4]. If for the purposes of illustration we consider a scenario where there are only two users, one intended and one interfering, and consider the additional noise due to the interfering user to be the only source of E_b/N_0 degradation in an otherwise ideal BPSK link, we get the results of Fig. 8. The processing gain and BER are used as parameters. As the ratio of the powers of the interfering to the desired user approaches 100, i.e., range ratio of 1:10, reliable communication becomes virtually impossible. The situation becomes even worse when more interfering users are present, since each additional user effectively adds more background noise. This simply serves to show that user power control is absolutely necessary if DSMA is considered for the space station application.

As mentioned earlier, for a command data link of 100 Kbps and a maximum allowable bandwidth of 50 MHz, about 24 dB of processing gain is available. This, provided user power control is employed, provides a certain degree of intentional interference (jamming) and radio frequency interference (RFI) rejection. An RFI source with a power comparable to a user is rejected along with the unaccepted users. The immunity to jamming depends to a great extent on the strategy of the jammer. Since the DSMA is not designed primarily as an antijam system, its antijamming capabilities are significantly limited by other overriding considerations like bandwidth.

Spectral spreading, nevertheless, provides some other gains. The power or gain requirements on the transmitting/receiving amplifiers and antennas are generally lower than in FDMA. For the same information data rate, the increase in the bandwidth utilized in DSMA compared to FDMA results in a reduced link power requirement. System growth has a

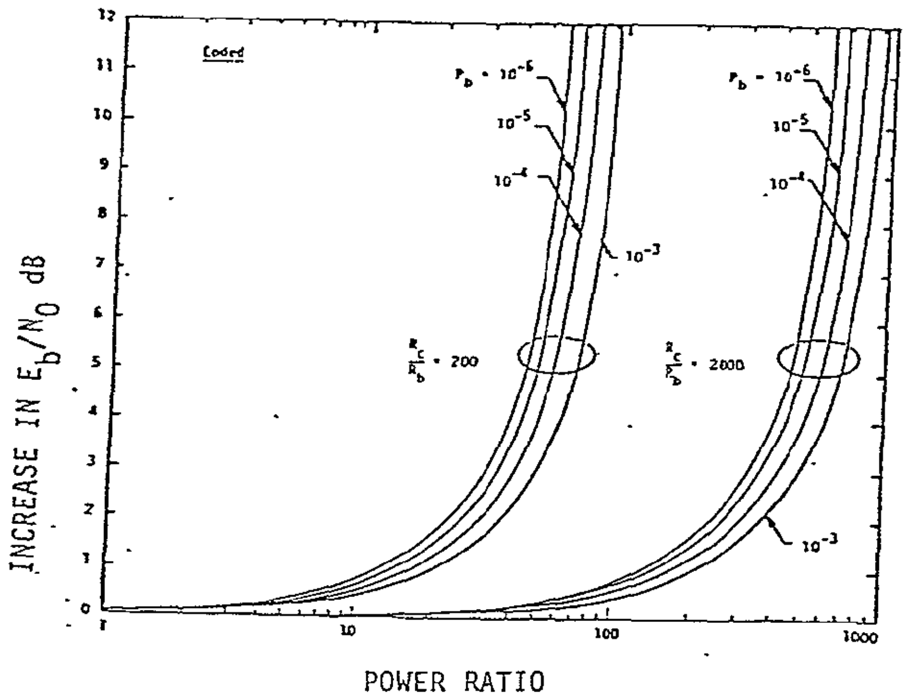


Fig. 8. Increase in E_b/N_0 versus Power Ratio for Two Users with K=7, R = 1/2 Convolutional Coding with Soft Decisions.

Power Ratio = Received Power from Interfering User/
Received Power From Intended User

R_c/R_b = Chip Rate/Bit Rate

P_b = BER

small impact on the DSMA system. As in FDMA, each additional user requires an additional receive and transmit set and the antenna system requires corresponding upgrading. The possibility of running out of bandwidth to accomodate extra users that may occur in FDMA is not present in DSMA. The addition of users, instead, manifests itself in a small degradation due to the extra interference.

Finally, a ranging capability is not automatically present in the DSMA system, but can be added by employing coherent PN code turnaround. The length of the PN code would have to accomodate the maximum range of interest. This may somewhat increase the PN acquisition time.

4. Frequency Hopping Multiple Access (FHMA)

An alternative to direct sequence multiple access is to use the PN sequence to produce frequency hopping. In this situation the available satellite bandwidth is partitioned into frequency bands and the transmission time partitioned into time slots. Each of the multiple users is assigned a unique PN code which determines its hopping pattern. With each hop the user carrier frequency is changed so that it occupies a designated band. Each user when observed over a long period of time appears to utilize the entire communication bandwidth although transmitting only within a sepcified band during any one time slot.

The space station receives all users signals simultaneously and demodulates each user signal after "dehopping" it using that user's unique PN code. This process obviously requires relatively complex and sophisticated hardware; first to hop the user carrier (or a space station carrier if on a forward link) and then to dehop, in a timely manner, the received carrier(s). In what follows we discuss briefly the system structures needed to effect FHMA and comment on the interference rejection capabilities provided by the FHMA scheme.

A functional block diagram of typical link in an FHMA system is shown in Fig. 9. This has been propsoed in [1] for the space station application. The data on a particular link is first encoded using a dual-4 convolutional code to improve system performance. The encoded data is then modulated using 16-ary FSK. The need for some form of FSK stems from the fact that as the carrier is dehopped in the receiver, loss of reference carrier phase occurs at the beginning of each hop thereby necessitating some form of noncoherent detection. Timing logic controls both the operation of the encoder and MFSK modulator, and the PN code generator and frequency synthesizer. The PN code generator controls the output frequency of the frequency synthesizer. The output of the synthesizer multiplies the output of the modulator and translates it to the appropriate frequency slot. This is then followed by RF upconversion and power amplification.

On the receiver side, a bank of IF filters followed by envelope detectors and integrators is used to detect PN alignment, and upon acquisition, perform MFSK demodulation. The integration output is compared to a threshold and is then used to control, through timing logic, the PN code generator which drives the frequency synthesizer in order for it to hop its frequency in synchronism with the received signal.

The timing clocks on both the transmitting and receiving sides have to be coherent in the sense that the time of hopping is roughly known. This makes the processes of initial PN acquisition or reacquisition take acceptably small time intervals.

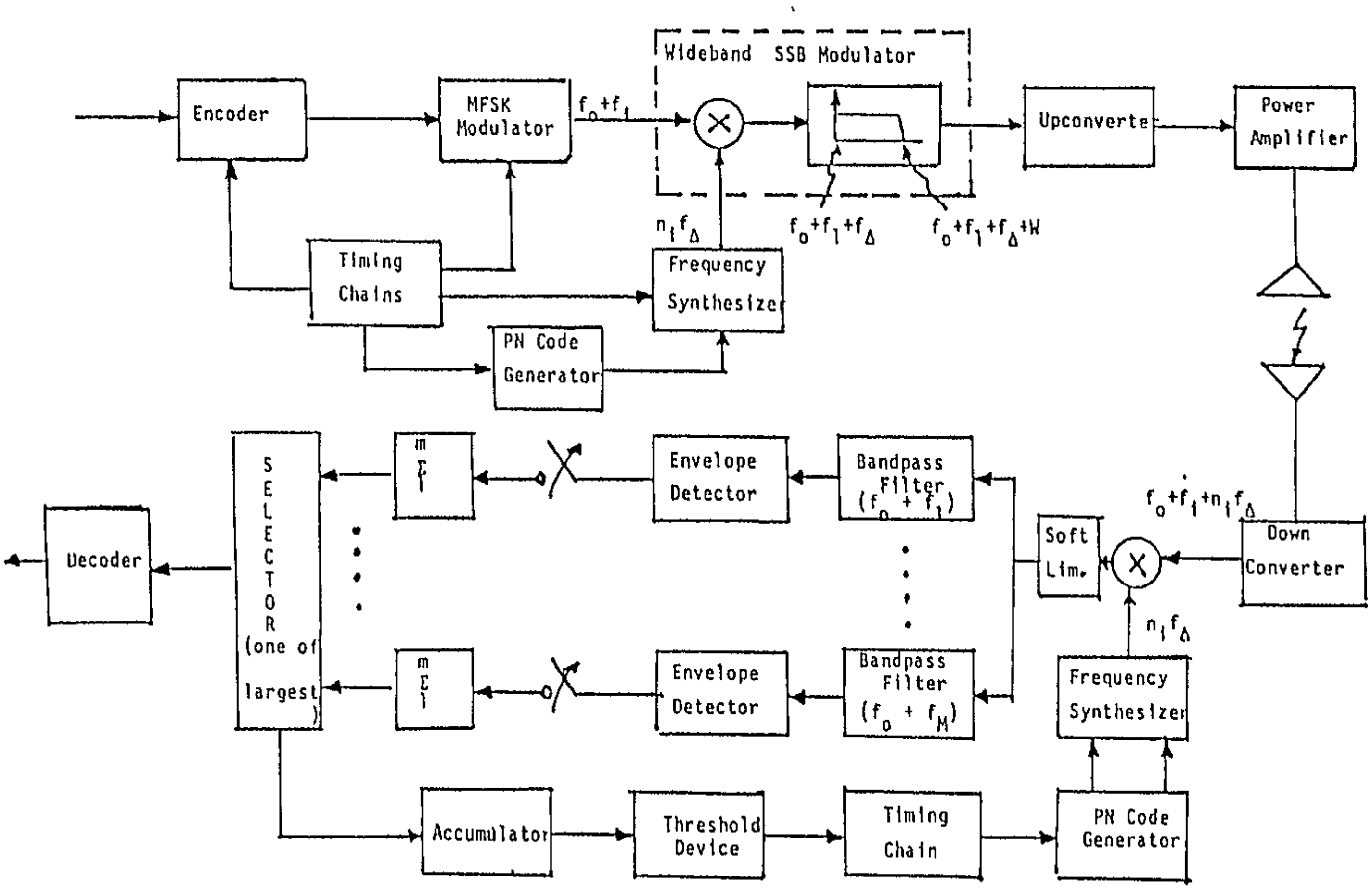


Fig. 9. Functional Block Diagram of a Typical FHMA Link.

The frequency synthesizers needed on the space station and users have to be capable of generating frequencies over a wide bandwidth with small settling time at every new frequency. This is a fundamental feature needed in any frequency hopped system. It becomes more difficult and costly as the hopping rate is increased. Unfortunately, a high hopping rate is essential if the frequency hopped system is to provide any appreciable immunity to interference. To fully comprehend this inescapable tradeoff in a system with bounded bandwidth resources, we discuss briefly in what follows the anti-interference capability of an FHMA system.

The anti-interference capability of an FHMA system operating in the space station/users environment can be characterized by the so-called J/S ratio. J/S stands for jamming (here actually interference from other users) power divided by the signal power of the intended user. From the military communication literature J/S is called the jamming to signal margin. In the space station multiple access scenario the interfering power J can be assumed spread over W the entire RF system bandwidth; and J/S can actually be taken as the number of interfering users or sources, of power equal to S, that can be accomodated by the FHMA system. The J/S margin is given by [1]

$$J/S = \left(\frac{2P_e}{M}\right)^{R_c/R_h k}\left(\frac{W}{R_h}\right) \qquad (6)$$

where R_h is the hopping rate (equal to the symbol rate R_s times the number of hops per symbol m), R_c is the PN chip rate, M is the symbol alphabet size referred to in MFSK, $k = \log_2 M$ and P_e is the probability of a chip error. For given values of the parameters M, P_e and R_c (6)

indicates that J/S increases linearly with the available bandwidth W, and more rapidly with the hopping rate R_h. The maximum available bandwidth is fixed. For example, 100 MHz on a return link. For the lowest return link rate of 100 kbps (telemetry cahnnel), and typical choices of M = 16, R_c = 128 kcps and rate 1/2 coding; we have R_s = 50 ksps, K = $\log_2 16$ = 4 and we get: J/S = 12.5 dB for m = 1 hop per symbol; J/S = 19.7 dB for m = 2; and 21.8 dB for m = 4 hops per symbol. Interpreting J/S as the number of users accomodated by the FHMA system, we find that with a hopping rate of 50 khps the system can support only 18 users. With a higher, and certainly significantly more expensive to produce, 100 khps the system can accomodate 93 users.

It has to be kept in mind that the above interference analysis is not really applicable to an intentional jammer. Such a jammer would not spread its power over the entire RF bandwidth W. If the FHMA system were to be resistant to such jamming, even more complex processing would be needed because of the limited available bandwidth.

From the brief description of the equipment needed for FHMA, and the results used above to illustrate the extent of interference rejection FHMA offers, it becomes apparent that FHMA is significantly more complex and expensive than, for example, FDMA if the sole purpose of the multiple access system is to separate between users communicating simultaneously. FHMA does not look attractive unless there is a requirement for jamming mitigation, and there is a willingness to support expensive frequency synthesizers that would hop at sufficiently fast rates to provide for such jamming resistance.

The complexity of FHMA systems, however, provide some advantageous features. First, it can support a large number of users if fast hopping (like 100 khps) is used. Secondly, coarse ranging can be provided easily by synchronizing the PN sequences at the transmit and receive ends. Thirdly, RFI interference rejection is inherent. Finally, the addition of users does not, if the system capacity is not exceeded, degraded performance appreciably. However, as with the other multiple access systems, the antenna systems would have to be upgraded. Also additional frequency synthesizers and PN generators would be needed to support the extra users.

REFERENCES

[1] "Trade Study for Space Station Multiple Access System," Presented to NASA-JSC by Lockheed Engineering and Management Services Company, June 11, 1984.

[2] R. M. Gagliardi, Satellite Communications, Lifetime Learning Publications, Belmont, California, 1984.

[3] P. Shaft, "Limiting of Signals and Its Effect on Communication," IEEE Trans. Comm. Tech., Vol. COM-13, December 1965, pp. 504-512.

[4] C. L. Weber, et al., "Performance Considerations of Code Division Multiple-Access Systems," IEEE Trans. on Vehicular Technology, February 1982, pp. 3-10.

APPENDIX J

DIFFERENT OPTIONS FOR THE LINKS BETWEEN THE ANTENNAS AND THE COMMUNICATION PROCESSOR

The space station (SS) will have a large platform that extends for more than 100 meters from end to end. The SS will also communicate with several users through a number of antenna systems located at different parts of the platform. At present it is envisioned that down-conversion from RF to IF will take place at the antenna units and that a distribution system will link the antenna units to a central communication processor. The communication processor will be located in a module which is close to one end of the SS platform. As a result, distances in excess of 100 meters may separate some of the antennas from the processor. Obviously, distribution of the signals using waveguides at RF entails severe and unacceptable losses for such distances. The conventional solution is the use of coaxial cables to carry the IF signals from the down-converters to the communication processor.

The SS free-space environment is a harsh one where cosmic radiation and charged particles bombard the platform. Coaxial cables provide a certain degree of resistence to electromagnetic interference. However, the cables themselves may in time deteriorate due to the severe environmental conditions. The integrity of the information carried on them may therefore be jeopardized. Since high reliability is desired, redundancy is generally needed to provide error protection measures that insure the integrity of the information transfer. The need for redundancy, together with superior cable shielding and low attenuation requirements at IF frequencies, lead to high overall cable weights. Typically, a coaxial cable of attenuation between 0.2 and 0.5 dB/m at 300 MHz (i.e., 20 to 50 dB attenuation over a 100 m run) has a diameter between 1/3 and 1 inch and weighs up to 1 Kg/m [1]. Lower attenuation is naturally achieved at the cost of larger size and heavier weight.

An alternative to the use of coaxial cables at IF is to do the data transfer at baseband. This would substantially reduce the cable weights since cables with low attenuation at a few megahertz are smaller and lighter than those operable at IF. This scheme, however, has the significant disadvantage of requiring a demodulator that demodulates down to baseband at every antenna location. This obviously entails additional cost and weight penalties.

An attractive alternative to the above approaches is to transfer the information between the antenna locations and the communication processor on optical fibers. The use of optical fibers on board the space station is attractive because of their light weight and immunity to electromagnetic radiation. Fiber-optic links, however, are suited for digital baseband transmission. As will be seen shortly, high bit rates can and have been achieved using intensity modulation of the light source. Intensity modulation is the most commonly used method of modulating the light sources, which are lasers or light-emitting diodes (LED's). Transmitter nonlinearities, fiber AM/PM conversion, as well as other subtle source/fiber interactions (like modal noise and distortions) make intensity modulation with pulses the scheme that best utilizes the broad bandwidth potential of fiber-optic links. The nonlinear effects mentioned above severely limit the bandwidth available for other modulation techniques such as direct analog intensity modulation.

Fig. 1 [2] illustrates the power characteristics of the different light sources and receivers that can be used for fiber-optic links. LED's emit less power than semi-conductor lasers but tend to be cheaper, simpler and more reliable. Si-APD's (avalanche photo-diodes) have more sensitivity than p-i-n-FET receivers (p-i-n photodiodes combined with low-noise FET amplifiers). Si-APD's, however, are more costly and nonlinear than p-i-n-FET receivers.

Several first generation systems (up till 1980) have been installed in commercial applications [3]. These systems utilized light sources at wavelengths around 0.85 μm operating at the dispersion minimum of multi-mode graded index fibers. Graded-index fibers have attenuations no less than 3 dB/Km [2]. Si-APD's were used to offset the cable loss since much higher link margins are needed in optical links compared to electrical links. As an example, we mention the system installed in New York City [4] which had a capacity of 450 Mbps$\cdot$Km and used a GaAlAs laser source and a standard Si-APD receiver.

Second generation systems that have been demonstrated since 1981 [2,5] fall into two categories: those using multimode fibers and operating at the material dispersion minimum around the longer wavelength of 1.3 μm, and those using the monomode fibers at one of the wavelengths of minimum attenuation. Both offer significantly higher capacities than first generation systems. At the longer wavelength of 1.3 μm cables having attenuation of only 1 dB/Km can be routinely installed. At these wavelengths, though, APD's have high noise factors

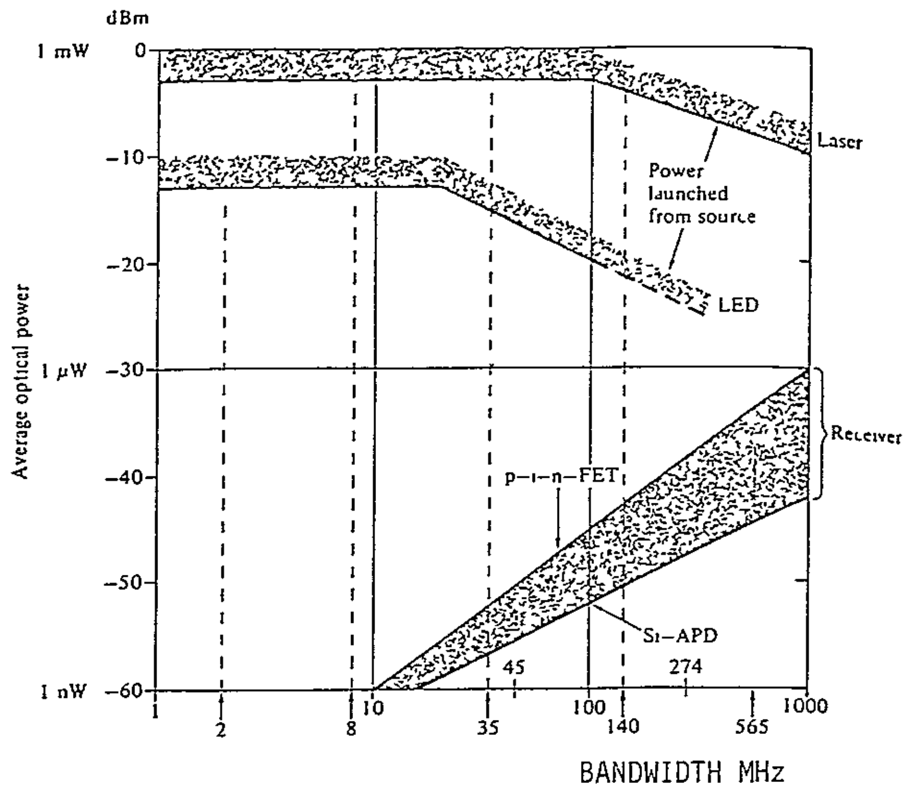


Fig. 1. Different Optical Source Powers Launched onto a Fiber and Ranges of Receiver Sensitivities.

and offer little advantage over p-i-n detectors. 1.5 Gbps·Km capacities with the use of LED sources have been reported [2]. This involves an increase by a factor of about 10 over first generation systems using LED sources. With InGaAsP/InP laser diodes, a capacity of 8 Gbps·Km (400 Mbps x 20 Km) has actually been demonstrated in the field [5] with single-mode fibers at 1.3 μm wavelength. An experimental system capable of handling pulse rates of 1.6 Gbps and a 20 Km range has also been reported in [5]. Among the reported factors that limit even further increase in data rates is the instabilities of the transversal and longitudinal single-mode oscillators of the laser diodes.

All the above systems demonstrate the superb bandwidth and capacity capabilities expected of fiber-optic links. One major problem in using a system as those described above for the SS application at hand is the need to provide the input signal in the form of baseband pulses. As mentioned earlier, this would require costly demodulation at each of the antenna sites. It would be far more advantageous if the IF signals at the output of the down-converters can be directly modulated onto the optical fibers. It has been mentioned [2,6] that intensity modulation by a frequency-modulated subcarrier is possible. This would be similar to what is best needed for the SS. This actually represents a compromise between intensity modulation with strictly digital and analog input signals, and has been accomplished on free space links using large laser sources. The feasibility of this approach on optical fibers is yet to be fully determined at this stage.

An alternate modulation technique that has also been mentioned [2] is pulse position (or frequency) modulation. A number of inherently nonlinear processes, particularly with LED sources, limit the utility of this approach.

It has been reported also [5] that recent advances in fiber manufacturing techniques have led to optical losses less than 0.5 dB/Km over a range extending from 1.2 to 1.6 μm without interruption. This has made feasible the technique of wavelength division multiplexing (WDM) (which is basically similar to FSK). Experimentally [7], a two-way WDM link of 100 Mbps signals was demonstrated over a 36 Km single-mode fiber using the two wavelengths 1.29 and 1.51 μm.

Finally, it should be mentioned that significant bandwidths can still be achieved on fibers with strictly analog intensity modulation. A prototype system has been demonstrated [8] for the nuclear testing application with a 3 dB bandwidth extending up to 200 MHz over 1 Km length. The nonlinear distortion is maintained at 1% up to 100 MHz and 3% up to 200 MHz. Careful high frequency equalization is used to compensate for the high frequency rolloff due to fiber delay distortion over the length of the cable. A similar system can presumably be used with the SS if the IF can be placed say around 100 MHz and the IF bandwidth is less than 200 MHz.

In conclusion, a careful cost/benefit tradeoff should be performed to determine which of the different options is best suited for the SS application. The options that have been pointed out here are: (1) coaxial cables carrying the IF signals from the down-converters at the antennas to the central communication processor; (2) simpler electrical cables conveying the signals to the processor at baseband; (3) wideband optical fibers pulse intensity modulated at baseband; (4) optical fibers directly modulated in intensity or otherwise at a suitably chosen IF within the distortion-limited bandwidth of the fiber.

REFERENCES

[1] Reference Data for Ratio Engineers, Howard Sams & Co., New York, 1975, Chapter 24.

[2] J. Gowar, Optical Communication Systems, Prentice-Hall, London, 1984.

[3] J. Midwinter, "First Generation Trunk Transmission Systems: Capabilities and Limitations," IEEE J. on Selected Areas in Comm., April 1983, pp. 381-386.

[4] I. Jacobs and J. Stauffer, "FT-3-A Metropolitan Trunk Lightwave System," Proc. IEEE, October 1980, pp. 1286-1290.

[5] K. Nakagawa, "Second Generation Trunk Transmission Technology," IEEE J. on Selected Areas in Comm., April 1983, pp. 387-393.

[6] R. Gagliardi and S. Karp, Optical Communications, John Wiley, New York, 1976.

[7] E. Yanai, et al., "Two-way Wavelength Division Multiplexing Transmission Experiment Using Single Mode Fibers," Paper Tech. Group Comm. Syst., IECE, Japan, CS81-162, January 1982.

[8] P. Lyons, et al., "Applications of Optical Fibers to Analog Telemetry Delay Lines and Sensing Systems," IEEE J. on Selected Areas in Comm., April 1983, pp. 555-561.

APPENDIX K

SPACE STATION MULTIPLE ACCESS SYSTEM CONCEPTUAL DESIGN

LinCom

AGENDA

- MA SYSTEM REQUIREMENTS
- KEY SYSTEM ASSUMPTIONS
  - ANTENNAS
  - POWER AMPLIFIERS
  - LOW NOISE AMPLIFIERS
- FREQUENCY SELECTION
  - FCC ALLOCATION
  - POWER FLUX DENSITY
- CODING/MODULATION
  - CODING
  - FADING/MULTIPATH
- LINK BUDGET
- MA SCHEMES
  - RANGING
- SIGNAL DESIGN AND FREQUENCY PLAN
  - INTERFERENCE
  - CHANNEL SEPARATION
- STRAWMAN SYSTEM
  - MOD/DEMOD
  - IF DISTRIBUTION
  - ANTENNA SYSTEMS
  - RISK ASSESSMENT

LinCom

LinCom

MA SYSTEM REQUIREMENTS

LinCom

LinCom

SPACE STATION MULTIPLE ACCESS SYSTEM

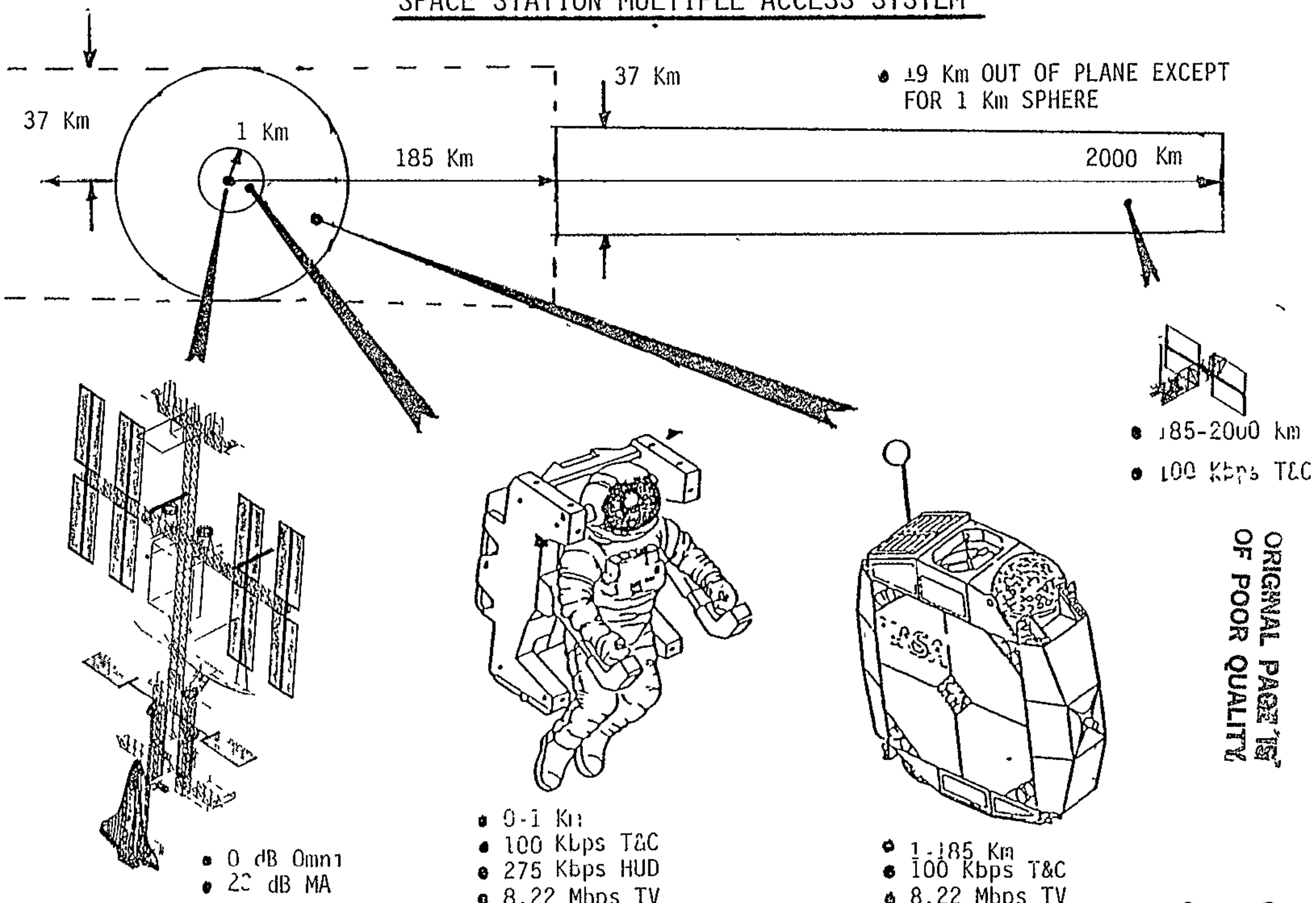
The space station multiple access system supports communications between the Space Station (SS) and the Extra-Vehicular Activity (EVA), the Orbit Transfer Vehicle (OTV), the Orbit Maneuvering Vehicle (OMV), Free Flyers (FF), and a co-orbitng platform. EVA support is for a 1 km sphere around the Space Station. OMV/OTV support is primarily around a disc of 37 km radius, +/- 9 km thick. There is also a requirement to communicate within a region defined by the 185 km rectangle, also +/- 9 km thick. The rectangle is actually a curvilinear slab since the horizontal lines represent sections of orbits parallel to the SS orbit. The FF/co-orbiting platform support is between 185 to 2000 km, also +/- 9 km thick. Only regions trailing the SS are shown. Regions leading the SS are defined identically.

The basic assumption for this study is that the SS antenna system consists of 0 db omnidirectional coverage antennas as well as a Multiple Access (MA) antenna system capable of supporting simultaneouly 5-6 users and providing 23 db gains.

LinCom

LinCom

SPACE STATION MULTIPLE ACCESS SYSTEM



LinCom

LinCom

SS MULTIPLE ACCESS REQUIREMENTS

The MA system consists of proximity operations and FF/platform links. The FF links are only for low data rate command and telemetry while the proximity operations also involve high data rate TV links. The primary purpose of the 0 db omni is to cover the EVA at close range where pointing the high gain MA antenna is impractical. Note that EMU stands for Extravehicular Manuevering Unit, used here interchangeably with EVA.

LinCom

LinCom

SS MULTIPLE ACCESS REQUIREMENTS

MULTIPLE ACCESS

- PROXIMITY OPERATIONS
  - EMU
    - 1 Km Sphere
    - 0/23 dB Gain Antenna
  - OTV/OMV
    - 37 Km disc*
    - 185Km slab
    - 23 dB Gain Antenna
- FREE FLYERS/PLATFORM
  - 185 Km to 2000 Km
  - 23 dB Gain Antenna

*37 Km radius, ±9 Km thick

LinCom

LinCom

FORWARD LINKS

Two more sets of voice & command and Head Up Display (HUD) links are included to account for the channel allocation for the EMU-Orbiter links, which can operate simultaneously with the EMU-SS links. The 100 kbps date rate shown is the maximum data rate to be accomodated.

LinCom

FORWARD LINKS

USER	LINK TYPE	DATA RATE	# OF LINKS IOC (GROWTH)
EMU*	VOICE & COMMAND	100 Kbps	4(6)
	HUD	275 Kbps	4(6)
	TV	22 Mbps	2
OMV	COMMAND	100 Kbps	1(2)
OTV	COMMAND	100 Kbps	0(1)
FREE FLYER	COMMAND	100 Kbps	4(8)
PLATFORM	COMMAND	100 Kbps	1(1)

*Includes 2 EMU/Orbiter Links--for channel allocation purpose only.

RETURN LINKS

Although both 8 and 22 Mbps TV channels are available, each user can only use one TV channel at a time. Hence the maximum number of simultanoeus TV links used can not exceed the number of users. Two sets of voice & telemetry, and 8 Mbps TV channel are reserved for the EMU-Orbiter.

RETURN LINKS

USER	LINK TYPE	DATA RATE	# OF LINKS IOC (GROWTH)
EMU*	VOICE & TELEMETRY	100 Kbps	4(6)
	TV[1]	8 Mbps	4(6)
	TV[1]	22 Mbps	2(4)
OMV	TELEMETRY	100 Kbps	1(2)
	TV[2]	8 Mbps	1(2)
	TV[2]	22 Mbps	1(2)
OTV	TELEMETRY	100 Kbps	0(1)
	TV[3]	8 Mbps	0(1)
	TV[3]	22 Mbps	0(1)
FREE FLYER	TELEMETRY	100 Kbps	4(8)
PLATFORM	TELEMETRY	100 Kbps	1(1)

[1] Simultaneous TV links will not exceed 4(6).

[2] Simultaneous links will not exceed 1(2).

[3] Simultaneous TV links will not exceed 0(1).

LinCom

MA LINKS SUMMARY

It is evident that the frequency specrtrum requirement for the MA system is dictated by the TV links.

LinCom

LinCom

MA LINKS SUMMARY

- FORWARD LINKS

DATA RATE	IOC (GROWTH)
100 Kbps	10(18)
275 Kbps	4(6)
22 Mbps	2

- RETURN LINKS

DATA RATE	IOC (GROWTH)
100 Kbps	10(18)
8 Mbps*	5(9)
22 Mbps*	3(7)

*Sum of TV links will not exceed 5(9).

LinCom

LinCom

KEY ASSUMPTIONS

The key driver for this study is the given baseline SS antenna system. In order to be cost effective, state of the art, commericially available microwave low noise amplifiers and power amplifiers are to be employed. This sets the noise figure for the receiver front end and the available transmit RF power.

LinCom

LinCom

KEY ASSLMPTIONS

LinCom

LinCom

ANTENNA SYSTEM

The baseline assumes that the MA hign gain (23db) antenna system serves all the proximity operations (prox ops) as well as the FF/platform links. That is a total of 8 users during IOC or 16 users during growth. Since the MA antenna system can serve only 5 or 6 users simultaneously, this means that the antenna beams must be assigned on a time-shared basis. To facilitate time-sharing, order-wire (OW) channels are used to allocate these beams to the users. Additional information will also be communicated through these channels to define the assigned link characteristics and perhaps provide some location information to assist antenna pointing. In order to lessen the load on the antenna system two sets of fixed beam antenna pairs, pointing forward and backward are recommended for the 5(9) FF/platform links. The fixed beam antennas are considerable cheaper than the MA antennas since no antenna acquisition and tracking system is required.

LinCom

LinCom

ANTENNA SYSTEM

- BASELINE
  - OMNI ANTENNAS (0 dB GAIN)
  - MULTI-BEAM ANTENNAS (23 dB GAIN)
    - 5-6 SIMULTANEOUS USERS
- RECOMMENDATION
  - SEPARATE FIXED BEAM ANTENNAS FOR FREE FLYERS (2 ON EACH SIDE)

LinCom

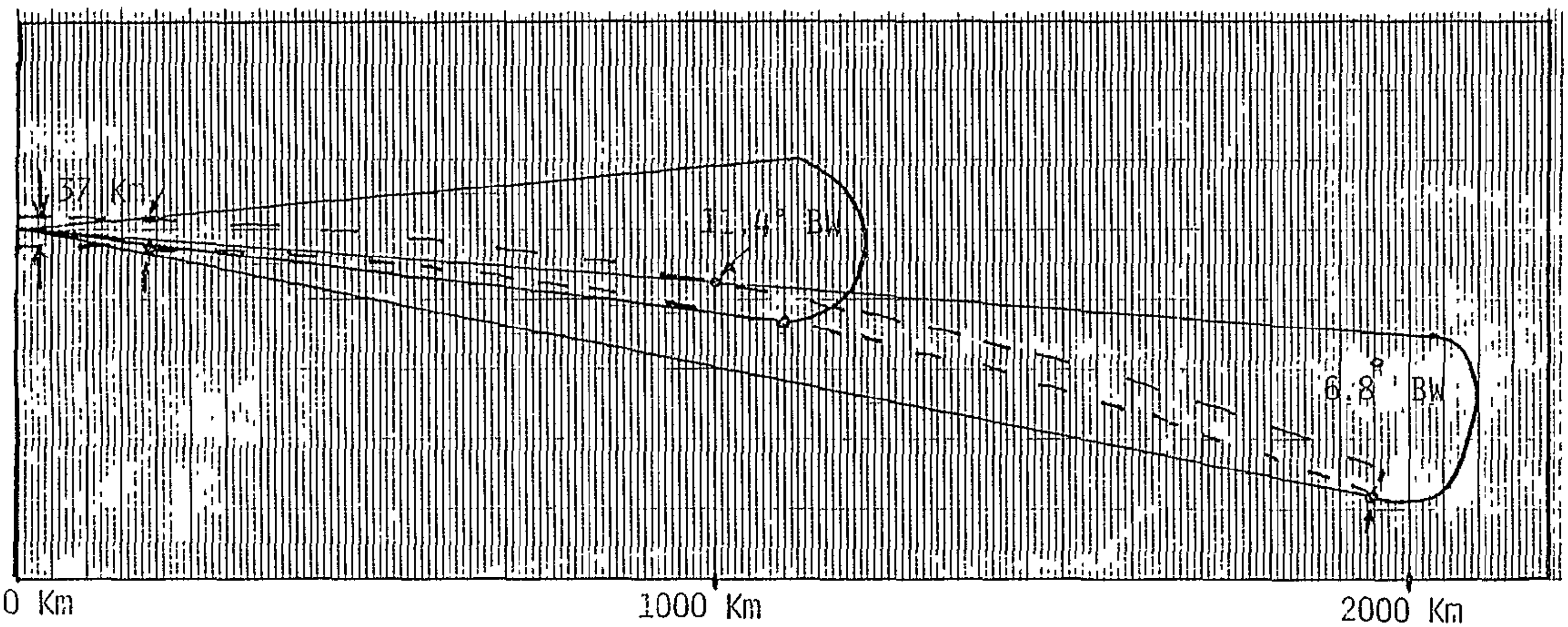
FIXED BEAM ANTENNAS FOR FREE FLYERS

This chart is drawn to scale to show how the curvillinear slab appears in Cartesean co-ordinate. The SS orbit altitude is assumed to be 200 km and an earth radius of 5370 km is used. The first antenna is chosen to cover the near edge at 185 km and the required 3 db beamwidth is 11.4 deg. The second antenna must be chosen so that the combined antenna pattern covers the whole region. The 6.8 deg antenna shown covers the range from 1000 to 2000 km. Notice there is an overlapped region.

At K-band, the diameters of the antennas are on the order of 6 and 12 inches. Since the beamwidths are rather wide, many types of antenna such as corrugated horns, lense antennas, or parabolic dishes can be used. They are commericially available and many are off the shelf items.

LinCom

FIXED BEAM ANTENNAS FOR FREE FLYERS



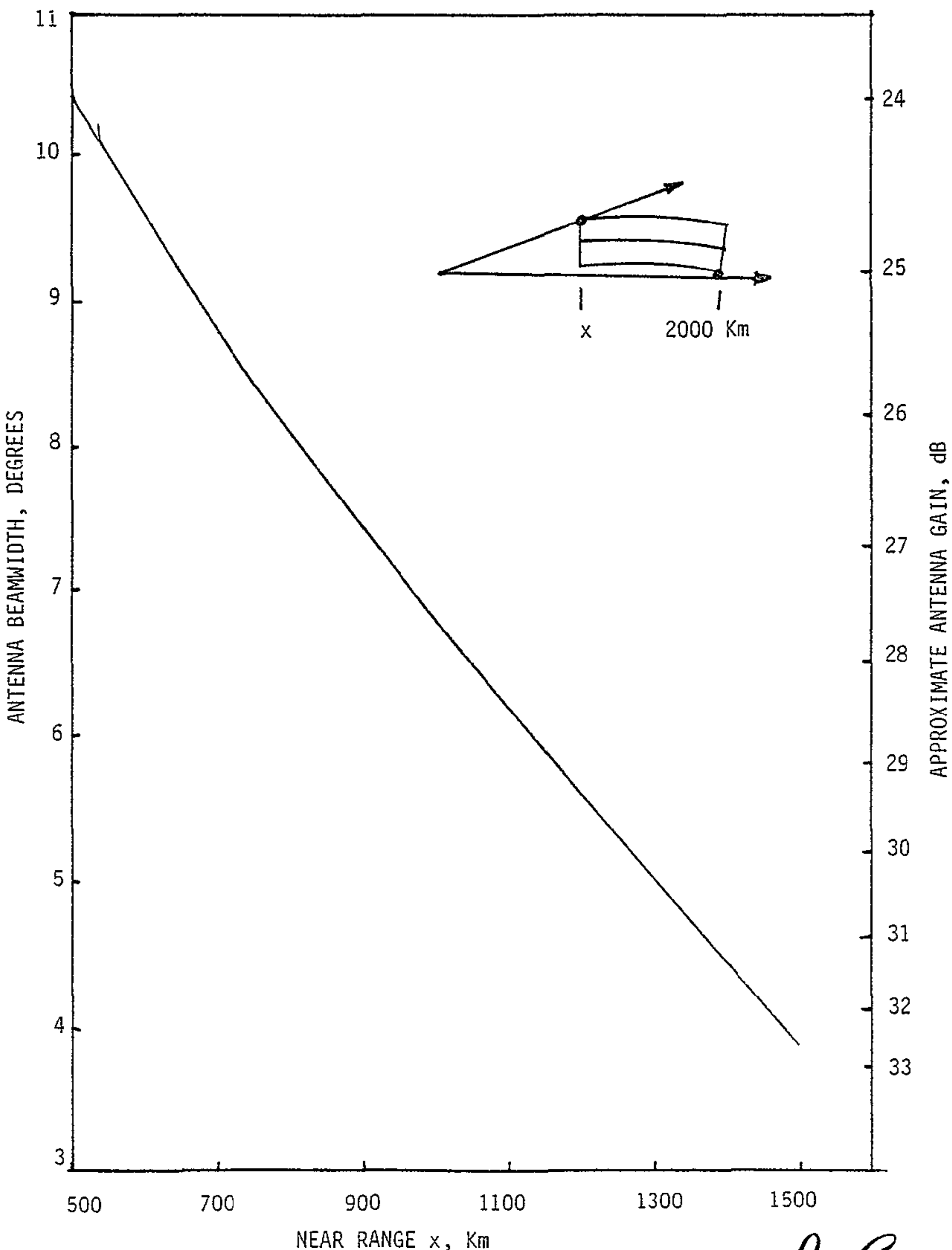
• 6 - 12 INCHES IN DIAMETER

LinCom

FAR-END FREE FLYER COVERAGE TRADEOFF

The coverage of the second beam can be adjusted to trade coverage overlap against antenna gain. For the crossover range of 1000 km chosen, the anteanna gain is approximately 28 db.

FAR-END FREE FLYER COVERAGE TRADEOFF



LinCom

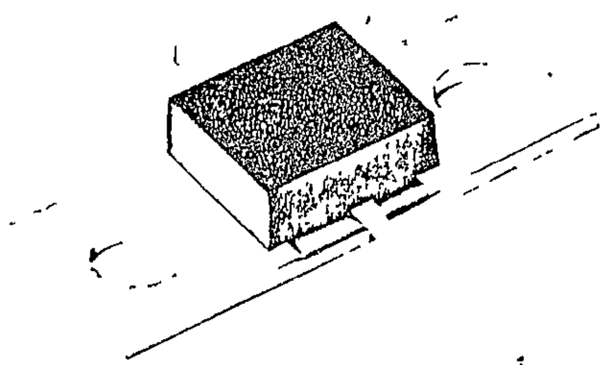
COMMERCIALLY AVAILABLE GaAs POWER FET CHIPS

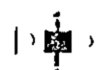
This is an example of commercially available power amplifiers at 15 Ghz. Many other vendors have similar products, e.g. Avantec, Harris, NEC, etc. The power output is on the order of 1 watt (30 dbm).

LinCom

LinCom

COMMERCIALLY AVAILABLE GaAs POWER FET CHIPS





Actual Size

FEATURES

1½ Watts Output

2 GHz 1 dB Bandwidth

24% Efficiency

50 ohms in/out

DESCRIPTION

Hughes model number C2421H-1500 and C2422H-1500 are single cell and dual cell 15 GHz broadband GaAs power FET chips mounted on internally matched chip carriers The passivated chips are protected against mechanical damage by a plastic cap The devices are guaranteed to operate in a 50 ohm system with no additional matching required by the user over a typically 2 GHz bandwidth The chip carriers are compatible with 25 mil (0 635 mm) alumina mircrostrip circuits and have 25 x 1 mil (0 635 x 0 0254 mm) gold ribbon leads The dual cell device is rated at 1¼ Watts minimum output power, and the single cell device is rated at ½ Watt minimum The devices can be cascaded directly for multi-stage amplifier applications and/or combined with quadrature couplers to achieve higher power levels

The matched transistors are 100% DC and RF tested with the measured values of power output, gain, power added efficiency, and the associated bias conditions supplied with each unit The devices are RF tested in an alumina microstrip test circuit with precision APC-7 to microstrip transitions This test fixture, Hughes model number Z1010H-1500, is available for sale separately

The 0 8 micron aluminum gate devices are 100% visually inspected at 1000X magnification Devices are screened by rejecting those with excessive change in DC characteristics after a 72-hour high temperature bake

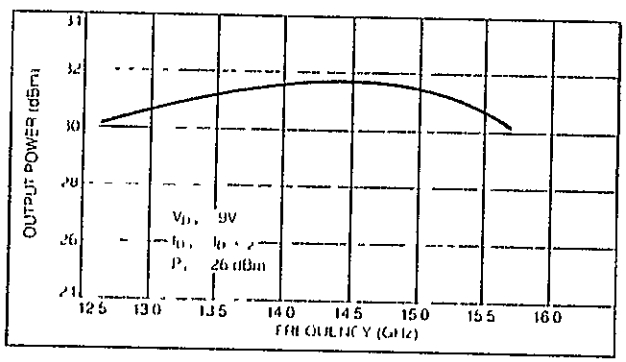
LinCom

GaAs FET PERFORMANCE

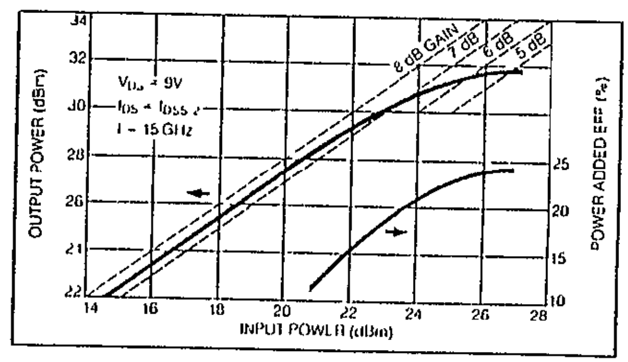
Note the wide bandwidth and linearity with a few db backoff.

GaAs FET PERFORMANCE

OUTPUT POWER VS FREQUENCY
(Model C2422H-1500 typical)



OUTPUT POWER AND EFFICIENCY VS INPUT POWER
(Model C2422H-1500 typical)



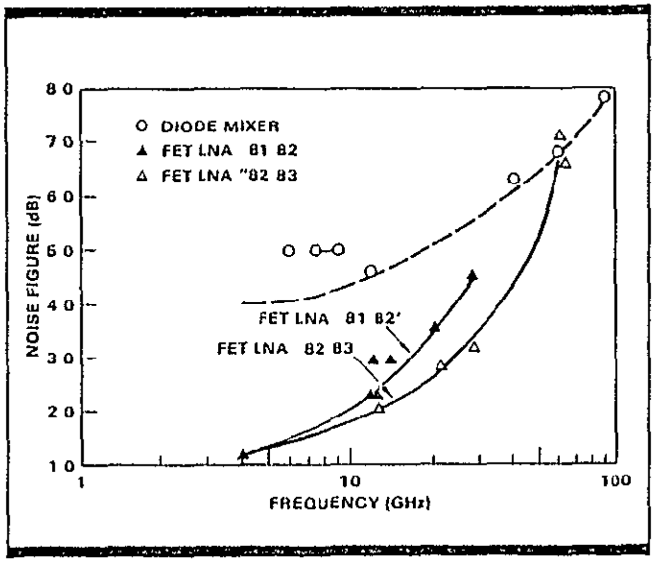
LinCom

FET LNA PERFORMANCE

This is a summary of typical FET low noise amplifier noise figure performance. Again, there are many vendors, perhaps more so than power FETs since there is a market for direct broadcast TV reception at the K-band frequencies. Typical noise figures range from 1.5 to 2.5 db.

LinCom

FET LNA PERFORMANCE



1 Comparison of mixer/IF amplifier and FET LNA performance vs frequency at microwave and mm-wave regions Below 18 GHz, FET amps dominate the high-performance, low-noise market

LinCom

FREQUENCY ALLOCATION AND REGULATORY LIMIT

Based on the total MA data rate requirement - close to a few hundred Mhz, the lowest frequency slot capable of handling this requirement is at K-band. Even if the requirement is reduced to the order of one hundred Mhz, the RFI problem makes S-band unattractive. The next set of charts addresses the NTIA/FCC band allocations at these frequencies, the associated regulatory limit on power flux density impringing on earth, and the proposed frequency plan for the MA system.

LinCom

LinCom

FREQUENCY ALLOCATION AND REGULATORY LIMIT

LinCom

INTER-SATELLITE BANDS (Ku-BAND)

The allocated service relevant to the MA system is under space research. Note that they are all secondary allocations. The primary allocations are given in capital letters. In case of conflict, the secondary allocation users must yield to the primary users. Note that there are also other secondary allocations within these bands to contend with. Radiolocation usually involves the use of radar, a potential source of RFI. Fixed services involve fiexed point-to-point communications such as commercial satellite services.

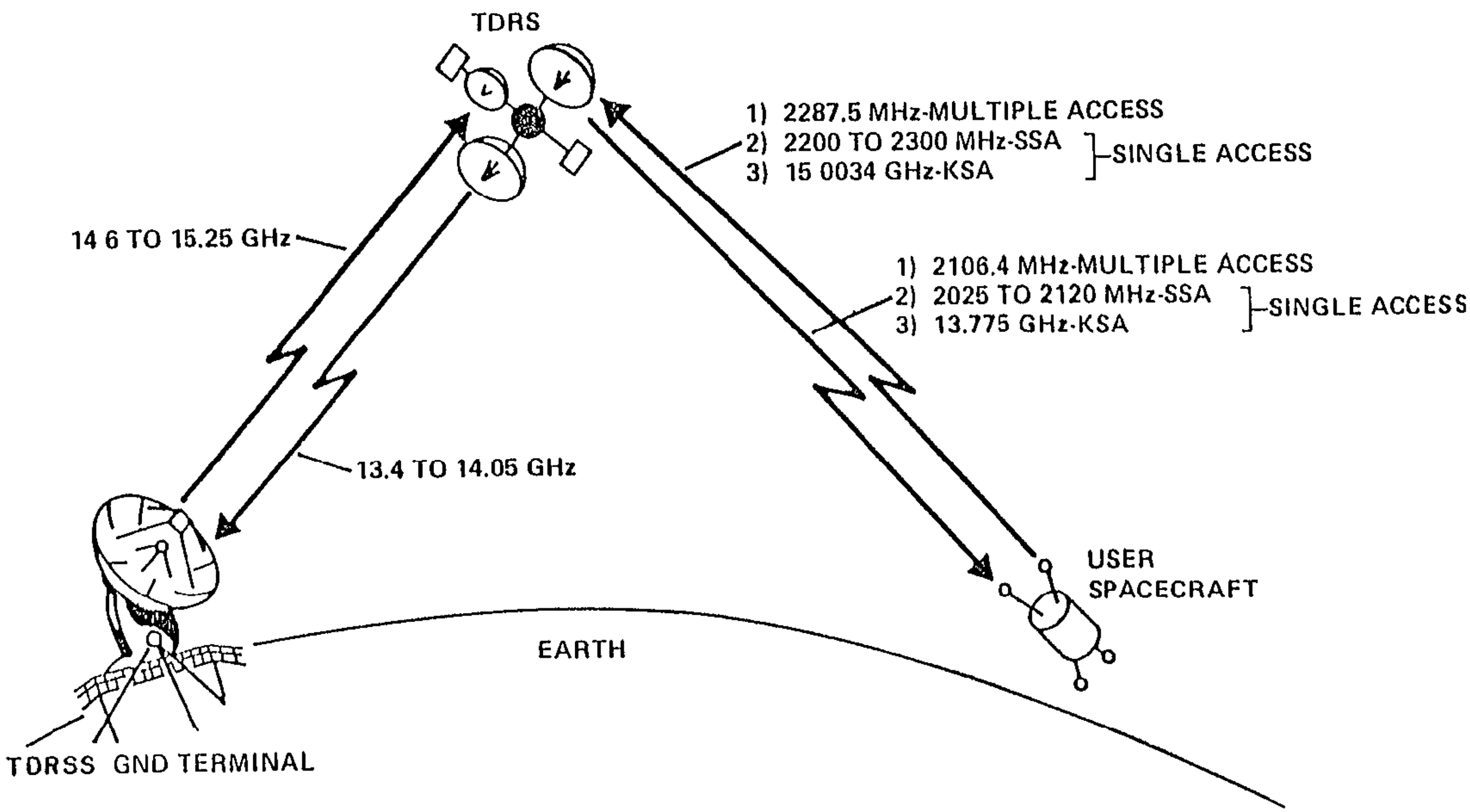
INTER-SATELLITE BANDS (Ku-BAND)

BANDS (GHz)	ALLOCATED SERVICES RELEVANT TO MA	OTHER ALLOCATED SERVICES	COMMENTS
13.4 - 14.0	Space Research	RADIOLOCATION	Not all data services authorized
		Standard Frequency & Time Signal Satellite (Earth-to-Space)	Secondary allocation only
14.0 - 14.2	Space Research	FIXED-SATELLITE (Earth-to-Space)	Not all data services authorized
		RADIONAVIGATION	Secondary allocation only
14.5 - 14.7145	Space Research	FIXED Mobile	Not all data services authorized Secondary allocation only
14.7145 - 15.1365	Space Research	FIXED Mobile	Not all data services authorized Secondary allocation only
15.1365 - 15.34	Space Research	FIXED Mobile	Not all data services authorized Secondary allocation only Passive sensing in this band Radio astronomy in upper adjacent band

TDRSS FREQUENCY PLAN

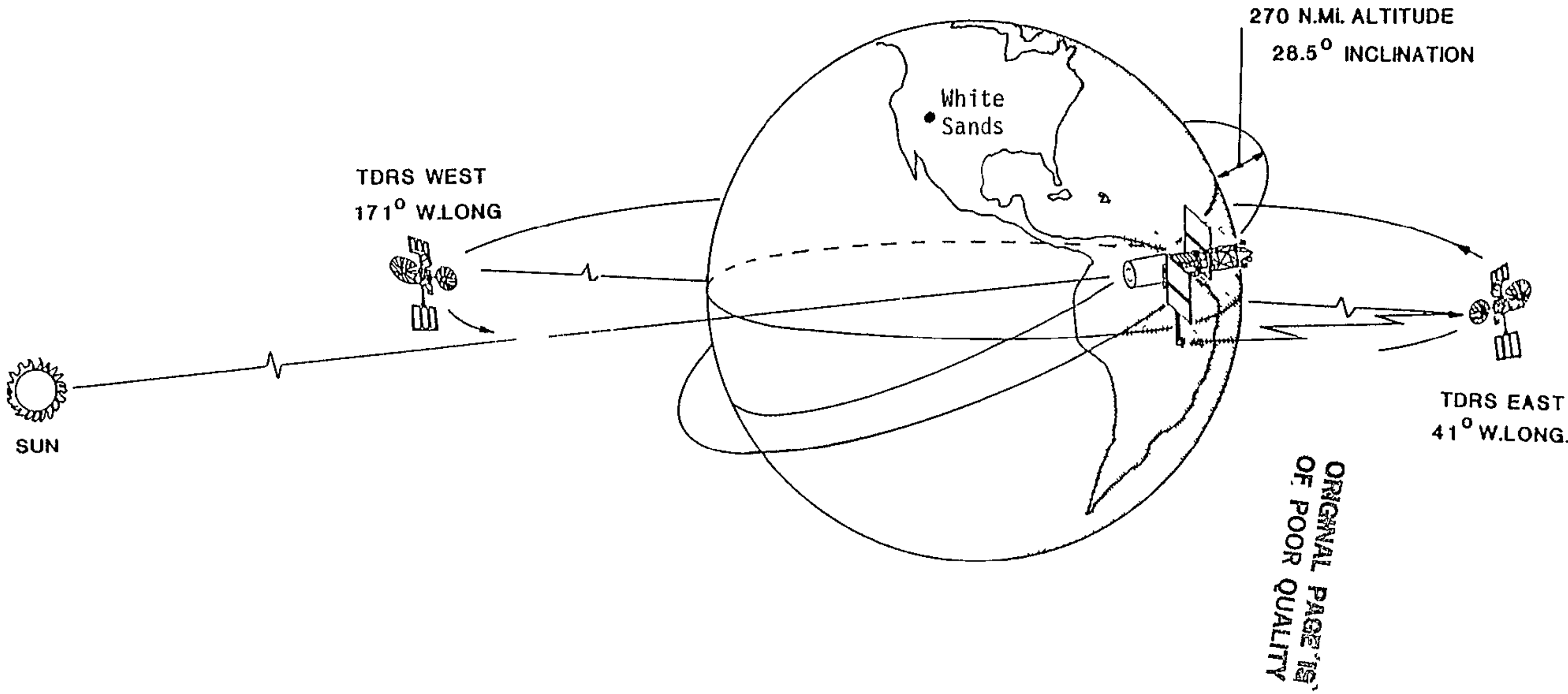
The TDRSS uses these frequencies. It is easier for the SS MA system to co-ordinate its frequency plan with the TDRSS, a single government user (controlled by NTIA) than with many non-government users (controlled by FCC).

TDRSS FREQUENCY PLAN



ADVANCED SOLAR OBSERVATORY CO-ORBITER PAYLOAD PLATFORM

This is taken from the SS reference configuration description document and shows the SS orbit. (The SS has the same orbit as the co-orbiting platform.) The TDRS locations are not drawn to scale and are actually at a distance about five times the earth radius away. However it is obvious even from this chart that the portion of the SS orbit that can be potentially interfered by the TDRS - White Sands space ground links (SGL) is very small if at all, considering that both the TDRS and the ground station use narrow beamwidth antennas. Hence the TDRSS SGL frequencies can be used safely by the SS MA system.



ADVANCED SOLAR OBSERVATORY CO-ORBITER PAYLOAD PLATFORM

LinCom

CANDIDATE FREQUENCY PLAN

The TDRSS KSA forward and return frequencies are off limit to the MA system. The SS must use these frequencies to communicate with the TDRSS. The proposed slots are shown shaded. The slot between 13.4 to 13.72 GHz is the TDRSS ground to space link and can also be considered.

LinCom

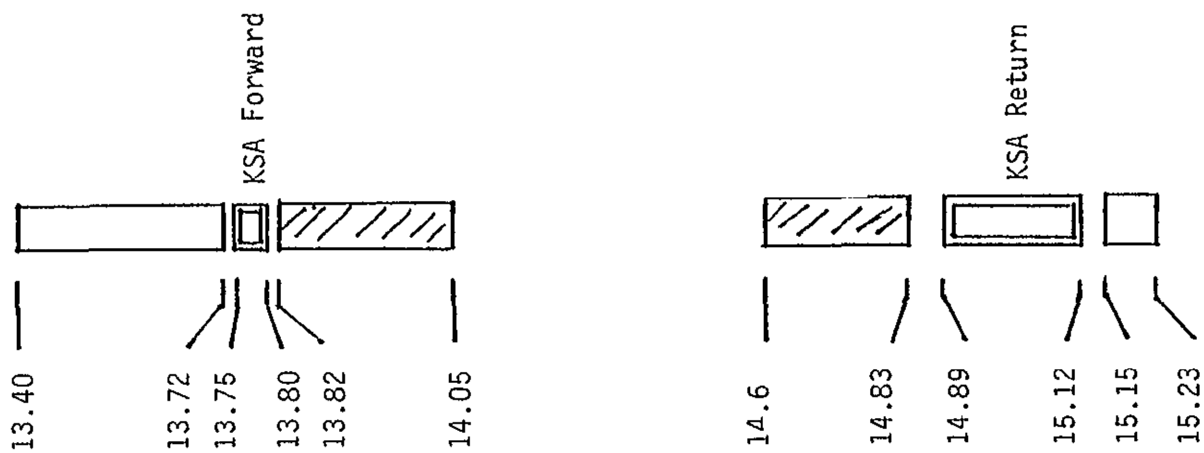
CANDIDATE FREQUENCY PLAN

FORWARD: 13.82 - 14.05 GHz 225 MHz

(SGL-KSA RETURN)

RETURN: 14.60 - 14.83 GHz 230 MHz

(SGL-KSA, SSA, MA FORWARD, TDRS COMMAND)



POWER FLUX DENSITY RESTRICTIONS

The next set of charts addresses the power flux density (PFD) issue at K-band. The user links likely to exceed the PFD limit are identified and options to control the PFD are discussed.

LinCom

POWER FLUX DENSITY RESTRICTIONS

- K-BAND FLUX DENSITY REGULATORY LIMITS
- WORST CASE SCENARIO
- IMPACT ON SS MA LINKS
    - PROXIMITY OPERATION (OMV/OTV)
    - FREE FLYER

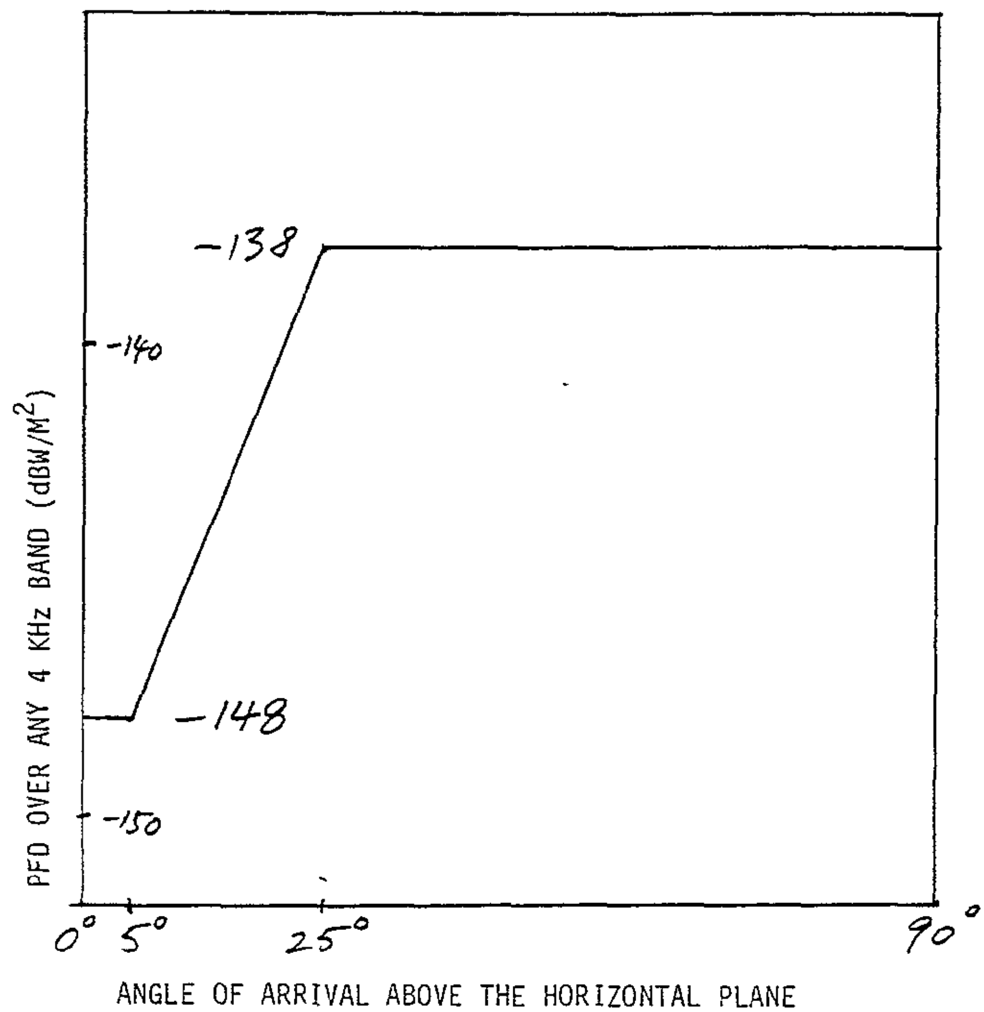
LinCom

K-BAND POWER FLUX DENSITY (PFD) RESTRICTION AT THE EARTH'S SURFACE

The PFD is measured over any 4 KHz band. For a given transmit power, a signal can be spread over a wider bandwith to reduce the PFD.

K-BAND POWER FLUX DENSITY (PFD) RESTRICTION

AT THE EARTH'S SURFACE

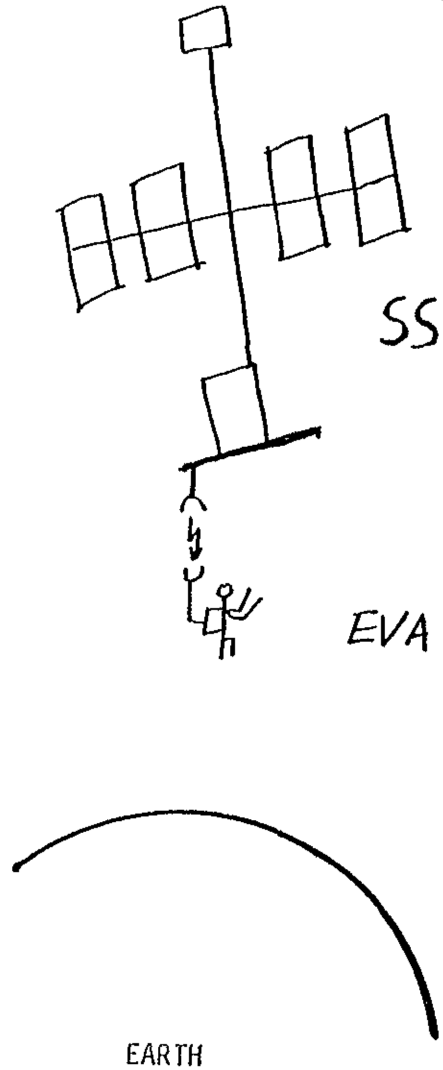


WORST CASE PFD SCENARIO

The MA high gain (23db) antenna would present more of a problem than the omni antenna for the same transmit power. However, since the data rate is also a factor here (affecting the distribution of PFD over a fixed frequency range), the situation is not so straightforward.

The worst case scenario involving the fixed beam antennas for the FF links are even more complicated since a few competing factors are involved. They have higher peak gains, yet the antenna will not be pointed directly toward the earth so the actual gain in the direction of earth is smaller. The slant range to the earth is higher, yet the PFD limit is more stringent because of the shape of the specification.

WORST CASE PFD SCENARIO



- 200 KM ORBIT
- WORST CASE WHEN EVA IS DIRECTLY ABOVE SUBSATELLITE POINT

TX EIRP, dBW	PFD, dBW/m^2
0	-117
23	-94

- SITUATION SIMILAR FOR PROXIMITY OPERATION

LinCom

OPTIONS TO ACCOMODATE PDF REQUIREMENT

The values used here are no longer current, however the chart is representative in that it shows the problematic links and the options available to control the PFD, if one decides to do so. There is always the alternative to ask for an exemption. In computing these numbers, a one watt transmitter is assumed. Also, the PFD is assumed to be uniformly distributed over a frequency range equal to the data rate. In cases where there are enough system margin, one can backoff the transmit power to satisfy the PFD requirement. If the system margin is not enough to cover the difference then the signal must be spread over a wider bandwidth. For low data rates, the signal can be spread over a large bandwidth by employing pseudo noise (PN) spreading. The decrease in PFD is proportional to the processing gain, i.e., the ratio of the chip to bit rate. For the TV links, the data rate is already high so that PN spreading becomes impractical. An alternative is to use low rate codes such as 1/n convolutional codes where n is 3 or higher. Since the overall bandwidth requirement is determined by the high rate TV links, the use of high rate codes is not desirable from a spectrum efficiency point of view.

LinCom

OPTIONS TO ACCOMODATE PFD REQUIREMENT

LINK	DATA RATE (Mbps)	MARGIN* (dB)	dB OVER PFD LIMIT	A	B	C	D
EVA (0 dB)	0.1	30	7		✓		
	0.275	26	2.5		✓		
	8	11	(-12)	✓			
	22	7	(-16)	✓			
EVA (23 dB)	0.1	43	30		✓		
	0.275	39	25.5		✓		
	8	24	11		✓		
	22	20	6.5		✓		
PROX OPS	0.1	11	30		✓		✓
	8	-8	11			✓	
	22	-12	6.5			✓	
FREE FLYER	0.1	-23.1	30**				✓

A = NO PFD COMPATIBILITY PROBLEM

B = POWER BACK-OFF

C = LOW RATE CODES

D = PN SPREADING

*MARGIN BASED ON SYSTEM DESCRIBED IN LinCom Memo TM-8411-3, 15 November 1984.

**6-10 dB LESS DEPENDING ON THE ANTENNA PATTERN.

LinCom

CODING/MODULATION TRADEOFF

Coding can potentially provide up to 5 or 6 db gains. There are however a number of problems associated with its use. The first major one is complexity which translates directly to size, weight and power. The second major one is bandwidth expansion. Both of which are concerns for the MA system. Modulation schemes can be picked to migitate problems associated with narrow bandwidth channels and HPA nonlinearity, none of which are of major concern presently. In the case of the MA system, the main driver for modulation selection is probably the potential multipath/fading problem assoicated with EVA operation near the SS structure using omnidirectional antennas.

LinCom

LinCom

CODING/MODULATION TRADEOFF

LinCom

SUMMARY

Codec chips (which solve the complexity problem) are good from a few hundred Kbps to a few Mbps. Hence they are particularly suited for the T&C links (100 Kbps). Another coding application which may be considered is the return links where the bulk of processing (decoding) is performed on the SS so that size, weight and power is less of a problem. If coding is employed for the TV links, then high rate codes are more desirable because of the bandwidth efficiency involved.

THe baseline MA system modulation selected is QPSK. This is a widely used scheme and its performance can be better or equal to some other more exotic schemes when the channel bandwidth is not severely limited and the channel is not highly nonlinear - envisioned currently for the MA system. If fading is of concern, then the other leading noncoherent candidate is perhaps FSK or differential PSK. Without multipath/fading, QPSK outperforms differential QPSK by 2.5 dB and uncoded quaternary FSK by 4.4 dB. When the multipath/fading loss is small, e.g., limited to 2 dB, the degradation relative to nonfading performance is similar so that QPSK still enjoys the advantage. Note that QPSK is a more bandwidth efficient modulation scheme.

LinCom

SUMMARY

- CODING
  - TO SIMPLIFY RECEIVER DESIGN, CODING IS AVOIDED WHERE LINK BUDGET ALLOWS
  - CODING IS USED FOR FREE FLYER T&C LINKS (BOTH FORWARD AND RETURN) BECAUSE CODEC CHIPS (e.g. HARRIS & TRW) ARE AVAILABLE -- CONVOLUTIONAL CODE, RATE 1/2, K = 7 WITH VITERBI DECODING
- MODULATION
  - QPSK IS USED FOR COST EFFECTIVENESS AND SIMPLICITY
  - IF A FADING LOSS OF 2 dB IS ASSUMED, PSK OUTPERFORMS FSK
  - IF MORE SEVERE FADING ENVIRONMENT IS ASSUMED, DIVERSITY IS REQUIRED

LinCom

LinCom

VLSI/VHSIC K=7 VITERBI CODEC

Note in particular the data rate, performance and power consumption. The TRW chip uses a different technology and is usable up to a few hundred Kbps.

LinCom

HARRIS CORPORATION
GOVERNMENT DATA COMMUNICATIONS DIVISION

VLSI/VHSIC K=7 VITERBI CODEC

PRELIMINARY SPECIFICATIONS

Data Rate	Up to 5Mb/s bit rate (10 Mb/s symbol rate)*
Performance	4.9 dB coding gain at 10^{-5} BER
Coding Polynomial	Rate 1/2, CL-7
Acquisition Time	TBD
Mean Time to Loss of Lock	TBD
Operating Modes	QPSK (dual input), BPSK (single input), both in continuous or burst mode (minimum 16 bit burst), with synchronous or asynchronous clocks (bit rate and trace-back clocks)
Input Data Formats	3 bits for each coded pair in either sign/magnitude or offset/binary, NRZ-L, -M, -S
Output Data Format	NRZ-L
Input Data/Clock Relationship	Accepts data changes at rising edge of clock
Output Data/Clock Relationship	Data changes at rising edge of clock; two output clocks available: one follows bit synchronization changes for true delay, the other is fixed for cripto data purposes
Branch Synchronization	Resolves swapped Ra & Rb coded pair for either QPSK or BPSK modes, plus swapped and staggered QPSK coded pair; accepts inverted/non-inverted Rb coded symbols
Throughput Delay	76 bits worst case (QPSK staggered)
Output Status	Lock, signal quality (4 Bits)
Testability	LSSD
IC Voltages	VDD = 3.3 ± 0.3 V, Ground
Interface	Low power schottky TTL with pull ups to chip's VDD (3.3 V)
Power Consumption	Less than 1/4 watts
Size	85 pins (package TBD)
Operating Temperature	-55° C to $+125^{\circ}$C
Rate 1/2, CL-7 Encoder	Included in chip with all features to match decoder

*Under total worst case conditions

PRELIMINARY ANALYSIS OF THE MULTIPATH PROBLEM FOR THE SPACE STATION

In this and the following charts, results of LinCom's preliminary analysis of the space station multipath problem are presented. The text accompanying the charts contains comments on the material presented therein, and serves as further explanation of that material.

The first chart outlines two important factors which influence the nature of the resulting multipath. The first is the proximity of the transmitting user to the space station platform (which is related to the type of user) and the second is the data rate (which depends on the type of link considered). This is explained in detail in the charts to follow.

LinCom

PRELIMINARY ANALYSIS OF THE MULTIPATH PROBLEM FOR THE SPACE STATION

GENERAL CONSIDERATIONS

- FOR EVA, A MULTITUDE OF REFLECTED PATHS MAY OCCUR (FIG. 1a). MANY PATHS WITH DIFFERENT ANGLES OF INCIDENCE ON REFLECTING SURFACES EXIST.
- FOR FREE-FLYER LINKS, REFLECTED PATHS WITH ONLY A FEW ANGLES OF INCIDENCE MAY BE CONSIDERED (FIG. 1b).
- ANALYSIS OF THE DATA LINKS WITH RATES > 1 Mbps SHOULD BE SEPARATE FROM THOSE LINKS WITH RATES < 1 Mbps, FOR REASONS RELATED TO THE RELATION BETWEEN POSSIBLE DIFFERENTIAL PATH DELAYS AND BIT DURATION.

LinCom

POSSIBLE MULTIPATH FROM TRANSMITTER IN NEAR ZONE

This chart depicts an EVA operating in the vicinity of the space station and transmitting to one of its receiving antennas. Due to the lack of directivity in the EVA's omni antenna pattern, rays of equal strength are transmitted towards the receiving antenna and other reflecting surfaces on the station. If the receiving antenna is also omni-directional, as in the case of the 100 Kbps T&C links, the main factor that attenuates the reflected rays compared to the direct path is the path length difference. It is easy to visualize situations where that difference is small, for example, when the EVA is close to the solar arrays. In those situations the signal fading due to multipath may be severe.

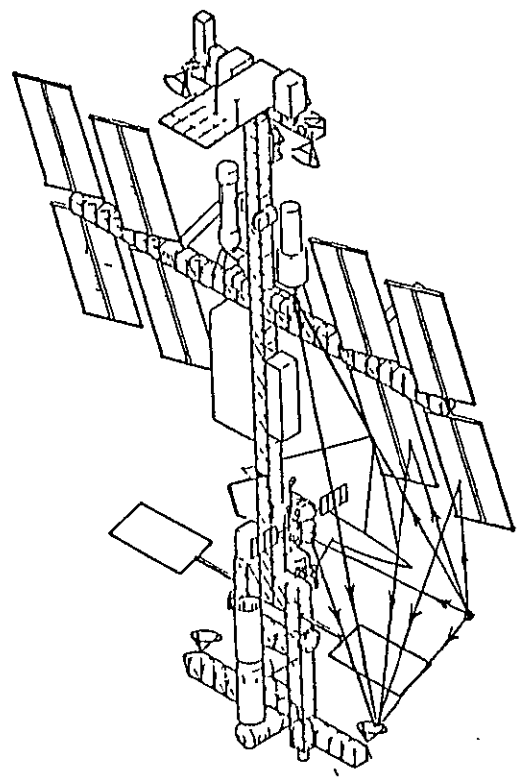


Fig. 1a. Possible Multipath From Transmitter in Near Zone.

POSSIBLE MULTIPATH FROM TRANSMITTER IN FAR ZONE

For the return links from users not in the immediate vicinity of the station, the number of reflected paths received by the SS antenna is generally smaller than in the case of the EVA's. The severity of the multipath may also be reduced by the directivity of the receiving antenna as, for example, in the case of the 100 Kbps T&C OMV/OTV return links.

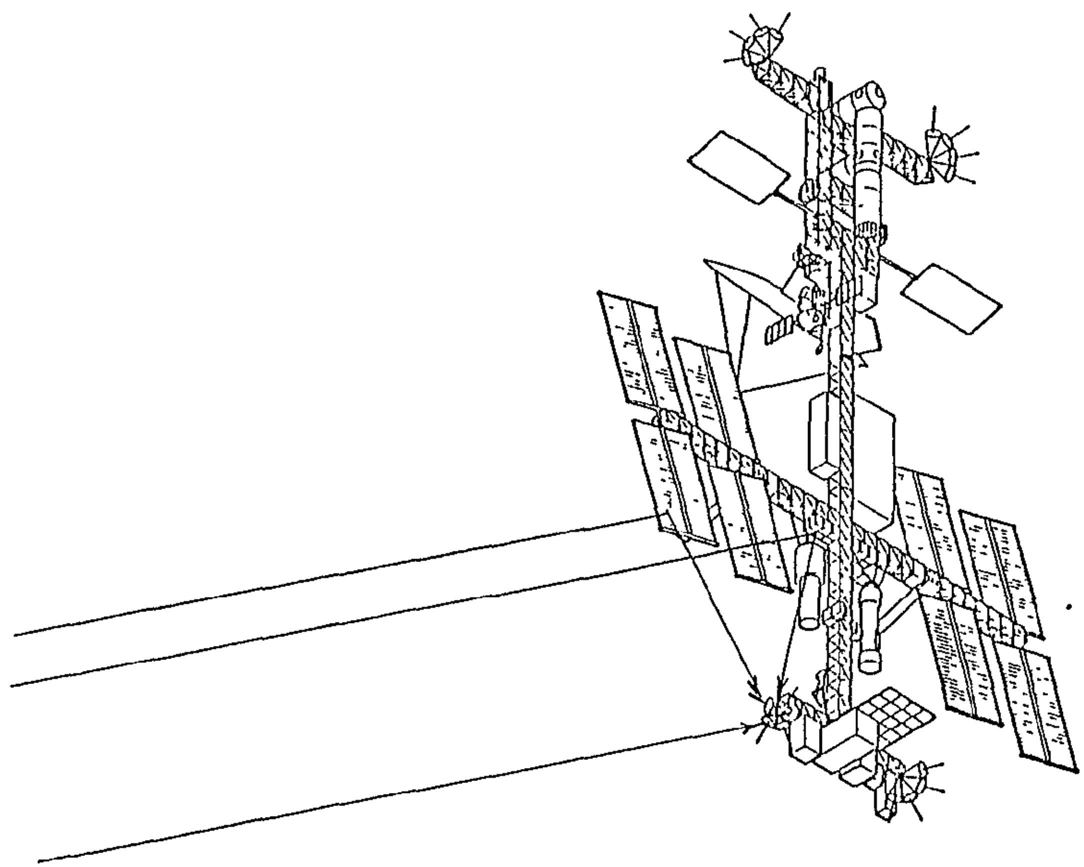


Fig. 1b. Possible Multipath From Transmitter in Far Zone.

The path differential delay is much smaller than the bit duration if its less than 10% of the bit time. In practice, actually 20% of the bit duration is accepted as the limit under which intersymbol distortion effects can be ignored and a slow fading model can be considered. For the 100 Kbps voice and telemetry links the bit duration is 10 μsec and a path difference of 600 meters causes a differential delay equal to 20% of the bit time. Since path differences of more than 600 meters are very unlikely to be created by the space station structure, moreover since the change in geometry is much slower than the data rate, a slow fading model is indeed appropriate for the 100 Kbps links.

LinCom

1. <u>LOWER RATE LINKS (R < 1 Mbps)</u>

   e.g. VOICE & TELEMETRY LINKS AT 100 Kbps

   - PATH DIFFERENTIAL DELAY << BIT DURATION (10 μsec NEEDS A PATH DIFFERENCE OF 3000 METERS)

   - SYSTEM CONFIGURATION GEOMETRY VARIES MUCH SLOWER THAN DATA RATE

- FROM ABOVE 2 REASONS, A "SLOW FADING", i.e., NO TIME OR FREQUENCY DISPERSION, MODEL IS APPROPRIATE

- RECEIVER CARRIER AMPLITUDE AND PHASE ARE APPROXIMATELY CONSTANT OVER A SINGLE BIT DURATION. THEY ARE, HOWEVER, RANDOM VARIABLES (DUE TO FADING) AND CAN CHANGE FOR DIFFERENT BITS IN THE DATA STREAM.

LinCom

The so-called Rayleigh criterion $h \gtrless \frac{\lambda}{8 \sin \gamma}$ is used in practice to determine the roughness (or smoothness) of the reflecting surfaces. It is obvious from the formula that any surface appears the roughest at normal incidence and smoother for incident rays which are almost parallel to the surface. (This is intuitively obvious in everyday life when the sun is reflecting off the pavement of the distant part of a highway; the same is true with sun reflection off the surface of the ocean.)

At 15 GHz λ = 2 cm and for γ as low as 10°, the surface would be rough if h > 14 mm (about 1/2 inch). If the incidence is normal (γ = 90°) the surface would appear rough if its granularity is any bigger than 2.5 mm (1/10 inch), a rather small granularity.

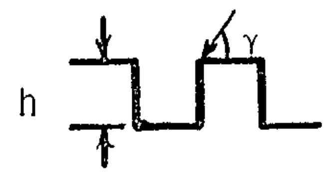
LinCom

POSSIBLE RECEIVED CARRIER COMPONENTS:

1. DIRECT PATH
2. ONE OR MORE SPECULAR REFLECTION COMPONENTS FROM SURFACES THAT APPEAR SMOOTH.
3. A SCATTER COMPONENT FROM SURFACES THAT APPEAR ROUGH.

FOLLOWING CRITERION MAY BE USED TO DETERMINE ROUGHNESS

$$h \gtrless \frac{\lambda}{8 \sin \gamma}$$



WHERE h IS GRANULARITY OF SURFACE. IF h IS $> \lambda/8 \sin \gamma$ THE SURFACE IS ROUGH.

RESULTING CARRIER STATISTICS:

- AMPLITUDE HAS A RICIAN-TYPE DISTRIBUTION
- PHASE IS UNIFORM

LinCom

RESULTS FOR SLOW FADING

It should be kept in mind that γ^2 is the ratio of the received direct power to the received reflected power. As such, the directivity of the receiving antenna plays an important role in determining the value of γ^2 for a particular scenario. In general, γ^2 is more likely to be large (little fading) for a highly directive receiving antenna.

LinCom

RESULTS FOR SLOW FADING

GENERAL COMMENTS

- γ^2 (RATIO OF RECEIVED DIRECT TO REFLECTED POWERS) IS KEY PARAMETER IN DETERMINING PERFORMANCE
- FOR HIGH γ^2 (AT LEAST 10 dB), PSK, DPSK OR FSK MAY BE USED.
- FOR LOWER γ^2, PSK IS IMPRACTICAL DUE TO TRACKING LOOP FAILURE.
- FOR SLOW FADING DPSK IS SIMILAR TO NONCOHERENT FSK, BUT WITH 3 dB SNR IMPROVEMENT (NEXT CHART)

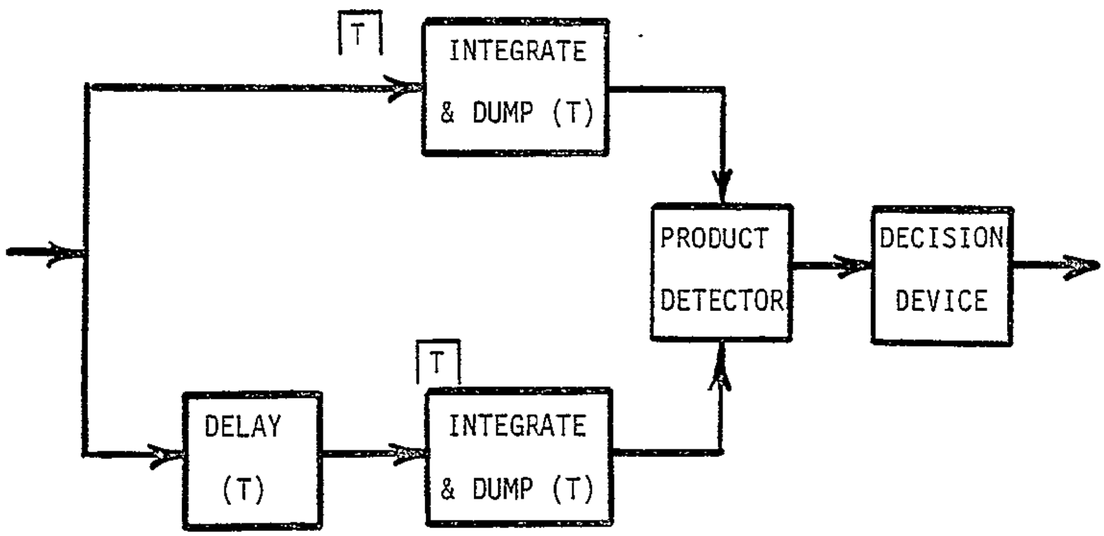
LinCom

C-3

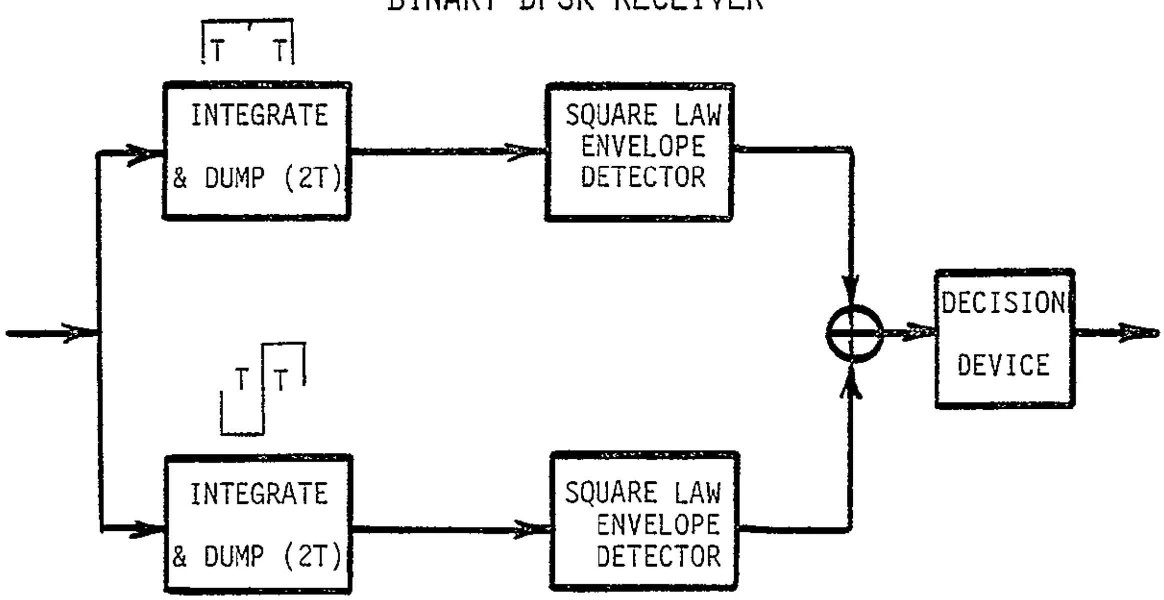
EQUIVALENCE OF DPSK AND FSK

This chart shows on top a coherent receiver of binary DPSK. The matched filters in the arms are matched to the symbol of duration T, as indicated on the block diagram. The product detector multiplies the outputs of the arm filters and extracts the baseband and rejects the double frequency terms. After some straightforward algebra it can be shown that the operation and performance of the coherent receiver is identical to the equivalent noncoherent receiver shown in the bottom of the chart. In the latter receiver the matched filters are matched to the shown symbols of length 2T; one symbol corresponding to two bits with no transition between them and the other two bits with a transition between them. After some additional algebra it can be shown that this noncoherent DBPSK receiver is equivalent to a noncoherent BFSK receiver, where the energy in the BFSK signals is equal to the energy in the DBPSK signals of duration <u>2T</u>. Hence, when DBPSK and noncoherent BFSK signals of duration T are compared, DBPSK is found to be 3 dB more efficient. The two schemes are otherwise similar in performance.

EQUIVALENCE OF DPSK AND FSK



BINARY DPSK RECEIVER



EQUIVALENT NONCOHERENT DETECTOR

BASELINE PERFORMANCE COMPARISON

AT $\gamma^2 = \infty$ (NO FADING) AND BER = 10^{-5}

- BPSK IS 0.7 dB SUPERIOR TO BINARY DPSK (DIFFERENTIAL ENCODING AND DIFFERENTIALLY COHERENT DETECTION).
- DBPSK, IN TURN, IS 3 dB BETTER THAN BFSK (NONCOHERENT DETECTION).
- FOR QUATERNARY SIGNALS, HOWEVER, DQPSK IS 2.5 dB INFERIOR TO QPSK.

BASELINE PERFORMANCE COMPARISON

In the following charts we will be comparing the increase in SNR needed to overcome the degradation in performance due to multipath for the different modulation schemes. The degradation in the performance of each scheme is relative to what may be called its "baseline" performance under no fading (i.e. $\gamma^2 = \infty$).

LinCom

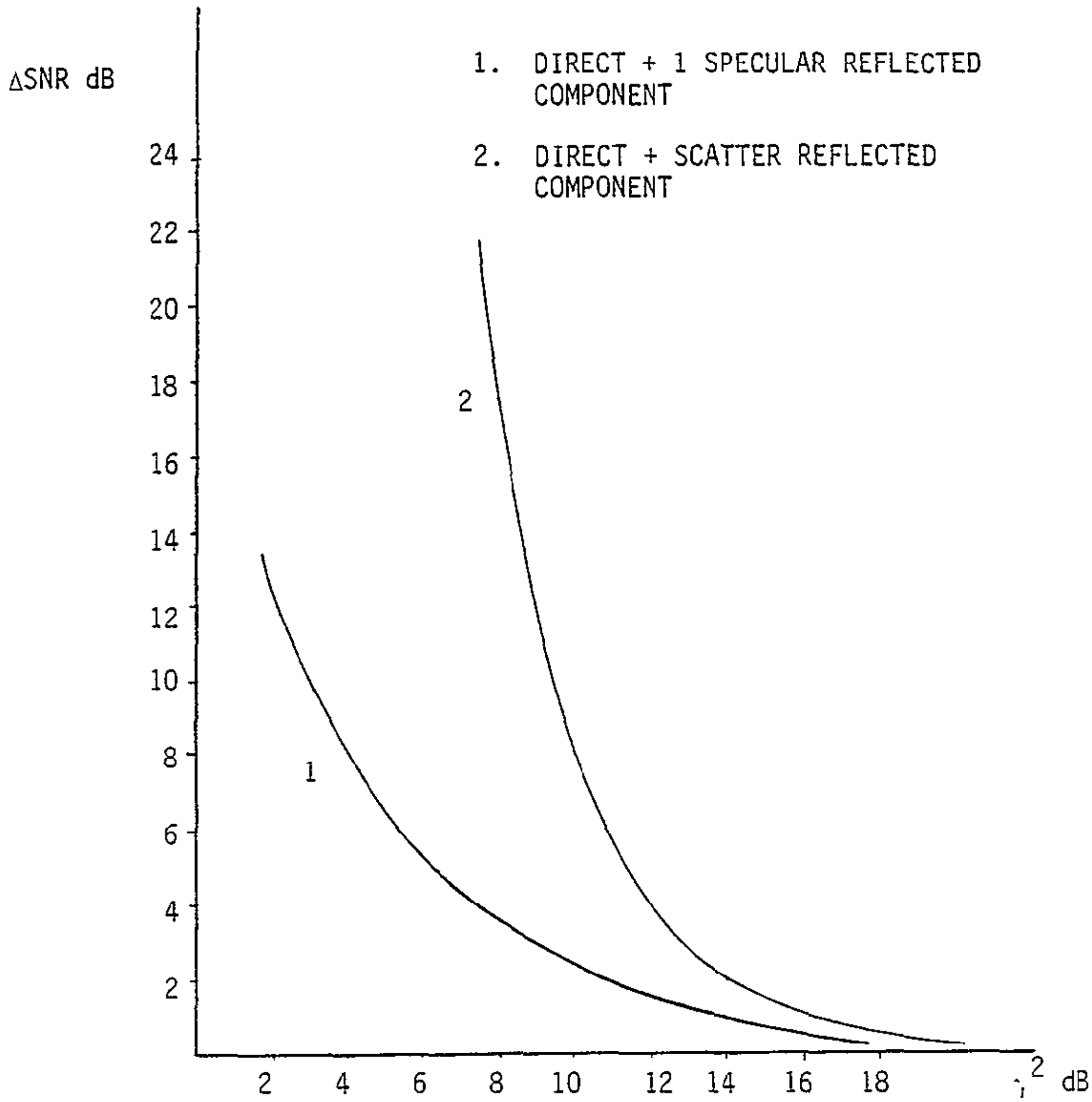
INCREASE IN DIRECT PATH SNR NEEDED TO MAINTAIN 10^{-5} BER PERFORMANCE FOR BFSK OR DPSK

In this and the following two charts it should be noted that for a certain fixed scenario, if a reflecting surface is considered rough (scattering) rather than smooth, the corresponding value of γ^2 should be different in each case. This is because the phenomenon of scattering implies a diffusion of the reflected power in different directions, as a result not all the reflected diffused power will be incident on the receiving antenna. This is in contrast to specular reflection where the entire reflected ray may be incident on the receiving antenna. The point to be made is that one should be cautious in interpreting the extent of performance degradation caused by the same value of γ^2 in the two types of reflection (since it may correspond to two different scenarios).

LinCom

LinCom

INCREASE IN DIRECT PATH SNR NEEDED TO MAINTAIN 10^{-5} BER PERFORMANCE FOR BFSK OR DPSK



$$\gamma^2 = \frac{\text{Power of Received Direct Component}}{\text{Power of Received Reflected Component}}$$

LinCom

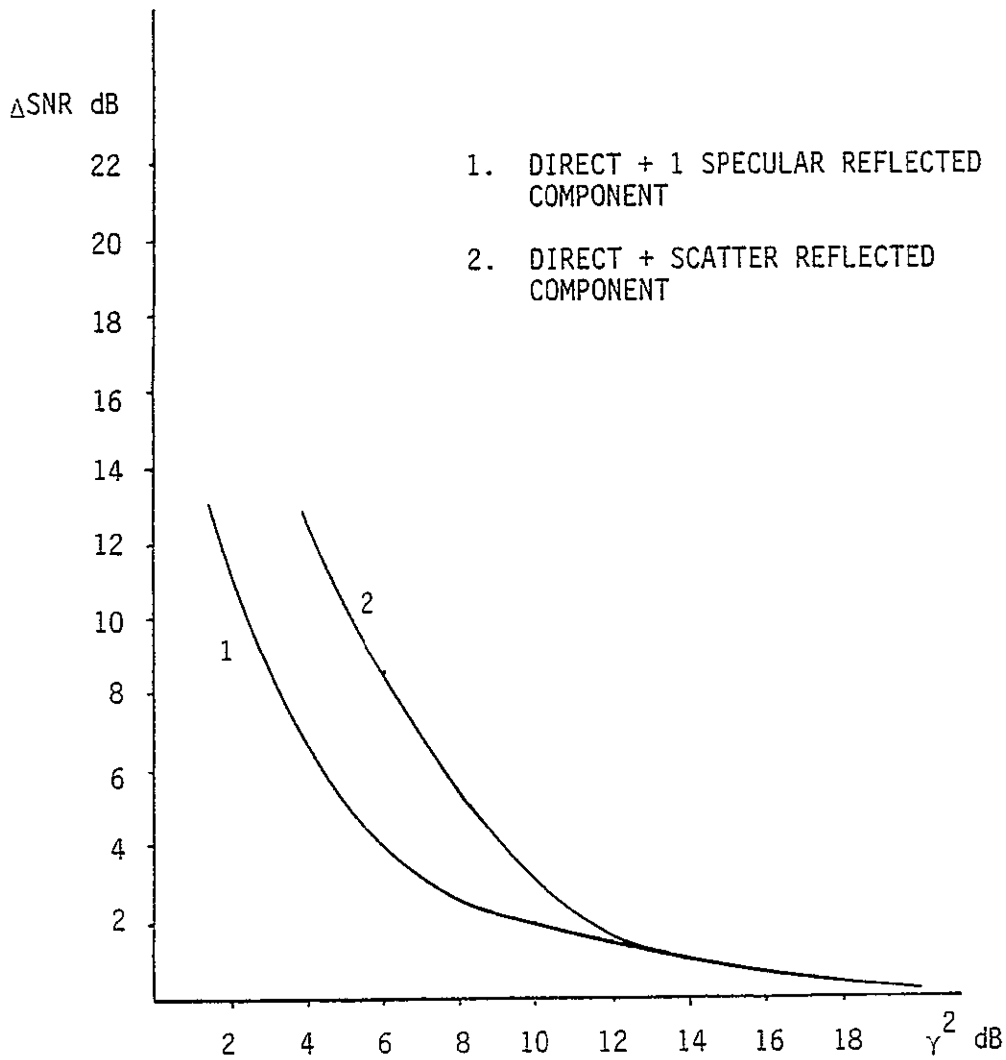
LinCom

INCREASE IN DIRECT PATH SNR NEEDED TO MAINTAIN 10^{-3} BER PERFORMANCE FOR BFSK OR DPSK

For a given value of γ^2, the increase in direct path SNR to preserve a 10^{-3} BER performance (this chart) is naturally less than the increase in SNR needed to preserved a 10^{-5} BER performance (last chart).

LinCom

INCREASE IN DIRECT PATH SNR NEEDED TO MAINTAIN 10^{-3} BER PERFORMANCE FOR BFSK OR DPSK



$$\gamma^2 = \frac{\text{Power of Received Direct Component}}{\text{Power of Received Reflected Component}}$$

LinCom

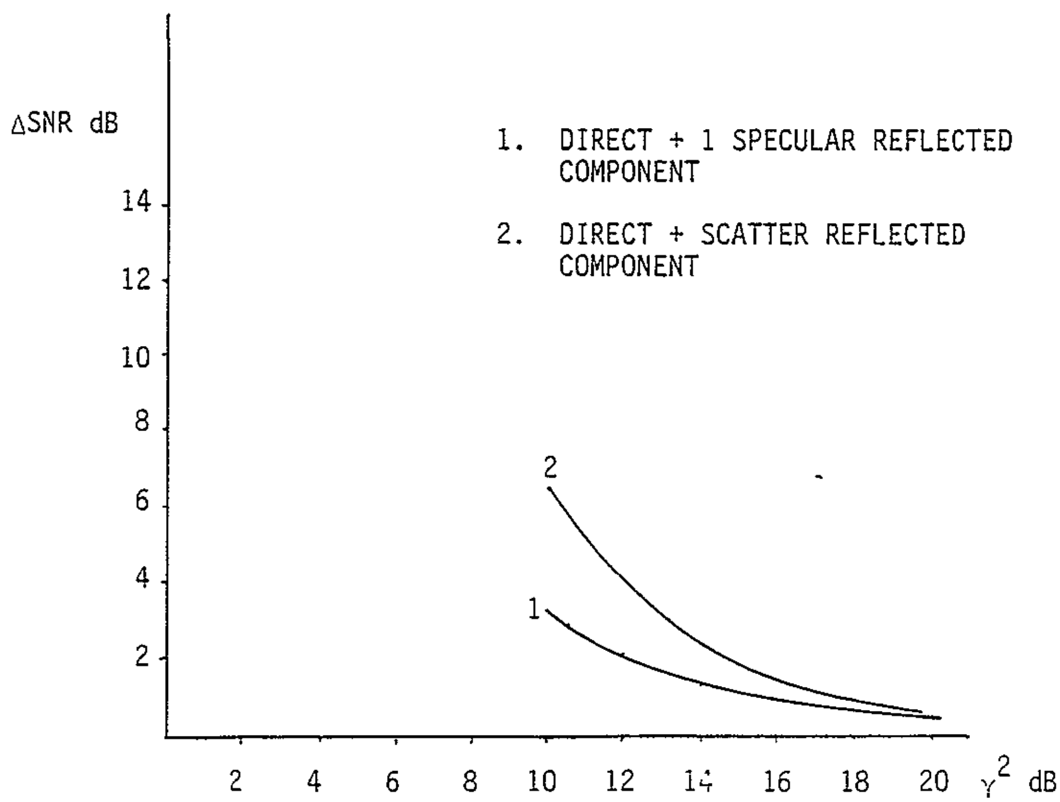
INCREASE IN DIRECT PATH SNR NEEDED TO MAINTAIN 10^{-5} BER PERFORMANCE FOR BPSK

The curves shown in this chart for coherent BPSK are not drawn for $\gamma^2 < 10$. This is because at $\gamma^2 < 10$ the degradation in performance due to carrier tracking loop anomalies (large tracking phase error, cycle slipping, loss of lock, etc.) can no longer be ignored. These deleterious effects make coherent BPSK unsuitable for multipath situations where $\gamma^2 < 10$.

LinCom

LinCom

INCREASE IN DIRECT PATH SNR NEEDED TO MAINTAIN 10^{-5} BER PERFORMANCE FOR BPSK



$$\gamma^2 = \frac{\text{Power of Received Direct Component}}{\text{Power of Received Reflected Component}}$$

LinCom

LinCom

COMMENTS ON RESULTS

The fact that the presence of a number of specularly reflected rays can be effectively treated as scattering (with an appropriate choice of γ^2) is established by invoking the central limit theorem. It has been observed[1] that the resulting approximation holds well even for a few specularly reflected rays.

[1] See for example: D. R. Morgan, "Error Rate of Phase-Shift Keying in the Presence of Discrete Multipath Interference," IEEE Trans. on Information Theory, July 1972, pp. 525-528.

LinCom

LinCom

COMMENTS ON RESULTS

- FOR ALL SCHEMES, LEAST PERFORMANCE DEGRADATION (FOR A GIVEN γ^2) IS DUE TO 1 SPECULARLY REFLECTED RAY (WITH RANDOM PHASE).
- FOR ALL SCHEMES, WORST PERFORMANCE DEGRADATION (FOR A GIVEN γ^2) IS DUE TO SCATTERING
- SEVERAL SPECULARLY REFLECTED PATHS QUICKLY APPROACH SCATTERING CASE.
- TYPICAL OPERATION LIES SOMEWHERE BETWEEN CURVES 1 AND 2.

LinCom

CONCLUSIONS (LOWER DATA RATES)

Since the penalty in going from coherent to differential PSK is only 0.7 dB, DPSK seems to be a more suitable choice in the multipath environment because its operation is more reliable in the event of deep fades ($\gamma^2 < 10$ dB).

Since DPSK and BFSK degrade by the same amount under slow fading, and since DPSK is 3 dB better than BFSK to start with, DPSK is a better choice for the low rate links (for which the fading is slow).

LinCom

CONCLUSIONS (LOWER DATA RATES)

- FOR LARGE γ^2 (> 10 dB) AND INFREQUENT SEVERE FADES (γ^2 < 10 dB), DEGRADATION FOR ALL SYSTEMS IS SIMILAR.

- DPSK IS MORE RELIABLE THAN PSK, 0.7 dB PENALTY IS INSIGNIFICANT.

- DPSK IS SUPERIOR TO FSK FOR LOW RATE LINKS.

LinCom

For multipath where the differential path delays are of the order of the bit duration, we cannot in general assume that the autocorrelation of the received signal over a single bit duration is approximately constant. As an immediate result slow fading (zero bandwidth fading) is not generally applicable.

As a first approximation, if the fading is not severe one can argue that the variation in the received signal autocorrelation over a bit duration is still not significant. One can then apply the slow fading model as a rough approximation.

The non-zero bandwidth fading caused by the long differential path delays generally causes intersymbol interference. This is equivalent to the presence of frequency selectivity on the multipath channel. The intersymbol interference may also be viewed as causing self-jamming if an FSK scheme is used (discussed next chart).

2. HIGHER DATA RATE LINKS (R > 1 Mbps)

e.g. TV CHANNELS AT 22 Mbps

- CONFIGURATION GEOMETRY STILL CHANGES MUCH SLOWER THAN DATA RATE
- DIFFERENTIAL PATH DELAYS ARE OF ORDER OF BIT DURATION (13.5 METERS IS PATH DIFFERENCE CORRESPONDING TO 1 BIT AT 22 Mbps)

TO A FIRST APPROXIMATION "SLOW FADING" MODEL CAN BE APPLIED.

A MORE ACCURATE TREATMENT NEEDS AN ASSESSMENT OF THE "TIME DISPERSION" EFFECTS CAUSED BY THE DIFFERENTIAL PATH DELAYS. THIS MAY BE MANIFESTED AS "FREQUENCY-SELECTIVE" FADING, SELF-JAMMING OR SYMBOL SMEARING.

CONCERNS

The significant changes in the phase of the received signal during a bit duration invalidate the assumption that was used in demonstrating the equivalence of DPSK and FSK under slow fading conditions. (The assumption is that the change in the received signal phase due to multipath is negligible over any two adjacent bit durations). Since these phase changes are equivalent to using a noisy reference, DPSK exhibits sensitivity and faster degradation with this type of multipath than does FSK which relies on energy detection.

The performance of FSK under long differential delays is inferior to its performance under insignificant differential delays. This is due to the interfering energy that corresponds to previous information bits and which may fall in alternate filters (depending on successive bit polarity). The fall of such interference energy in alternate filters becomes more likely with the increase in M the number of tones in MFSK.

LinCom

CONCERNS

- DPSK NO LONGER SIMILAR (BUT SUPERIOR) TO FSK
- DPSK MAY DEGRADE FASTER THAN FSK; FSK LESS SENSITIVE TO FREQUENCY SELECTIVE MULTIPATH (SEE NEXT CHART)
- PERFORMANCE OF FSK CAN BE SIGNIFICANTLY INFERIOR THAN ITS PERFORMANCE IN SLOW FADING
  - LONG DIFFERENTIAL DELAYS CAUSE SELF-JAMMING (RECEPTION OF POWER IN ALTERNATE FILTER)
  - MORE SEVERE SELF-JAMMING WITH M-ARY FSK

LinCom

FSK/DPSK BREAK-EVEN BER AS A FUNCTION OF FADING BANDWIDTH

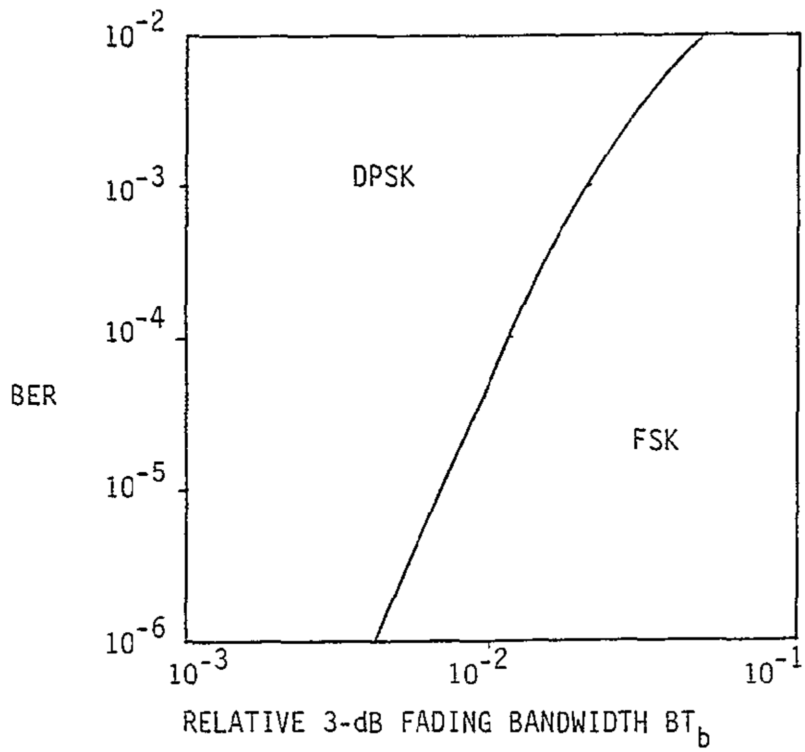
Since DPSK is 3 dB superior to BFSK when no multipath is present, and since it degrades in certain cases faster than FSK, a point comes where break-even occurs (for a given γ^2).

For high relative fading bandwidth, which signifies fast phase changes over a single bit duration, DPSK suffers considerably and FSK has the better performance. This is the region to the right of the break-even curve on the chart. Otherwise, i.e., if the fading bandwidth is small (closer to zero), DPSK remains superior to BSFK.

LinCom

FSK/DPSK BREAK-EVEN BER AS A FUNCTION OF FADING BANDWIDTH

- GAUSSIAN FADING AUTOCORRELATION (GIVES GAUSSIAN FADING SPECTRUM)
- ZERO FADING BANDWIDTH (CONSTANT AUTO-CORRELATION OVER T_b) IS SLOW (FLAT) FADING CASE



LinCom

RECOMMENDATIONS FOR FUTURE ANALYSIS OF MULTIPATH

An estimate of the extent of the multipath problem for the space station can be obtained from a knowledge of its geometry and system specs (data rates, antenna patterns, surface areas and dielectric constants, etc.).

Upper and lower bounds for the degradation in communication links performance can be obtained by using the specular reflection and scattering models highlighted in the previous charts. The determination of the key parameter γ^2 for each scenario is the task that requires the most significant effort.

LinCom

RECOMMENDATIONS FOR FUTURE ANALYSIS OF MULTIPATH

- SPACE STATION MULTIPATH PROBLEM IS DEPENDENT ON GEOMETRY
- MULTIPATH MODEL PARAMETERS ARE OBTAINABLE FROM GEOMETRY AND SYSTEM SPECS
- SPECIFICATION OF MODEL PARAMETERS REQUIRES:
  - DETERMINATION OF ANTENNA POSITIONS AND PATTERNS FOR SPACE STATION
  - KNOWLEDGE OF DIELECTRIC CONSTANTS, AREAS, LIKELY RANGES OF INCLINATION ANGLES OF SOLAR ARRAYS
  - SPECIFICATION OF SURFACE CHARACTERISTICS FOR OTHER LARGE SURFACE-AREA MODULES

LinCom

MA LINK BUDGETS

The next set of charts shows the link budgets for the MA links.

MA LINK BUDGETS

LinCom

GROUND RULES

The RF power at 1 watt and recevier noise figure at 3 db are slightly pessimistic estimates. When a 5 watt RF power is required, it can be provided by either combining multiple power FETs or by using a TWTA with sufficient backoff, e.g., using a 20 W TWT. This assures linear operation of the power amplification stage. The 3 db pointing loss used assumes that the crossover boundary for the MA antenna system is 3 db. It also assumes that the user antenna has a worst case 3 db loss. This assumption is realistic for the EVA antenna.

LinCom

LinCom

GROUND RULES

- 1 WATT GaAs POWER FET ARE USED WHERE POSSIBLE
- 3 dB POINTING LOSS ARE ASSUMED FOR TRANSMIT AND RECEIVE ANTENNAS
- RECEIVER NOISE FIGURE IS 3 dB
- RF LOSSES ARE 1 dB (TRANSMIT AND RECEIVE)
- 3 dB FADING LOSS IS ASSUMED WHENEVER A 0 dB GAIN ANTENNA IS USED
- POWER CONTROL IS NOT USED AS RANGE VARIES
- 5 dB SYSTEM MARGIN TARGET
- CODING FOR FF T&C LINKS ONLY
- 2 dB IMPLEMENTATION LOSS

LinCom

LinCom

MA FORWARD LINK ANTENNA

This chart shows the allocation of SS antennas for the MA user links. With the dedicated fixed-beam (FB) antenna system for the free flyers, the omni-antenna order-wire (OW) link is only used for the OMV/OTV users. The function of the OW link for the EVAs is combined with the T&C link since it already uses the omni antenna. The OW link is active continuously between the SS and the user. The main purpose of the OW is to allocate the MA high gain antenna beams to the users as needed, since the 23 db antenna system can support only 5-6 users at a time.

LinCom

LinCom

MA FORWARD LINK ANTENNA

ANTENNA	DATA CHANNEL	USERS			
		EVA	OMV/OTV	FF	TOTAL
OMNI (0 dB)	100 Kbps T&C	4(6)	--	--	4(6)
	275 Kbps HUD	4(6)	--	--	4(6)
	8/22 Mbps TV	2	--	--	2
	100 bps OW	--	1(3)	--	1(3)
MA (23 dB)	100 Kbps T&C	--	1(3)	--	1(3)
	8/22 Mbps TV	2	--	--	2
FB (23 dB)	100 Kbps T&C	--	--	5(9)	5(9)
FB (28 dB)	100 Kbps T&C	--	--	5(9)	5(9)

LinCom

LinCom

MA RETURN LINK ANTENNA

The return link antenna allocation is basically a reciprocal of the forward link.

LinCom

LinCom

MA RETURN LINK ANTENNA

ANTENNA	DATA CHANNEL	USERS			
		EVA	OMV/TV	FF	TOTAL
OMNI (0 dB)	100 Kbps T&C	4(6)	--	--	4(6)
	8/22 Mbps TV	4(6)	--	--	4(6)
	100 bps OW	--	1(3)	--	1(3)
MA (23 dB)	100 Kbps T&C	--	1(3)	--	1(3)
	8/22 Mbps TV	4(6)	1(2)	--	5(8)
FB (23 dB)	100 Kbps T&C	--	--	5(9)	5(9)
FB (28 dB)	100 Kbps T&C	--	--	5(9)	5(9)

LinCom

MA FORWARD LINK BUDGET

For the EVA links, it is assumed that the the T&C and HUD are combined into a single link so that the total transmit power in the chart is 1 watt. The antenna gain is switched from 0 db to 23 db at 200 meters for the TV links. It is assumed that power control is not used, so that the far zone coverage from .2 to 1 Km has considerably higher margins.

The OMV/OTV link shows a 3.1 db margin. However, only a 0 db antenna is assumed for the user. For the range from 37 to 185 Km, it is very likely that the OMV will use a higher gain antenna, since antenna pointing is not a problem for the OMV at these distances. (Only a knowledge of an approximate location of the SS to a few Km is needed to point a medium gain antenna.) Hence this link should have a healthy margin.

A 17 db gain antenna is assumed for the free flyers. Again, at a minimum distance of 185 Km away and with that kind of gain (beamwidth), antenna pointing is not a real problem.

LinCom

MA FORWARD LINK BUDGET

	EVA						OMV/OTV		FREE FLYER	
			TV 8		TV 22				T&C	
LINK PARAMETERS	T&C	HUD	NEAR	FAR	NEAR	FAR	T&C	DW	NEAR	FAR
TRANSMIT POWER, W	0.5	0.5	1.0	1.0	1.0	1.0	5.0	1.0	5.0	5.0
(DBW)	-3.0	-3.0	0.0	0.0	0.0	0.0	7.0	0.0	7.0	7.0
TRANSMIT LOSS, DB	-1.0	-1.0	-1.0	-1.0	-1.0	-1.0	-1.0	-1.0	-1.0	-1.0
TRANSMIT ANTENNA GAIN, DB	0.0	0.0	0.0	23.0	0.0	23.0	23.0	0.0	23.0	28.0
POINTING LOSS, DB	-3.0	-3.0	-3.0	-3.0	-3.0	-3.0	-3.0	-3.0	-3.0	-3.0
EIRP, DBW	-7.0	-7.0	-4.0	19.0	-4.0	19.0	26.0	-4.0	26.0	31.0
RANGE, KM	1.0	1.0	0.2	1.0	0.2	1.0	185.0	185.0	1000.0	2000.0
FREQUENCY, GHZ	15.0	15.0	15.0	15.0	15.0	15.0	15.0	15.0	15.0	15.0
SPACE LOSS, DB	-116.0	-116.0	-102.0	-116.0	-102.0	-116.0	-161.3	-161.3	-176.0	-182.0
RECEIVE ANTENNA GAIN, DB	0.0	0.0	0.0	0.0	0.0	0.0	0.0	0.0	17.0	17.0
POLARIZATION LOSS, DB	0.0	0.0	0.0	0.0	0.0	0.0	0.0	0.0	0.0	0.0
POINTING LOSS, DB	-3.0	-3.0	-3.0	-3.0	-3.0	-3.0	-3.0	-3.0	-3.0	-3.0
RF LOSS, DB	-1.0	-1.0	-1.0	-1.0	-1.0	-1.0	-1.0	-1.0	-1.0	-1.0
RECEIVE POWER, DBW	-127.0	-127.0	-110.0	-101.0	-110.0	-101.0	-139.3	-169.3	-137.0	-138.0
RECEIVER NOISE FIGURE, DB	3.0	3.0	3.0	3.0	3.0	3.0	3.0	3.0	3.0	3.0
RECEIVE SYSTEM TEMP, DBK	27.6	27.6	27.6	27.6	27.6	27.6	27.6	27.6	27.6	27.6
NOISE DENSITY, DBW/HZ	-201.0	-201.0	-201.0	-201.0	-201.0	-201.0	-201.0	-201.0	-201.0	-201.0
RECEIVED C/NO, DB-HZ	74.0	74.0	91.0	100.0	91.0	100.0	61.7	31.7	64.0	63.0
DATA RATE, KBPS	100.0	275.0	8000.0	8000.0	22000.0	22000.0	100.0	0.1	100.0	100.0
(DB-HZ)	50.0	54.4	69.0	69.0	73.4	73.4	50.0	20.0	50.0	50.0
AVAILABLE EB/NO, DB	24.0	19.6	22.0	31.0	17.6	26.6	11.7	11.7	14.0	13.0
REQUIRED EB/NO, DB	9.6	9.6	9.6	9.6	9.6	9.6	9.6	9.6	9.6	9.6
CODING GAIN, DB	0.0	0.0	0.0	0.0	0.0	0.0	5.0	5.0	5.0	5.0
IMPLEMENTATION LOSS, DB	-2.0	-2.0	-2.0	-2.0	-2.0	-2.0	-2.0	-2.0	-2.0	-2.0
FADING LOSS,DB	-2.00	-2.00	-2.00	-2.00	-2.00	-2.00	-2.00	-2.00	0.00	0.00
SYSTEM MARGIN, DB	10.4	6.0	8.4	17.4	4.0	13.0	3.1	3.1	7.4	6.4

LinCom

MA RETURN LINK BUDGET

The return link budget is basically the reciprocal of the forward link. The main difference is that the OMV/OTV return links include TV. For the OMV OW and T&C links, a 185 Km maximum range is used as in the forward link. For the TV links, a maximum range of 37 Km (end of the disc region) is used. From 37 to 185 Km, an additional margin of 14 db must be provided. If required, this can be realized by a comination of the following: employment of coding, increased transmit power, increased antenna gain, and decrease in data rate.

LinCom

MA RETURN LINK BUDGET

	EVA					OMV/OTV				FREE FLYER	
		TV 8		TV 22						T&C	
LINK PARAMETERS	T&C	NEAR	FAR	NEAR	FAR	DW	T&C	TV 8	TV 22	NEAR	FAR
TRANSMIT POWER, W	0.2	1.0	1.0	1.0	1.0	1.0	5.0	2.0	5.0	5.0	5.0
(DBW)	-7.0	0.0	0.0	0.0	0.0	0.0	7.0	3.0	7.0	7.0	7.0
TRANSMIT LOSS, DB	-1.0	-1.0	-1.0	-1.0	-1.0	-1.0	-1.0	-1.0	-1.0	-1.0	-1.0
TRANSMIT ANTENNA GAIN, DB	0.0	0.0	0.0	0.0	0.0	0.0	0.0	17.0	17.0	17.0	17.0
POINTING LOSS, DB	-3.0	-3.0	-3.0	-3.0	-3.0	-3.0	-3.0	-3.0	-3.0	-3.0	-3.0
EIRP, DBW	-11.0	-4.0	-4.0	-4.0	-4.0	-4.0	3.0	16.0	20.0	20.0	20.0
RANGE, KM	1.0	0.2	1.0	0.2	1.0	185.0	185.0	37.0	37.0	1000.0	2000.0
FREQUENCY, GHZ	15.0	15.0	15.0	15.0	15.0	15.0	15.0	15.0	15.0	15.0	15.0
SPACE LOSS, DB	-116.0	-102.0	-116.0	-102.0	-116.0	-161.3	-161.3	-147.3	-147.3	-176.0	-182.0
RECEIVE ANTENNA GAIN, DB	0.0	0.0	23.0	0.0	23.0	0.0	23.0	23.0	23.0	23.0	28.0
POLARIZATION LOSS, DB	0.0	0.0	0.0	0.0	0.0	0.0	0.0	0.0	0.0	0.0	0.0
POINTING LOSS, DB	-3.0	-3.0	-3.0	-3.0	-3.0	-3.0	-3.0	-3.0	-3.0	-3.0	-3.0
RF LOSS, DB	-1.0	-1.0	-1.0	-1.0	-1.0	-1.0	-1.0	-1.0	-1.0	-1.0	-1.0
RECEIVE POWER, DBW	-131.0	-110.0	-101.0	-110.0	-101.0	-169.3	-139.3	-112.3	-108.3	-137.0	-138.0
RECEIVER NOISE FIGURE, DB	3.0	3.0	3.0	3.0	3.0	3.0	3.0	3.0	3.0	3.0	3.0
RECEIVE SYSTEM TEMP, DBK	27.6	27.6	27.6	27.6	27.6	27.6	27.6	27.6	27.6	27.6	27.6
NOISE DENSITY, DBW/HZ	-201.0	-201.0	-201.0	-201.0	-201.0	-201.0	-201.0	-201.0	-201.0	-201.0	-201.0
RECEIVED C/NO, DB-HZ	70.0	91.0	100.0	91.0	100.0	31.7	61.7	88.7	92.6	64.0	63.0
DATA RATE, KBPS	100.0	8000.0	8000.0	22000.0	22000.0	0.1	100.0	8000.0	22000.0	100.0	100.0
(DB-HZ)	50.0	69.0	69.0	73.4	73.4	20.0	50.0	69.0	73.4	50.0	50.0
AVAILABLE EB/NO, DB	20.0	22.0	31.0	17.6	26.6	11.7	11.7	19.6	19.2	14.0	13.0
REQUIRED EB/NO, DB	9.6	9.6	9.6	9.6	9.6	9.6	9.6	9.6	9.6	9.6	9.6
CODING GAIN, DB	0.0	0.0	0.0	0.0	0.0	5.0	5.0	0.0	0.0	5.0	5.0
IMPLEMENTATION LOSS, DB	-2.0	-2.0	-2.0	-2.0	-2.0	-2.0	-2.0	-2.0	-2.0	-2.0	-2.0
FADING LOSS,DB	-2.00	-2 00	-2.00	-2.00	-2.00	-2.00	-2.00	-2.00	-2.00	0.00	0.00
SYSTEM MARGIN, DB	6.4	8.4	17.4	4.0	13.0	3.1	3.1	6.0	5.6	7.4	6.4

LinCom

FREE FLYER LINK BUDGET USING MA ANTENNA

If the fixed beam antenna is not used, then an OW link must be included for the FF links. Because of the differences in antenna gains, the T&C link transmit power must be increased from the previous charts.

LinCom

FREE FLYER LINK BUDGET USING MA ANTENNA

	FREE FLYER			
	FORWARD		RETURN	
LINK PARAMETERS	T&C	DW	T&C	DW
TRANSMIT POWER, W	12.0	4.0	12.0	4.0
(DBW)	10.8	6.0	10.8	6.0
TRANSMIT LOSS, DB	-1.0	-1.0	-1.0	-1.0
TRANSMIT ANTENNA GAIN, DB	23.0	0.0	17.0	17.0
POINTING LOSS, DB	-3.0	-3.0	-3.0	-3.0
EIRP, DBW	29.8	2.0	23.8	19.0
RANGE, KM	2000.0	2000.0	2000.0	2000.0
FREQUENCY, GHZ	15.0	15.0	15.0	15.0
SPACE LOSS, DB	-182.0	-182.0	-182.0	-182.0
RECEIVE ANTENNA GAIN, DB	17.0	17.0	23.0	0.0
POLARIZATION LOSS, DB	0.0	0.0	0.0	0.0
POINTING LOSS, DB	-3.0	-3.0	-3.0	-3.0
RF LOSS, DB	-1.0	-1.0	-1.0	-1.0
RECEIVE POWER, DBW	-139.2	-167.0	-139.2	-167.0
RECEIVER NOISE FIGURE, DB	3.0	3.0	3.0	3.0
RECEIVE SYSTEM TEMP, DBK	27.6	27.6	27.6	27.6
NOISE DENSITY, DBW/HZ	-201.0	-201.0	-201.0	-201.0
RECEIVED C/NO, DB-HZ	61.8	34.0	61.8	34.0
DATA RATE, KBPS	100.0	0.1	100.0	0.1
(DB-HZ)	50.0	20.0	50.0	20.0
AVAILABLE EB/NO, DB	11.8	14.0	11.8	14.0
REQUIRED EB/NO, DB	9.6	9.6	9.6	9.6
CODING GAIN, DB	5.0	5.0	5.0	5.0
IMPLEMENTATION LOSS, DB	-2.0	-2.0	-2.0	-2.0
FADING LOSS, DB	0.00	-2.00	0.00	-2.00
SYSTEM MARGIN, DB	5.2	5.4	5.2	5.4

LinCom

LinCom

MA SCHEMES

The next set of charts compares 4 candidate multiple access schemes. Explainations and additional details are dcoumented in a seperate report. For the SS MA application, FDMA appears to be the best technique. However one drawback of FDMA systems is that they are not inherently compatible with ranging. To circumvent this, a simple coarse ranging system is proposed.

LinCom

LinCom

MA SCHEMES

LinCom

LinCom

(THIS PAGE IS INTENTIONALLY BLANK)

LinCom

LinCom

COMPARISON SUMMARY FOR MULTIPLE ACCESS SCHEMES

	FDMA	TDMA	DSMA	FHMA
COMPLEXITY	LEAST	HIGH	MODERATE	HIGH
POWER REQUIREMENTS	MODERATE	HIGH	LOWEST	LOW
BW EFFICIENCY	MEDIUM	HIGHEST	LOW	LOW
INTERFERENCE REJECTION	NO	NO	YES	YES
RANGING CAPABILITY	NO	YES	EASILY ADDED	YES (COARSE)
GROWTH POTENTIAL	LIMITED (BW)	LIMITED (POWER)	GOOD	LIMITED (HOP RATE)
TECHNOLOGY LIMITATIONS	NONE	ANTENNA	HARDWARE LOSSES	NONE
COST	LOWEST	HIGH	MODERATE	HIGH

LinCom

LinCom

(THIS PAGE IS INTENTIONALLY BLANK)

LinCom

LinCom

COMPARATIVE FEATURES OF MULTIPLE ACCESS SCHEMES

(1) FDMA

ADVANTAGES

- SIMPLEST TECHNOLOGY; SIMPLEST HARDWARE ON SPACE STATION AND USERS; LOWEST COST

DRAWBACKS

- NEEDS USER POWER CONTROL TO RESOLVE NEAR FAR PROBLEM
- CROSSTALK, INTERMODULATION DISTORTION, WEAK SIGNAL SUPPRESSION MAY OCCUR
- LIMITED GROWTH POTENTIAL DUE TO INEFFICIENT USE OF BANDWIDTH
- NO INTERFERENCE REJECTION
- NO RANGING CAPABILITY

LinCom

PRECEDING PAGE BLANK NOT FILMED

(THIS PAGE IS INTENTIONALLY BLANK)

LinCom

(2) TDMA

ADVANTAGES

- HIGH BANDWIDTH EFFICIENCY
- INHERENT RANGING CAPABILITY
- NO NEED FOR USER POWER CONTROL

DRAWBACKS

- HIGHEST GAIN REQUIREMENTS; ANTENNA TECHNOLOGY A LIMITING FACTOR
- NEEDS HIGH SPEED AND NETWORK TIMING HARDWARE; MAY BE TOO MASSIVE TO DEPLOY IN SPACE
- NO INTERFERENCE REJECTION
- GROWTH SHOULD BE PRE-PLANNED

PRECEDING PAGE BLANK NOT FILMED

LinCom

LinCom

(THIS PAGE IS INTENTIONALLY BLANK

LinCom

LinCom

(3) DSMA

ADVANTAGES

- NETWORK TIMING CONTROL NOT NEEDED
- GOOD REJECTION OF RFI
- LOWEST POWER REQUIREMENTS
- GOOD GROWTH POTENTIAL

DRAWBACKS

- DESPREADING HARDWARE NEEDED ON SPACE STATION AND USERS; MODERATELY EXPENSIVE
- CHIP RATE MAY REQUIRE SPECIAL HARDWARE DESIGN
- USER POWER CONTROL ABSOLUTELY NECESSARY
- NO INHERENT RANGING CAPABILITY BUT CAN BE ADDED WITH MINOR IMPACT
- INTENTIONAL JAMMER RESISTENCE LIMITED BY BANDWIDTH

LinCom

(THIS PAGE IS INTENTIONALLY BLANK)

LinCom

(4) FHMA

ADVANTAGES

- INHERENT REJECTION OF RFI
- NETWORK COORDINATION GENERALLY NOT NEEDED
- INHERENT COARSE RANGING CAPABILITY
- POWER REQUIREMENTS LESS THAN FDMA
- CAN ACCOMODATE MANY USERS IF COST OF HIGH HOPPING RATE IS PERMISSABLE

DRAWBACKS

- COMPLEX AND EXTENSIVE HARDWARE ON SPACE STATION AND USERS
- FREQUENCY SYNTHESIZERS NEEDED TO PROVIDE MINIMUM HOPPING RATES (TO ACCOMODATE 16 USERS) ARE EXPENSIVE
- MEANINGFUL JAMMING RESISTENCE LIMITED BY BANDWIDTH AND COST OF HIGHER HOPPING RATES

LinCom

<u>A SIMPLE COARSE RANGING SYSTEM</u>

The proposed system is applicable to FDMA systems. In the SS application, in particular for the free flyers operating from 185 to 2000 Km, coarse range information with accuracy on the order of 1 Km is often adequate.

A SIMPLE COARSE RANGING SYSTEM

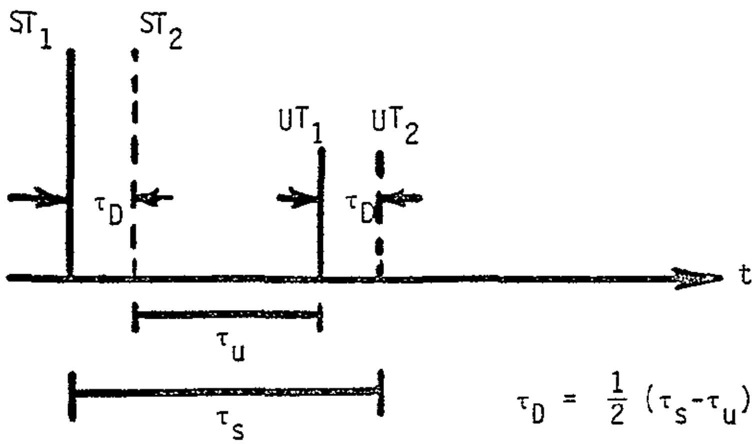
- CDMA CAN PROVIDE RANGING ACCURACY PROPORTIONAL TO CHIP DURATION; HOWEVER EQUIPMENT BIAS LIMITS REALIZABLE ACCURACY TO TENS OF METERS

- A SIMPLE RANGING SYSTEM CAN BE IMPLEMENTED USING FRAME SYNC INFORMATION AND YIELDS RAW ACCURACY (ASSUME 48 Kbps) OF ABOUT 500 m OR SMOOTHED ACCURACY OF ABOUT 200 m. THIS ACCURACY CAN BE TOLERATED FOR APPLICATIONS SUCH AS POWER CONTROL.

COARSE RANGING SYSTEM PRINCIPLE OF OPERATION

The propagation delay can be computed from half of the difference of two counter reading clocked by the transmit and receive frame sync epochs. The first counter reading is obtained on board the SS. The second counter reading must be transmitted back from the user to the SS via data links. The range can then be determined from the propagation delay. Because a difference technique is used, biases common to both measurements will cancel out.

COARSE RANGING SYSTEM
PRINCIPLE OF OPERATION



$$\tau_D = \frac{1}{2}(\tau_s - \tau_u)$$

ST_1 = SS Frame Sync Epoch transmitted at SS

ST_2 = SS Frame Sync Epoch received at User

UT_1 = User Frame Sync Epoch transmitted at User

UT_2 = User Frame Sync Epoch received at SS

τ_D = Propagation delay due to SS to User range

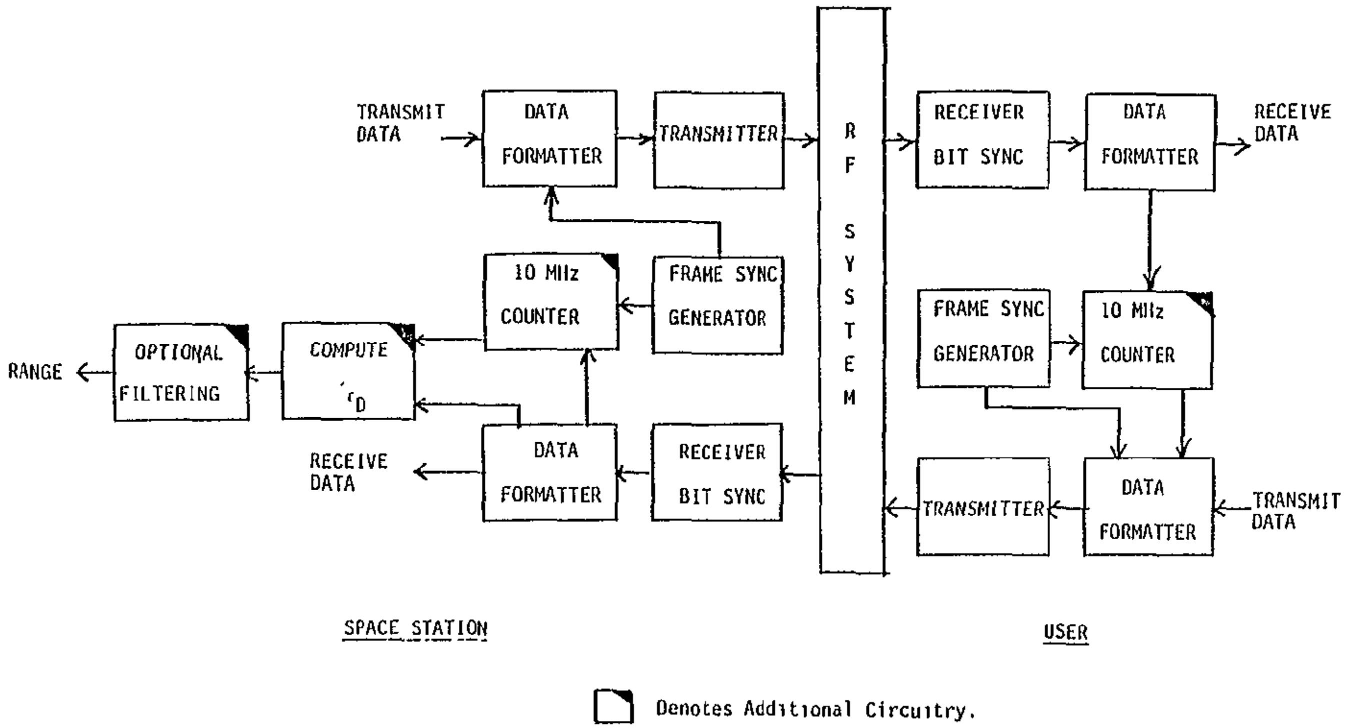
τ_s = SS measured time difference of transmitted SS Frame Sync Epoch and received User Frame Synch Epoch

τ_u = User measured time difference of received SS Frame Sync Epoch and transmitted User Frame Synch Epoch

COARSE ANGING SYSTEM HARDWARE IMPLEMENTATION

Notice that the amount of additional hardware is minimal. As long as the transmit and received frame rates are the same, the ranging system can be added easily at any time. The 10 Mhz counter provides a resolution of 100 ns, or 30 meters.

COARSE RANGING SYSTEM
HARDWARE IMPLEMENTATION



LinCom

SIGNAL DESIGN AND FREQUENCY PLAN

The next set of charts treats the proposed MA techniques and provides a tentative frequency plan and channel allocation for the MA user links. For the OW, CDMA is used. The rest of the links are FDMA. Here, the issues of adjacent channel seperation, and forward and return link adjacent channel interferences are addressed. Issues associated with further growth are also considered.

LinCom

LinCom

SIGNAL DESIGN AND

FREQUENCY PLAN

LinCom

LinCom

MA CHANNELS

The OW is a CDMA system so it can potentially accomodate up to hundreds of users. The overall strategy used here is to allocate as many channels as required to cover the projected need of the growth configuration. In the cases involving low date rate links, there are more channels allocated than required by growth since they do not really impact the overall bandwidth.

LinCom

LinCom

MA CHANNELS

	LINK	DATA RATE	# CHANNEL ASSIGNED (REQUIRED)	MA TECHNIQUE
FORWARD	T&C	100 Kbps	30(18)	FDMA
	HUD	275 Kbps	6(6)	FDMA
	TV	8/22 Mbps	2(2)	FDMA
	OW	100 bps	1* (3 Mcps)	CDMA
RETURN	T&C	100 Kbps	36(18)	FDMA
	TV	8/22 Mbps	5(5)	FDMA
	OW	100 bps	1* (3 Mcps)	CDMA

*CDMA

LinCom

CONSIDERATIONS FOR ORDER-WIRE CDMA CHANNEL

The CDMA OW channel allows very good ranging accuracy if coherent turnaround ranging is implemented. (The TDRSS system advertises a ranging accuracy of 20 ns, or 6 meters.) This is very useful for the OMV/OTV proximity operation as well as for traffic control. It is also anticipated that the OMV/OTV will have TDRSS capability so that the cost of the OW system can be shared if they are designed to be interoperable.

CONSIDERATIONS FOR ORDER-WIRE CDMA CHANNEL

- TDRSS CDMA FORMAT IS USED TO ASSURE COMPATIBILITY, AND TO TAKE ADVANTAGE OF SSA USER TRANSPONDER AND KSA MONOPULSE RECEIVER DEVELOPMENT (MOTOROLA)
- PROCESSING GAIN OF 30,000 (CHIP RATE = 3 Mcps, DATA RATE = 100 bps) CAN ACCOMODATE 1,000 USER SIMULTANEOUSLY
- ORDER-WIRE USED FOR OMV/OTV LINK (1 to 185 Km) ONLY SO THAT NEAR-FAR PROBLEM IS NOT SEVERE (ONLY 45 dB DYNAMIC RANGE) AND CAN BE HANDLED BY DIVIDING INTO 2 POWER CONTROL REGIONS WITH THE 37 Km BOUNDARY
- CDMA PROVIDES GOOD RANGING CAPABILITY FOR TRAFFIC CONTROL
- 5 MHz CHANNEL ALLOCATED AS IN TDRS MA SYSTEM

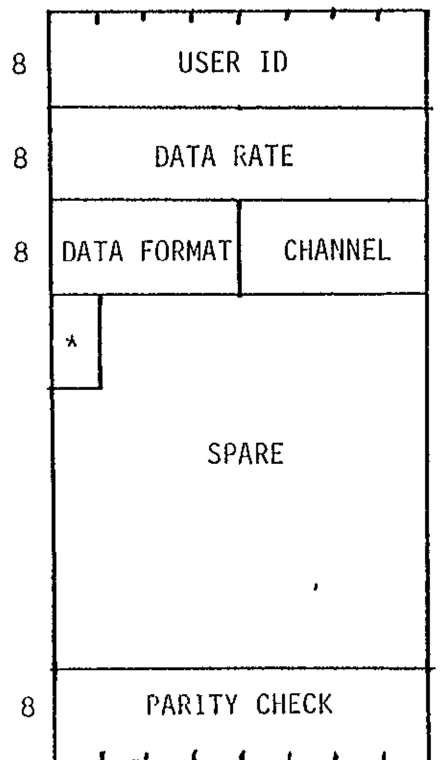
ORDER-WIRE FRAME FORMAT

The 64 bit frame assures that service request can be initiated on the order of seconds at 100 bps. The rest of the roughly 30 bits can be used to transmit location information if for example the user has a GPS receiver on board. Due to the relatively wide beamwidths involved, the location information for antenna pointing do not have to be very accurate. Also, since the link is active all the time, the positon data can be broken up into smaller segments and transmited over many frames, prior to any service request.

LinCom

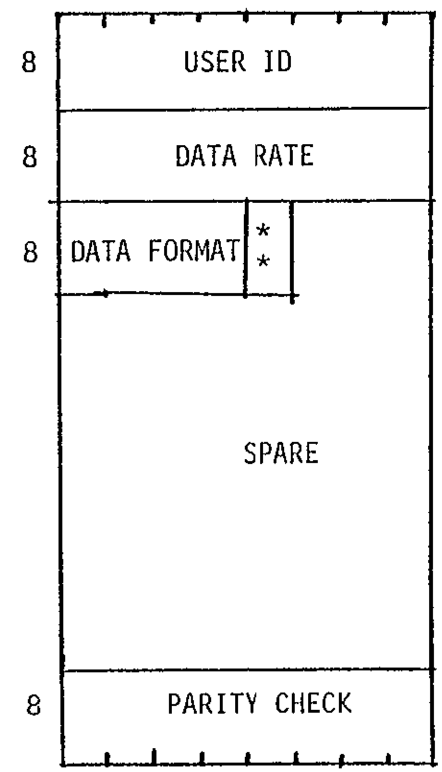
ORDER-WIRE FRAME FORMAT

- FORWARD
  - 64 BIT FRAME



* Clear to Transmit

- RETURN
  - 64 BIT FRAME



**Request to Transmit

LinCom

ADJACENT CHANNEL SEPARATION FOR FDMA

The issue of cost associated with channel filtering is covered by the satellite example. Since commercial satellite communication is multi-user, the technology used must be inexpensive. The channel spacing used reflects realizable, cost effective filter designs. The issue of filtering degradation is addressed by the 7-pole filter example. The 7-pole filter is extremely sharp and represents practically the worst kind of filtering in term of tail-end signal energy suppression and inter-symbol interference. Even in this case, the degradation is tolerable for the relatively narrow (noise)bandwidth shown. The selected spacing for the MA system is wider by close to 50 %.

LinCom

ADJACENT CHANNEL SEPARATION FOR FDMA

- CONSIDERATIONS
  - COST (FILTERING)
  - DEGRADATION
- COMMERCIAL SATELLITE EXAMPLES
  - CHANNEL SPACING/SYMBOL RATE $\leq$ 1.2
  - NOISE BW/SYMBOL RATE $\leq$ 1

PROGRAM	MODULATION	DATA RATE	NOISE BW	CHANNEL SPACING
SBS	QPSK	43 Mbps	43 MHz	49 MHz
INTELSAT (SPADE)	QPSK	64 Kbps	38 KHz	45 KHz

- DEGRADATION
  - 7-POLE CHEBYSHEV FILTERING (SHARP)
  - DEGRADATION < 1 dB at 10^{-6} IF $\begin{cases} BW > 2 \times \text{ data rate (BPSK)} \\ BW > \text{data rate (QPSK)} \end{cases}$

LinCom

LinCom

FORWARD LINK MUTUAL INTERFERENCE

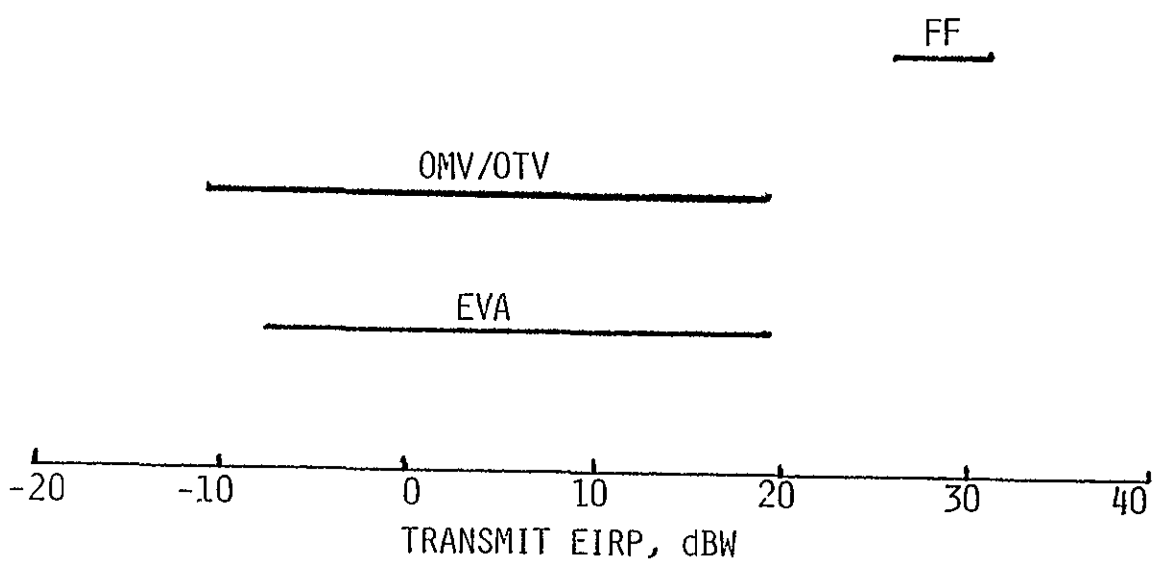
It appears that mutual interfernce is not a severe problem for the forward links. Since the dynamic range for the receiver front end is about 50 db higher than the signal dynamic range, one can safely first downconvert and then filter at IF for the FDMA signals. The values for the transmit power ranges are taken from the link budget charts.

LinCom

LinCom

FORWARD LINK MUTUAL INTERFERENCE

- SIGNAL DYNAMIC RANGE ~ 40 dB
- DYNAMIC RANGE OF TYPICAL LNA ~ 90 dB
    - NF = 3 dB
    - 1 dB COMPRESSION POINT AT 10 dBm
    - 1 GHz NOISE BW



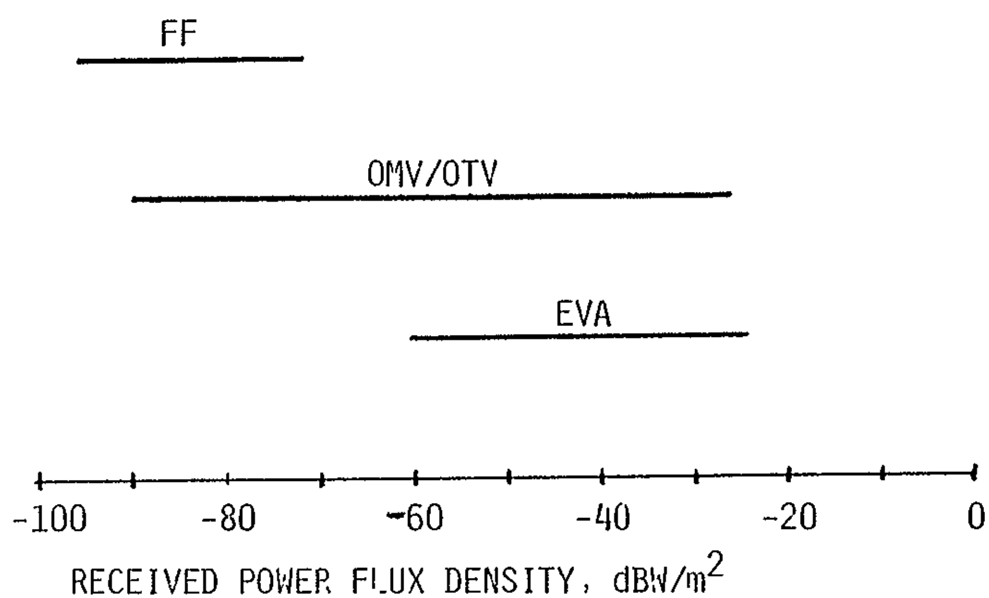
LinCom

RETURN LINK MUTUAL INTERFERENCE

The situation here may require some sort of RF filtering before IF processing. The main problem is the interferences from the high rate TV links to the low rate links. If they are seperated far enough in frequency, then the requirement on the RF filter skirt can be less stringent.

RETURN LINK MUTUAL INTERFERENCE

- SIGNAL DYNAMIC RANGE ~ 70 dB
- ONLY 20 dB OF GAIN CAN BE APPLIED BEFORE FILTERING AND DOWNCONVERSION
- DESIRABLE TO SEPARATE TV AND T&C LINKS IN FREQUENCY PLAN



LinCom

RETURN LINK MUTUAL INTERFERENCE BUDGET

This is a backup for the values used in the previous chart.

LinCom

LinCom

RETURN LINK MUTUAL INTERFERENCE BUDGET

LINK PARAMETERS	EVA T&C	EVA TV 8 NEAR	EVA TV 8 FAR	EVA TV 22 NEAR	EVA TV 22 FAR	OMV/OTV DW	OMV/OTV T&C	OMV/OTV TV 8	OMV/OTV TV 22	FREE FLYER T&C NEAR	FREE FLYER T&C FAR
TRANSMIT POWER, W	0.2	1.0	1.0	1.0	1.0	0.2	1.0	2.0	5.0	5.0	5.0
(DBW)	-7.0	0.0	0.0	0.0	0.0	-7.0	0.0	3.0	7.0	7.0	7.0
TRANSMIT LOSS, DB	-1.0	-1.0	-1.0	-1.0	-1.0	-1.0	-1.0	-1.0	-1.0	-1.0	-1.0
TRANSMIT ANTENNA GAIN, DB	0.0	0.0	0.0	0.0	0.0	0.0	0.0	17.0	17.0	17.0	17.0
POINTING LOSS, DB	-3.0	-3.0	-3.0	-3.0	-3.0	-3.0	-3.0	-3.0	-3.0	-3.0	-3.0
EIRP, DBW	-11.0	-4.0	-4.0	-4.0	-4.0	-11.0	-4.0	16.0	20.0	20.0	20.0
INTENDED PFD, DBW/M/M	-60.0	-39.0	-53.0	-39.0	-53.0	-91.4	-84.4	-64.4	-60.4	-89.0	-95.0
AT RANGE, KM	1.00	0.20	1.00	0.20	1.00	37.0	37.0	37.0	37.0	1000.0	2000.0
INTERFERING PFD, DBW/M/M	-31.0	-24.0	-36.0	-24.0	-36.0	-57.0	-50.0	-30.0	-26.0	-71.4	-86.0
AT RANGE, KM	0.05	0.05	0.20	0.05	0.20	1.0	1.0	1.0	1.0	185.0	1000.0

LinCom

LinCom

MA LINK CHARACTERISTICS

There is no separate allocation for 8 or 22 Mbps TV. TV channels are allocated in 35 MHz chunks and can be used for either one 22 Mbps channel or two 8 Mbps channels. The notation 2/5 means 2 forward and 5 return. Note that the minimum guard band between 2 channels in a given link is about 50 KHz. This is sufficient to handle the 3-5 Khz frequency uncertainty due to the frequency stability of a typical transmitter.

LinCom

MA LINK CHARACTERISTICS

LINK	DATA RATE	MODULATION	CODING*	NOISE BW	CHANNEL SPACING	# CHANNEL	TOTAL BANDWIDTH
OMV T&C	100 Kbps	QPSK	YES	300 KHz	400 KHz	5	2 MHz
EVA HUD**	275+100 Kbps	QPSK	NO	800 KHz	1 MHz	6	6 MHz
FF T&C	100 Kbps	QPSK	YES	300 KHz	400 KHz	20	8 MHz
TV	22 Mbps	QPSK	NO	30 MHz	35 MHz	2/5	70/175
TV	8 Mbps	QPSK	NO	15 MHz	17.5 MHz	4/10	70/175

*RATE 1/2, K=7 CONVOLUTIONAL CODE

** EVA T&C AND HUD USE SAME CHANNEL. EITHER COMBINED 375 Kbps DATA STREAM OR INDEPENDENT I,Q MODULATION

LinCom

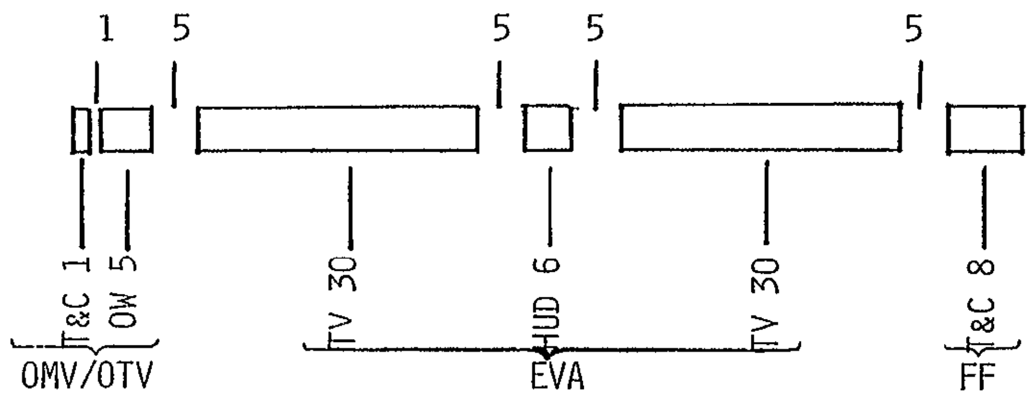
FORWARD LINK CHANNEL ASSIGNMENT

Note the frequency assignment for EVA. An EVA user only needs to work with a contiguous band of approximately 50 MHz (HUD and one of the TV channels). The next chart shows that the return link uses the same arrangement.

LinCom

FORWARD LINK CHANNEL ASSIGNMENT

• TOTAL BW ~100 MHz



• EVA CHANNELS SELECTED TO SIMPLIFY IF

• FF AND OMV/OTV ARE SEPARATED AS THEY HAVE EXTREME TX EIRPs

RETURN LINK CHANNEL ASSIGNMENT

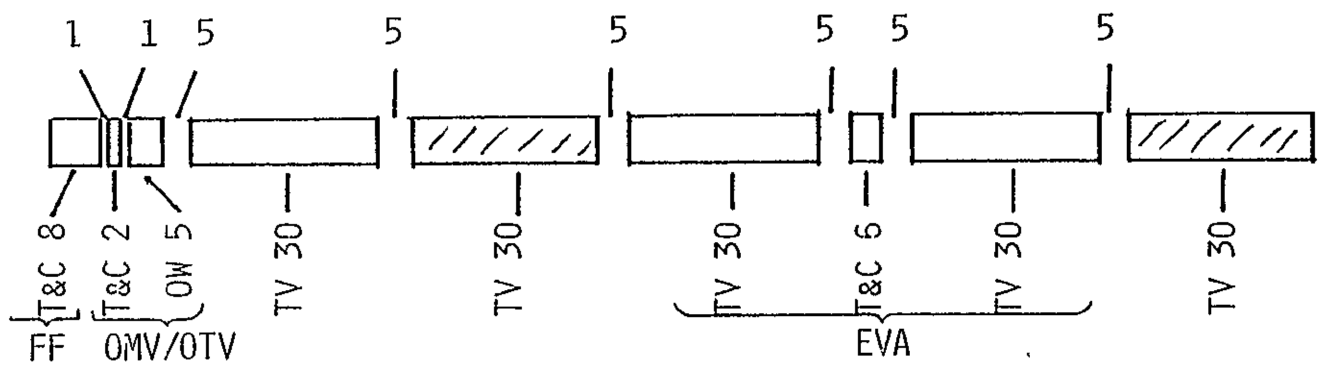
The TV links designated EVA is normally assigned to EVA users. However they can be assigned to other users on a secondary basis. If the required bandwidth is not available, the shaded TV channels are to be eliminated. In this case, there will be more users than available channels at times and the channels must be used on a time-shared basis.

At times it will be worthwhile to split one 22 Mbps channel into two 8 Mbps channels to accomodate more simultaneous users. The channel allocation will be accomplished through the T&C link for the EVA users and the OW link for the OMV/OTV users.

LinCom

RETURN LINK CHANNEL ASSIGNMENT

- TOTAL BW ~ 200 MHz



- FF T&C (LOW RX POWER) IS FARTHEST AWAY FROM TV (HIGH RX POWER)
- OW MORE TOLERANT TO TV INTERFERENCE BECAUSE OF SPREADING
- EVA FORWARD AND RETURN ALLOCATION CHOSEN TO SIMPLIFY IF
- SHADED TV LINKS TO BE ELIMINATED IF BW NOT AVAILABLE

LinCom

GROWTH CONSIDERATIONS

The employment of an ALOHA type of random access technique for the OW channel will virtually make the OW channel available to any user at all times. This is true because of the infrequent service request nature of the MA system - most likely a link will be established for a period of hours, yet infrequently enough so that the chance of collision for a service request is remote.

LinCom

GROWTH CONSIDERATIONS

- CURRENTLY, MA SYSTEM AS CONCEIVED CAN ACCOMODATE ALL USERS DURING IOC AND GROWTH
- EVA AND OMV/OTV
  - NUMBER OF USERS PRIMARILY LIMITED BY THE NUMBER OF MA BEAMS
  - OMV/OTV ORDER-WIRE ALLOWS ADDITIONAL USERS TO TIME-SHARE EASILY
  - FOR EVA, IF FURTHER GROWTH REQUIRES TIME-SHARING THEN ONE HUD CHANNEL CAN BE DESIGNATED AS THE ORDER-WIRE CHANNEL. AN ALOHA SCHEME WILL BE REQUIRED FOR THAT CHANNEL
- FREE FLYER
  - IF FURTHER GROWTH REQUIRES TIME-SHARING, THEN ONE T&C CHANNEL CAN BE DESIGNATED AS THE ORDER-WIRE CHANNEL. AN ALOHA SCHEME WILL BE REQUIRED FOR THAT CHANNEL. IN REALITY, TDRSS WILL BE USED TO ESTABLISH INITIAL CONTACT UNLESS POCC IS ON SS.

LinCom

STRAWMAN SYSTEM

The next set of charts shows a strawman MA system architecture. Its main purpose is to identify the required subsystem blocks and to highlight the potential risk areas.

STRAWMAN SYSTEM

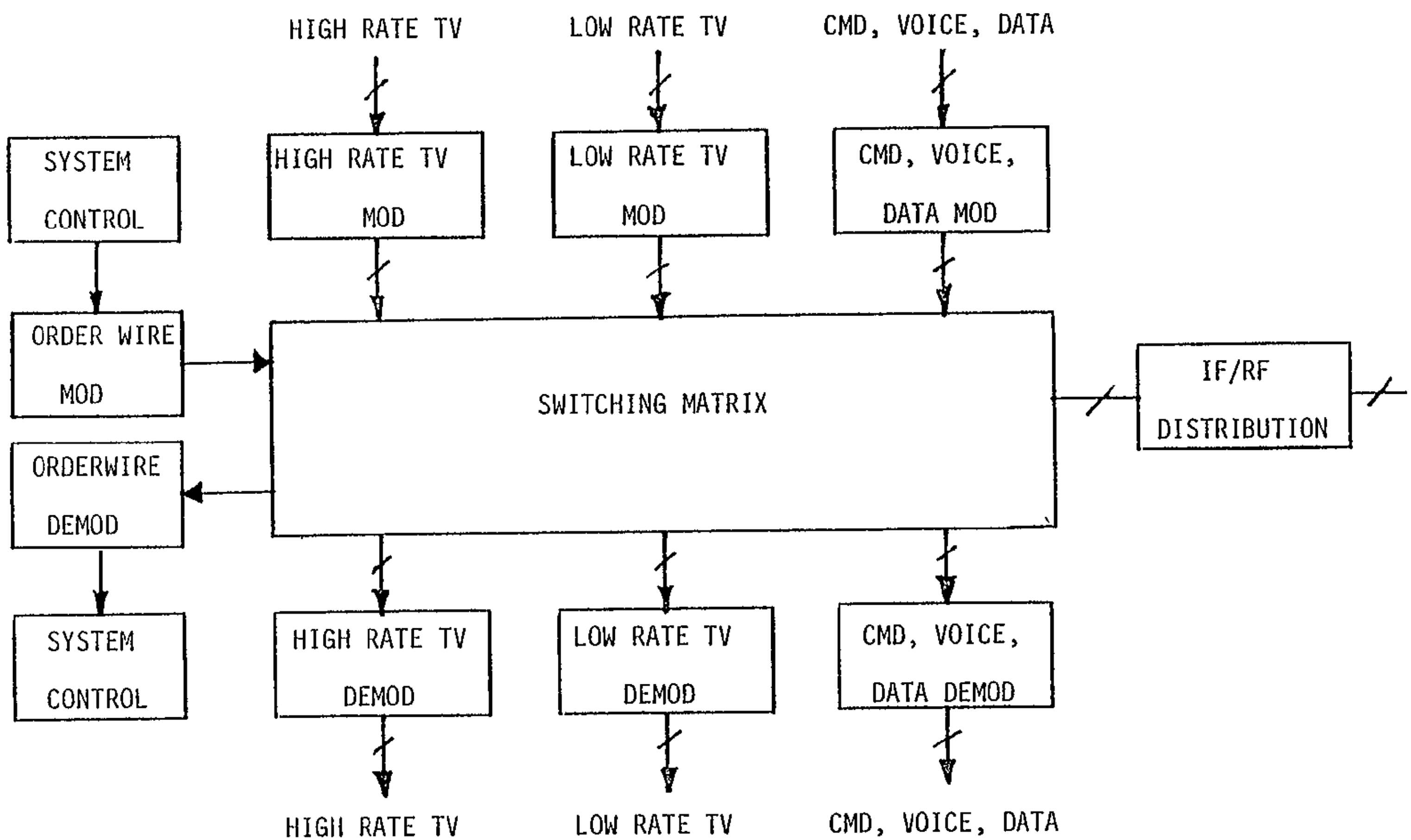
LinCom

MULTI-ACCESS BREADBORAD BASE XCVR

This block diagram highlights the various subsystems of a potential SS breadboard transceiver. Notice that the subsystems themselves are not high technology items; however, because of the sheer amount involved, it will pay to develop perhaps a single modulator/demodulator unit capable of handling the different data rates. The control of the MA system, including the switching matrix may be complicated due to its large number of subsystem elements involved. The choice of IF or RF distribution to the antenna site offers some interesting tradeoffs.

LinCom

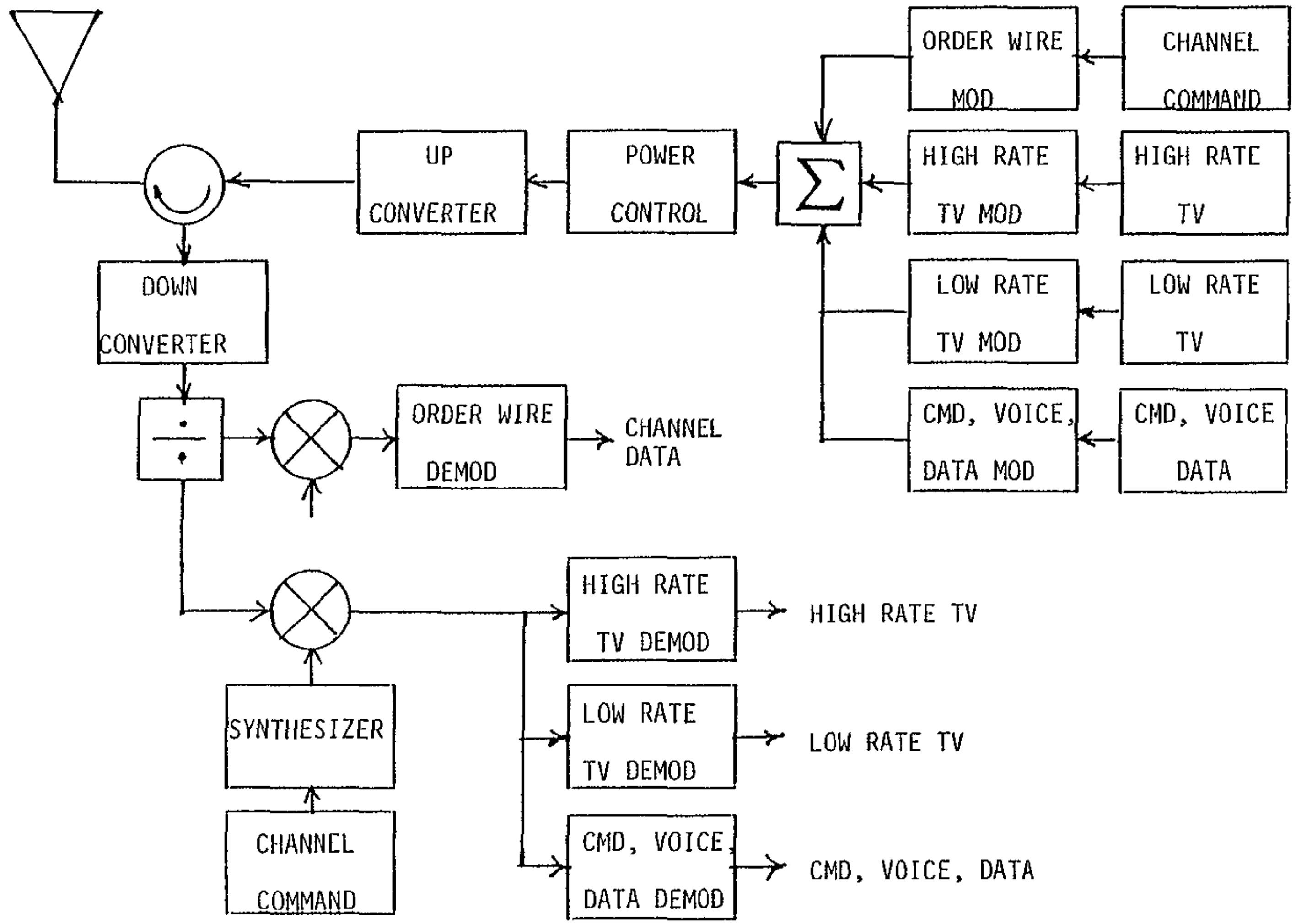
MULTI-ACCESS BREADBOARD BASE XCVR



MULTI-ACCESS BREADBOARD USER XCVR (REMOTE)

This block diagram shows the architecture of a universal user transceiver. A typical user only requires some of the indicated capabilities. Compared to the SS breadboard, the user breadboard is simple. In order to simplify the hardware on the EMU, the synthesizer unit for channel selection may be substituted by switched-selectable crytal sources.

MULTI-ACCESS BREADBOARD USER XCVR (REMOTE)



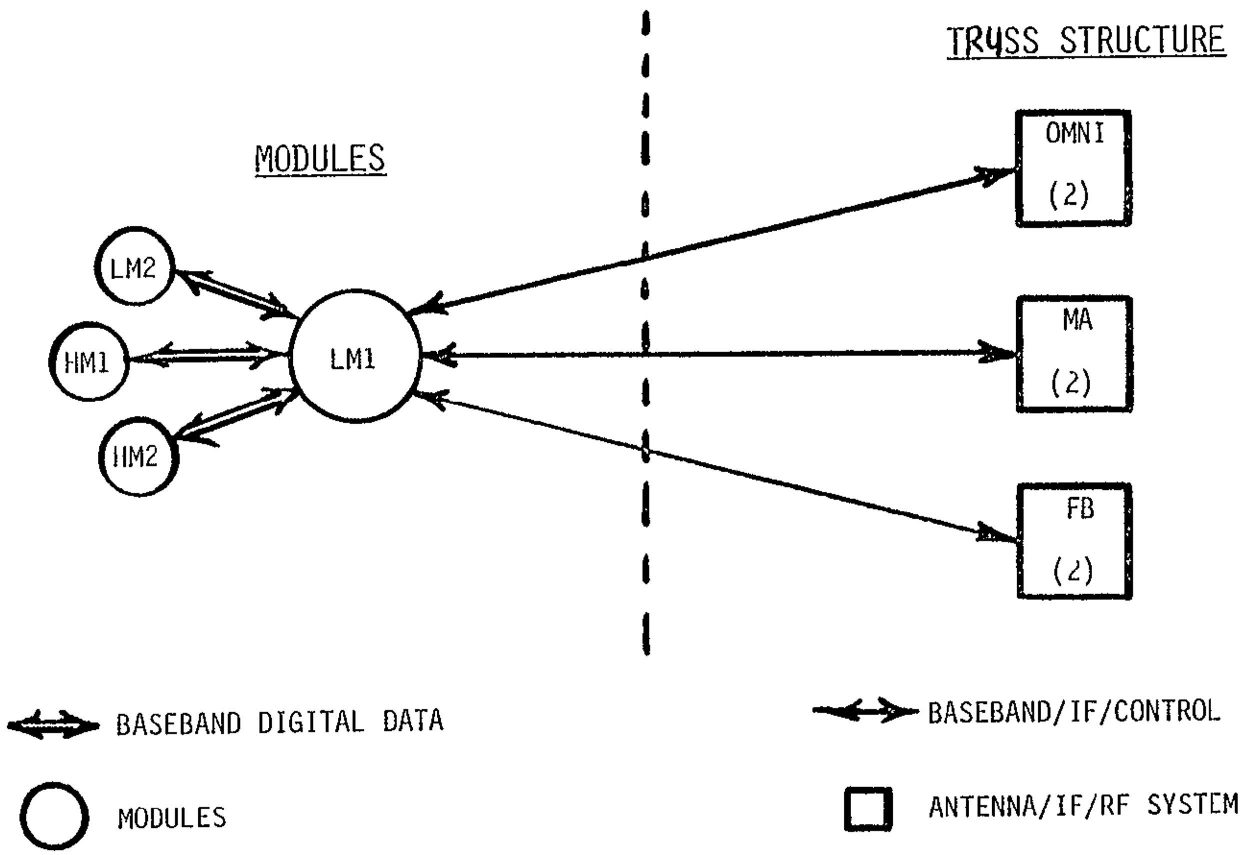
LinCom

MA SIGNAL DISTRIBUTION

The baseband data orginating from the SS modules - Habitat Modules (HM) and Logistics Modules (LM) must be distributed to the RF subsystems on the SS truss structure, and vice versa. The distances between the modules and the RF subsystems can be on the order of a Km so that RF distribution may be lossy and impractical. Note that the omni antenna system may have many gap fillers, although they only handle EVA traffic. The fixed beam system consists of two units, one pointing forward and the other pointing backward. The omni and MA system are assumed to provide hemispherical coverage so that each system has 2 units.

LinCom

MA SIGNAL DISTRIBUTION



BASEBAND VS IF DISTRIBUTION

This comparision indicates that IF distribution may be more practical because of the multiple antenna sites involved.

LinCom

BASEBAND VS IF DISTRIBUTION

SCHEME	ADVANTAGES	DISADVANTAGES
BASEBAND DISTRIBUTION	• SIMPLE DISTRIBUTION • WELL-SUITED FOR FIBER OPTICS	• MULTIPLE MODEMS
IF DISTRIBUTION	• CENTRALIZED MOD/DEMOD	• REQUIRES INTENSITY MODULATION FOR FIBER OPTICS

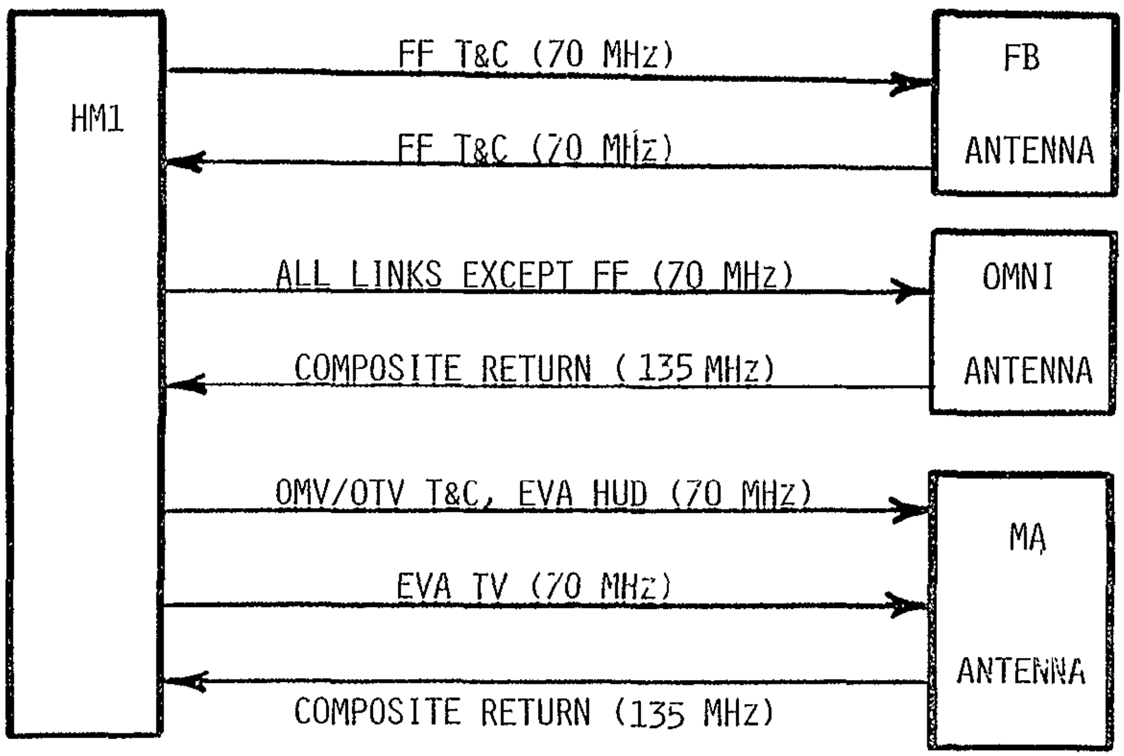
LinCom

STRAWMAN SYSTEM FOR SIGNAL DISTRIBUTION

The IF frequency is chosen to support the required signal bandwidth. Two 70 MHz IF signals are used in the forward link to separate high rate TV data and low rate T&C data. Since the signals are to be multiplexed in frequency and upconverted to RF, a composite 135 MHz IF does not offer any advantages. Instead it forces the multiplexing to take place at a higher frequency. The 135 MHz IF is purposely selected to be a non-integer multiple of 70 Mhz.

STRAWMAN SYSTEM FOR SIGNAL DISTRIBUTION

- FIBER OPTIC OR COAX CABLE



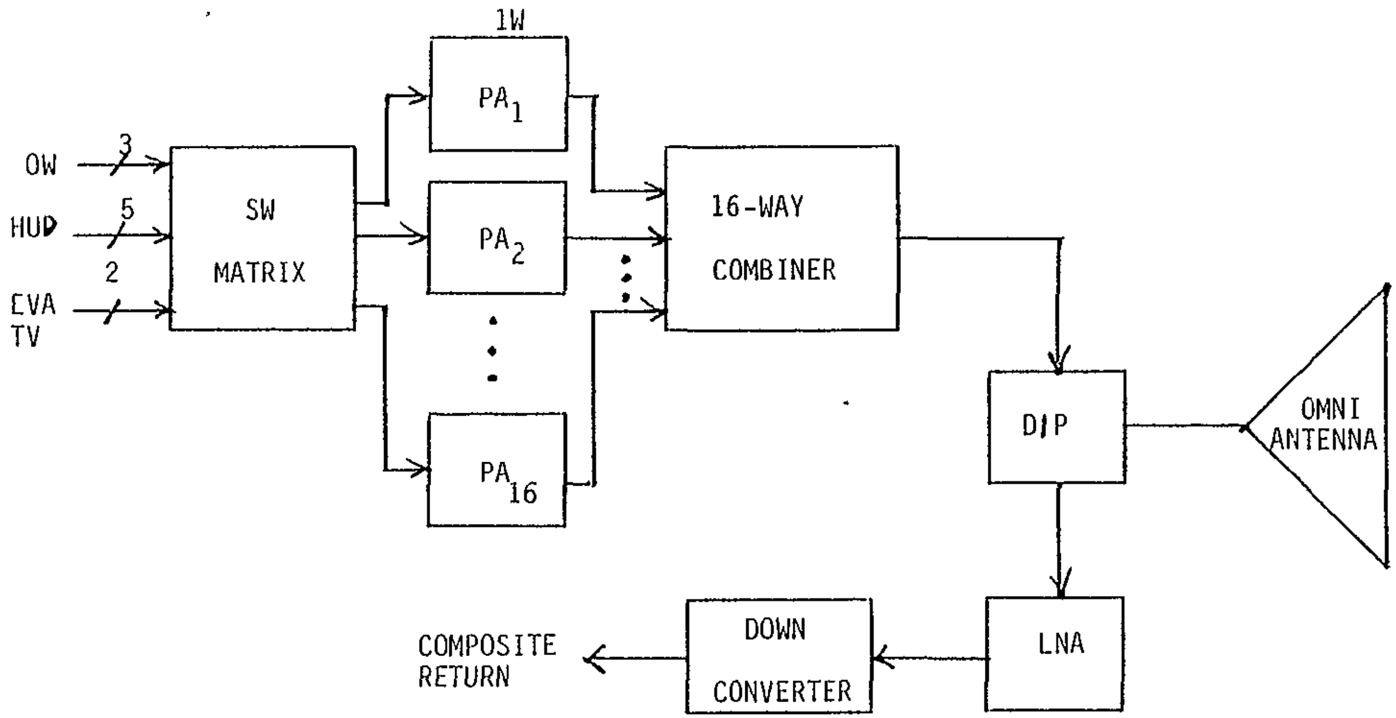
LinCom

OMNI SYSTEM (2 + GAP FLLERS)

Sixteen 1 W PAs are shown to provide some redundancy. The PA outputs can be combined directly since noninterfering FDMA and CDMA signals are involved. Note that there is no requirement for coherent power combining because of the signal format.

LinCom

OMNI SYSTEM (2 + GAP FILLERS)

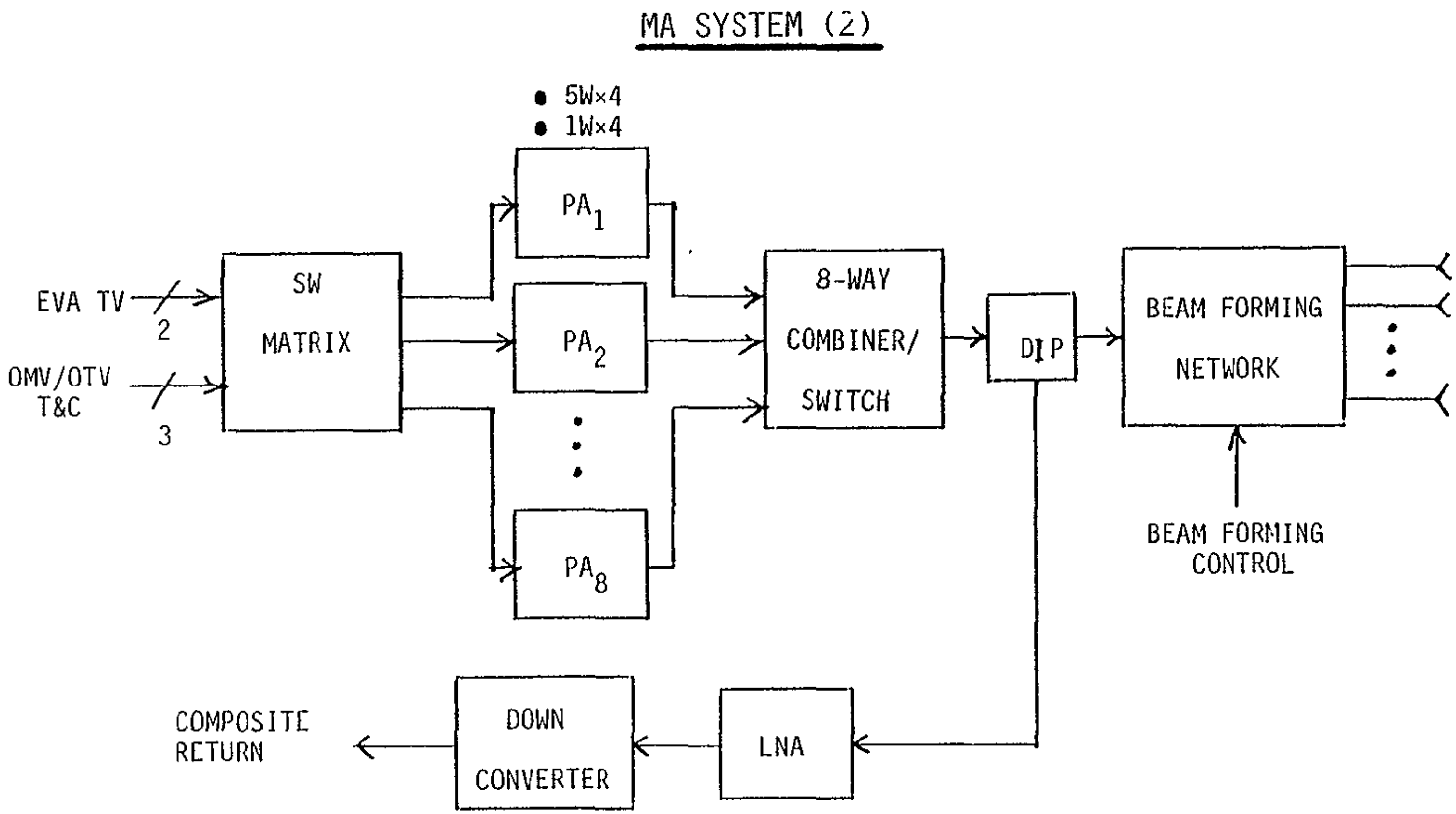


LinCom

MA SYSTEM(2)

The 5 W PAs are for the OMV/OTV links. The 8-way combiner/switch routes the signals to the appropriate port of the beamforming network. Notice that more than one signal may be assessing a particular beam. Again, no coherent power combining is required because of the signal design.

LinCom


MA SYSTEM (2)
● 5W×4
● 1W×4
EVA TV
2
OMV/OTV
T&C
3
SW
MATRIX
PA1
PA2
PA8
8-WAY
COMBINER/
SWITCH
DIP
BEAM FORMING
NETWORK
BEAM FORMING
CONTROL
COMPOSITE
RETURN
DOWN
CONVERTER
LNA

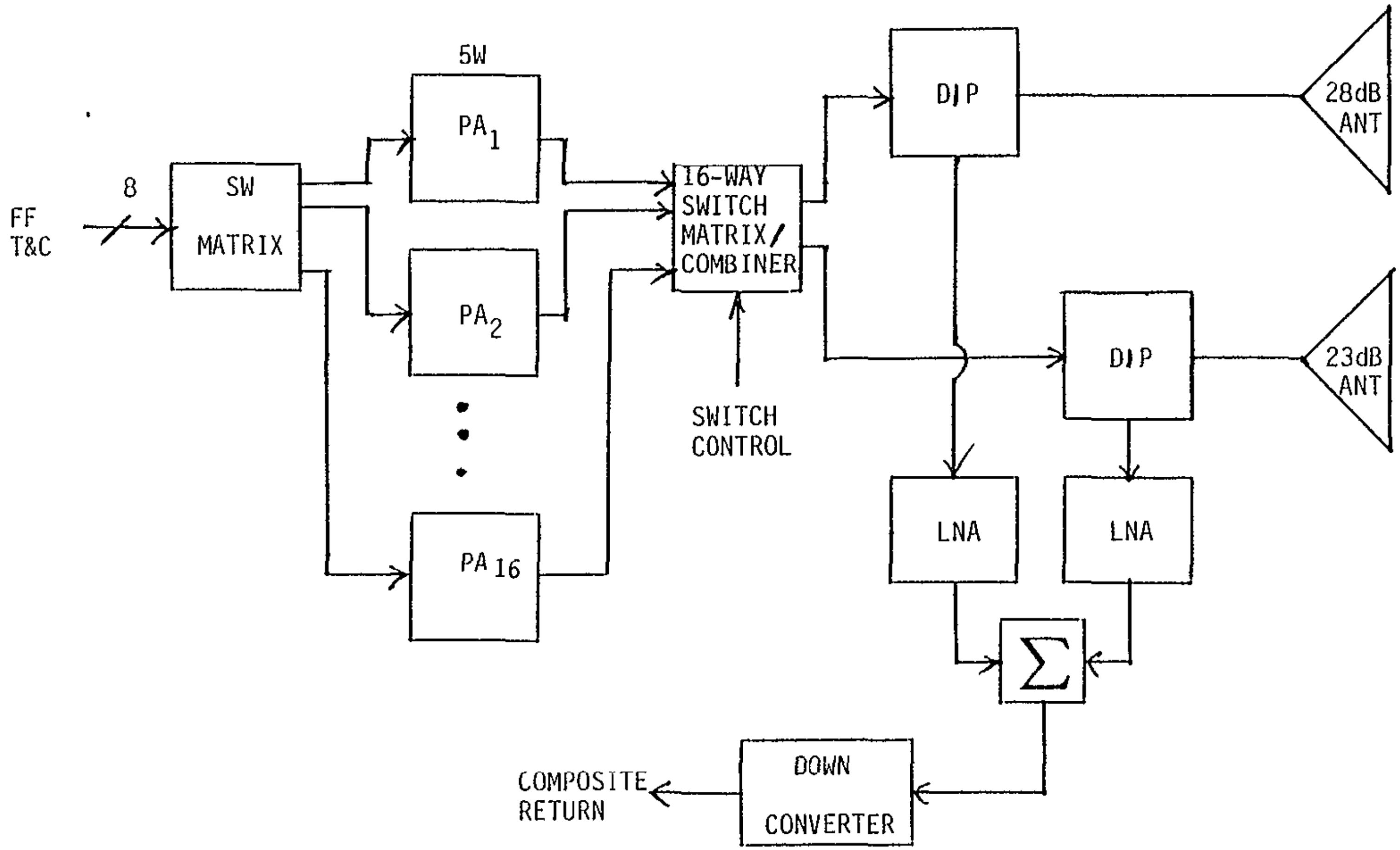
LinCom

FIXED-BEAM SYSTEM (2)

Sixteen 5 W PAs are used to provide redundancy. The combiner/switch combines the appropriate signals into two groups and routes them to the appropriate antenna. Again, no coherent power combining is required because of the signal design.

LinCom

FIXED-BEAM SYSTEM (2)



HARDWARE RISK ASSESSMENT

The hardware involved appears to be of low to medium risk. The medium risk areas are all associated with the MA antenna system. The challenge is to design a complicated system with many subsystems in a cost effective manner.

HARDWARE RISK ASSESSMENT

SUBSYSTEM	RISK
DATA MULTIPLEXER	LOW
CODEC	LOW
MODEM	LOW
IF FILTERING	LOW
IF DISTRIBUTION	MEDIUM
RF FILTERING	LOW/MEDIUM
RF POWER COMBINING	LOW/MEDIUM
OMNI ANTENNA	LOW
FIXED BEAM ANTENNA	LOW
MA ANTENNA	MEDIUM
BEAM FORMING	MEDIUM
BEAM SWITCHING	MEDIUM

SYSTEM ENGINEERING ISSUES

The first two issues can force changes in the modulation selection and frequency plan presented in these charts. If multipath is indeed a problem then diversity with noncoherent detection must be used and can make the system more complicated. Also, the modulation techniques suitable for combatting fading can drive up the system margin required.

If cost is a big driver for the user and if a user must communication with TDRSS and can only affort a single communication system, then there is a lot of incentive to make at least part of the MA system compatible with the TDRSS format. The other operational alternative for the user is to talk to the SS via the TDRSS

What is not treated here in detail is the issue of compatibility with the Shuttle EVA. We have only assigned two return TV channels for the Shuttle EVA and assume that they use the same EVA link design. SInce there is less room in modifying the existing Shuttle than designing the SS from scratch, Shuttle compatibility can conceivably impact the SS MA system design.

LinCom

SYSTEM ENGINEERING ISSUES

- MULTIPATH FADING/ANTENNA BLOCKAGE
- FREQUENCY/POWER FLUX DENSITY COORDINATION
- TDRSS COMPATIBILITY (SSA, KSA)
- SHUTTLE EVA COMPATIBILITY

LinCom

APPENDIX L

COMMENTS ON "SPACE STATION OPERATIONS--OPERATIONAL CONTROL ZONES"

Here are some thoughts and comments on the above document relating to the C&T aspects of the MA System:

1. It appears that the drivers for the control zone concept are primarily user traffic regulation, collision avoidance, rendezvous, and proximity operations requirements. As such it can only be used to define portions of the MA C&T system requirements. The user data requirements will have to come elsewhere.
2. On the second paragraph of Section 3.0. There is a reference to a priori knowledge of target location and is not clear where this information comes from. Is it from ground track?
3. For zones 1 and 2, the Space Station has the capability to (1) track and monitor all vehicle's system health telemetry and (2) command and track all unmanned vehicles. Does one assume that the vehicles must subscribe to the MA system requirements? Or the Space Station will accomodate their normal ground control link characteristics?
4. For zones 3-6, there is a requirement to actively control the satellites. The same questions for item 3 apply.

5. As far as the operating zone concept is concerned, the communication link is for command, control and system health status telemetry. This can be supported with a communications link on the order of 2000 bps. The total system data rates will then be on the order of a Mbps for approximately 100 users. It seems practical to allow the users to use standard S-band TDRSS MA signal formats.
6. It may be helpful if the MPAD people can come up with a few typical mission scenarios that establish the sequences of events while a user is traversing different zones and comment on the C&T requirements.

C-4

APPENDIX M

HIGH RATE CODES FOR SPACE STATION APPLICATION

Summary

The familiar rate 1/2 constraint length 7 convolution code used by TDRSS provides up to 5 dB coding gain at bit error rates around 10^{-6}. This coding gain is achieved at the expense of a 100% bandwidth expansion. This does not cause a problem if the excess bandwidth is available. When the available bandwidth becomes a precious commodity such as the situation encountered by the space station at S- and Ku-band, higher rate codes (with code rate close to 1) can still be used to provide comparable coding gains. This memo represents a preliminary look at a few promising high rate codes whose decoders are of low complexity. Concatenated codes based on these high rate codes can provide a 5 dB gain with a 5/6 code rate. References are included in the appendix.

It is recommended that some sort of high rate codes be employed for all space station communication systems.

1.0 Threshold Decodable Convolutional Codes

Several high rate threshold decodable convolutional codes are currently in use for high rate (up to 100 Mbps) commercial satellite communications. Table 1 summarizes the properties of four such codes.

The performance characteristics of the two example rate 7/8 codes are shown in Figure 2. These codes should provide adequate error correction capability with significant reduction in decoder complexity, as compared to the Viterbi algorithm.

The performance of these codes (as well as others) are traditionally shown in the coding literature as a plot of the decoded bit error rate vs raw channel bit error rate. To convert to the more familiar coding gain, consider the following example. Figure 2a shows that with the DITEC code, a raw channel BER of 7.1×10^{-4} is required for a decoded BER of 10^{-6}. The coding gain at 10^{-6} for an additive white Gaussian channel is determined simply by

$$\begin{aligned} \text{coding gain} &= E_b/N_0 \text{ required for design BER (dB)} \\ &\quad - E_b/N_0 \text{ required for channel BER (dB)} \\ &\quad + 10 \log \text{(code rate)} \\ &= 10.5 \text{ dB} - 7.1 \text{ dB} - 0.6 \text{ dB} \\ &= 2.8 \text{ dB} \end{aligned}$$

2.0 Reed-Solomon Codes

The advantages of Reed-Solomon codes are well-known. Recently, Cyclotomics has been able to build low complexity, high speed decoders using special purpose microprocessors to perform the required time consuming Galois Field arithmetic. Figure 3 shows the performance of their 120 Mbps rate 243/255 decoder. The coding gain at 10^{-6} is 10.5 - 7.3 - 0.2 = 3dB.

3.0 Punctured Convolutional Codes

Punctured convolutional codes are obtained by periodically deleting a part of the output symbols of low-rate convolutional codes. For example, rate $(n-1)/n$ $(n=3,4,\ldots,14)$ can be derived from a rate 1/2 code. The coding gain performance for a rate 7/8, K=7 code is about 3.8 dB at 10^{-6}.

Since there is enough market for the rate 1/2, K=7 Viterbi decoders, manufacturers (e.g. Harris and TRW) are beginning to develop and market a codec chip using VSHIC technology operating at about 5 Mbps, for about 500 dollars. A number of these chips can be multiplexed to form a high speed codec using the punctured concept. The cost and complexity for this scheme can be very competitive.

4.0 Concatenated Codes

Even though the above coding schemes only give moderate coding gains, i.e., around 3 dB, concatenating them is a viable technique to provide higher gains. Based on published data on concatenation using an RS outer code (encoded first) and a Viterbi decoded convolutional inner code, it is anticipated that the concatenated coding gain is roughly equal to the sum of the individual gains, or about 5 dB. Using the rate 7/8 and the rate 243/255 example, the concatenated code rate is $0.83 \approx 5/6$.

Table 1. High Rate CSOC's

	Code Rate R	Constraint Length n_A	Effective Constraint Length n_e	Number of Syncrome Equations J	Minimum Distance dim
SPADE	3/4	80	31	4	5
INTELSAT SCPC	7/8	376	71	5	5
DITEC	7/8	1176	148	6	7
TDMA	8/9	1233	81	4	5

LinCom

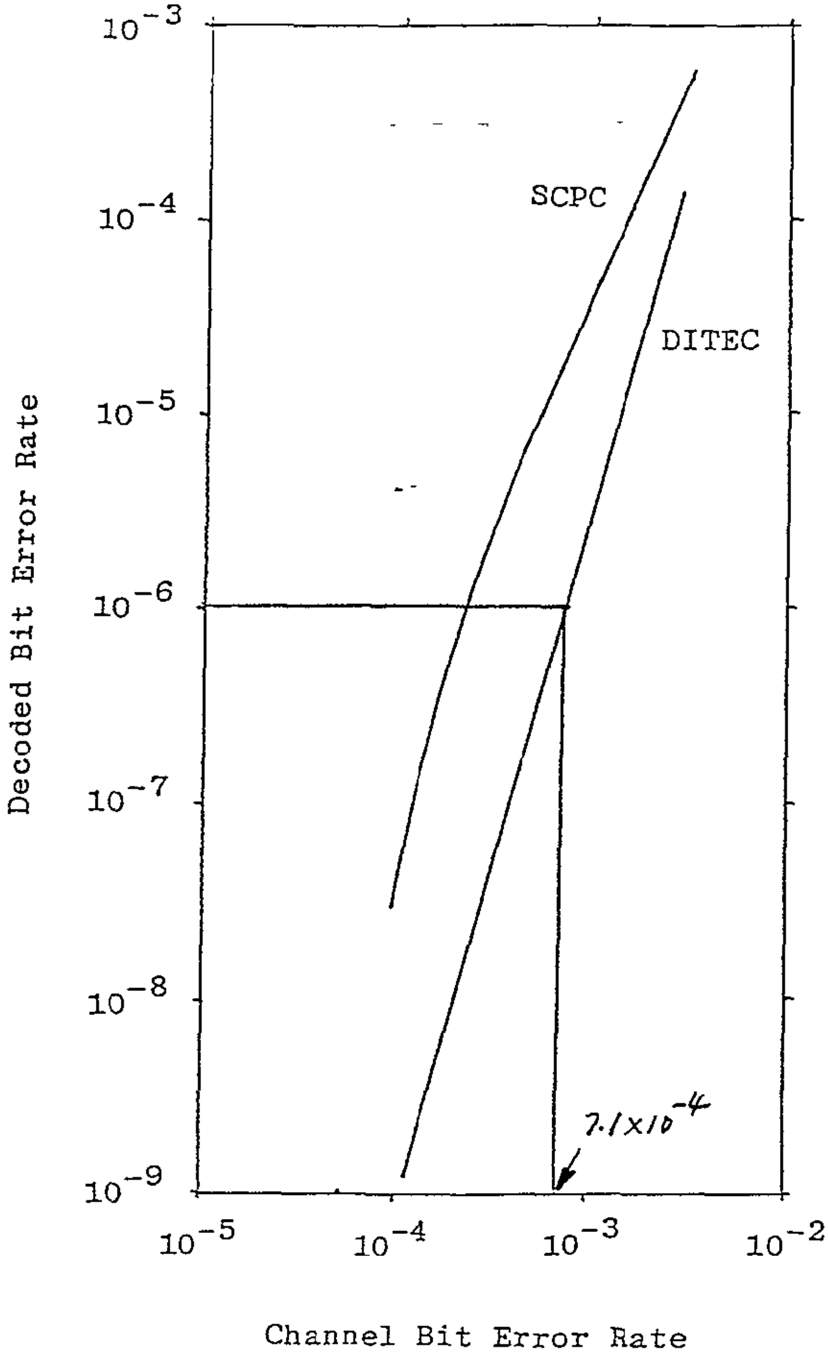


Figure 2a Error Rate of Rate 7/8 Codes with Threshold Decoding

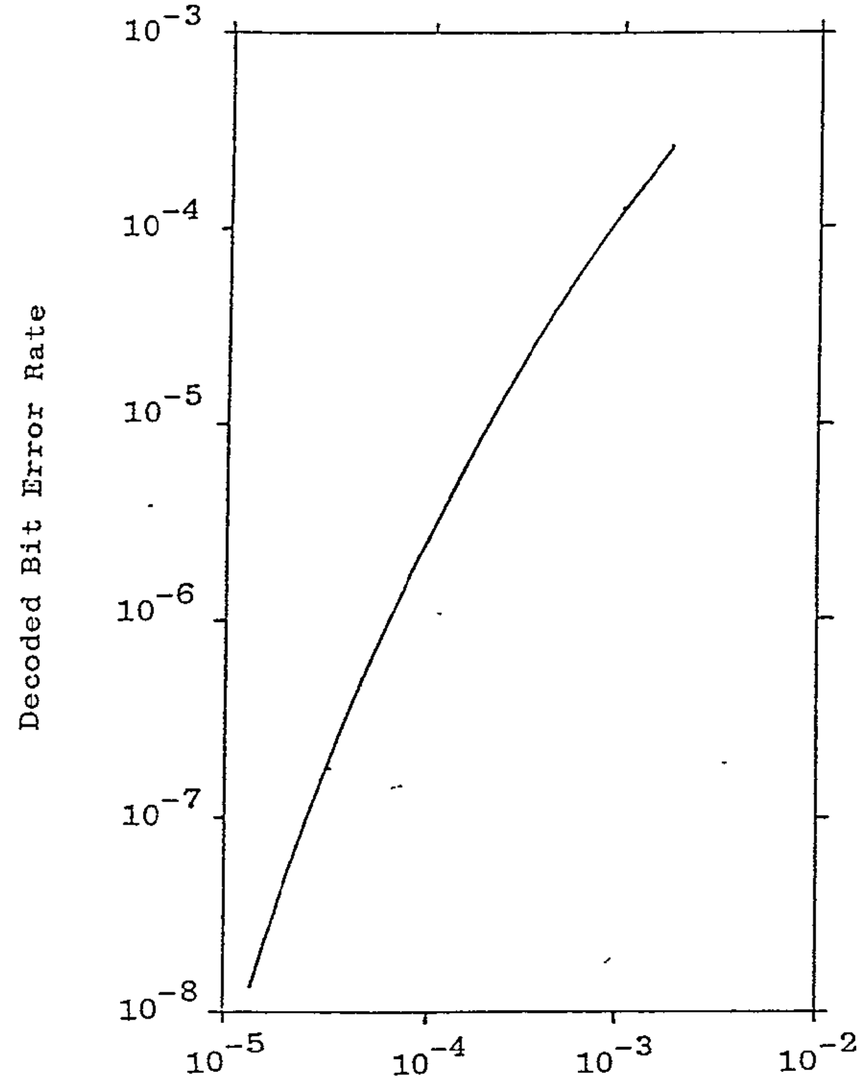


Figure 2b Error Rate of Rate 8/9 Code (TDMA) with Threshold Decoding

RS, m=8; n=255

4.7% Redundancy

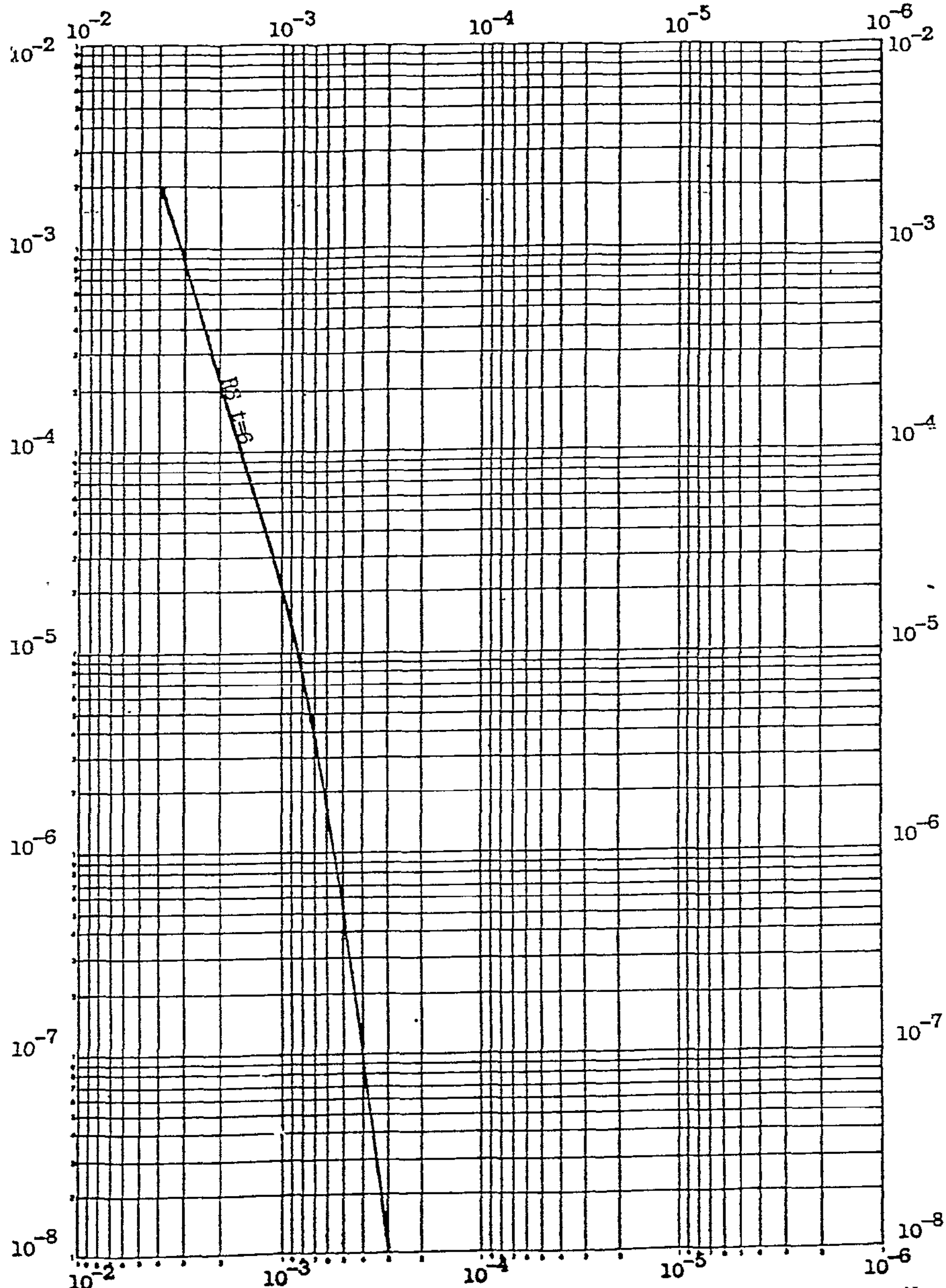


Figure 3. Performance Curve for Cyclotomics' Model 120 GF1® Decoder

ORIGINAL PAGE IS
OF POOR QUALITY

APPENDIX N

A REVIEW OF THE TDRSS USERS GUIDE, REV. 5

SUMMARY

The purpose of this memo is to review the current operational capabilities of the TDRSS pertaining to the Space Station as defined by the latest revision (Rev. 5) of the Users' Guide. In general Rev. 5 is a much expanded version of the previous one. In particular a lot of material on the operational aspects of the TDRSS has been added. As far as the Space Station is concerned, the most noticeable difference between the two versions is the absence of the IF service in the current revision.

The current capabilities are as follows:

	FORWARD	RETURN
MA	10 KBPS	50 KBPS
SSA	300 KBPS	3.15 MBPS
KSA	25 MBPS	300 MBPS

However, it appears that the TDRSS system bandwidth should be capable of supporting the following SA services:

	FORWARD	RETURN
SSA	50 MBPS	25 MBPS
KSA	75 MBPS	560 MBPS

assuming the IF service is used and a modulation/coding scheme with a

throughput of 2.5 bits/Hz can be used. Whether this goal of 2.5 bits/Hz can be achieved under the TDRSS type of environment will be the subject of further study.

1.0 Maximum Data Rates Supported

1.1 Forward Services

The maximum data rates for the MA and SA services are summarized in Table I. The forward link data rates are for uncoded data using NRZ format. For data rates not greater than 300 Kbps, the data is spread with a 3 Mhz PN code. (Data is transmitted on the command channel only. There is another quadrature range channel transmitted with a different 3 Mhz PN code). This applies to all MA and SSA users. This also applies to the low rate KSA users. The main purpose of this spreading is to satisfy the NTIA imposed flux density restrictions. For KSA data rates greater than 300 kbps, direct BPSK modulation is used and the range channel is not transmitted.

The MA and the SSA forward services are already supporting 6 Mbps QPSK data if one considers the command and range channnel PN chips to be data. The MA forward service can only support one user at a time. This is also true for the SA users. Hence the operational constraints are mainly the user G/T and the flux density restrictions, but not interference to other users.

1.2 Return Services

Both the MA and SSA links are <u>required</u> to be Rate 1/2 convolutionally encoded. The purpose of the encoding (together with interleaving) is to minimize RFI degradation at S-band. The MA service can accomodate 20 users simultaneously using (PN) code division multiplexing. The convolutional encoding also helps to minimize the interference among MA users by reducing the required EIRP through the coding gain. Since RFI is not a major problem at K-band, the KSA link does not have to be coded.

Although the MA channel can potentially support more than 50 Kbps, it will be difficult to do so operationally. This is because a high rate user, with its higher EIRP will interfere with other users of the MA service. This is quite different from the single user MA forward link considered. Hence if the SS is to use the MA link, it must conform with the Users' Guide requirements. This may not be too bad since the MA is probably going to be used as a backup T&C link.

The SSA link consists of two quadrature channels. One channel carries 3 Mbps NRZ rate 1/2 encoded data (6 Msps). The other channel can either be spread (DG1 mode 3) or unspread (DG2) and carries 150 Kbps rate 1/2 coded data. According to a conversation with GSFC, the main reason behind this 150 Kbps restriction was purely operational: They did not want to procure the additional 3 Mbps de-interleaving and decoding hardware at WSGT. However, since the release of Rev. 5, they found out that the cost of procuring the additional hardware is minimal and they are planning to accomodate 3 Mbps on both the I and Q channel for DG2 users. This will be reflected in a later version of the Users' Guide. With this being the case, the SSA channel can therefore carries 6 Mbps NRZ data in its normal mode of operation. Since the rate 1/2 coding is only used to mitigate RFI, if the SS can provide enough EIRP for RFI degradation margin, then it can bypass the coding and transmit 12 Mbps uncoded QPSK NRZ data through the SSA channel.

The KSA supports 300 Mbps uncoded QPSK NRZ data.

2.0 Channel Bandwidths

Table II summarizes the 3 db RF bandwidths associated with the services. The bandwidths for the forward services are covered by Rev. 5. The bandwidths for the return services are from the previous version, Rev. 4. They have been deleted from the new Users' Guide because the IF service has been eliminated. The 3 db bandwidths are considered to be parameters pertaining to the IF service only. The MA return bandwidth was not given because the signal must be processed at WSGT by the MA processor for beamforming so that the IF signal is virtually useless. According to [1, p.8-16 to 8-26], the bandwidths are set by the bandpass filters/diplexers in the forward and return processor onboard the TDRS spacecraft.

2.1 Data Rate Supportable by the Channels.

If a channel is linear, then [2] showed that the QPSK modulation format can support about 1.6 bits/Hz and the 8PSK format can support about 2.4 bits/Hz with a few tens of db degradation under certain idealized models. Of course, the TDRSS is not an idealized channel and the performance will be worse. However, there are other types of so-called bandwidth efficient schemes such as the MSK which are more suited for the TDRSS type of satellite repeater nonlinear bandwidth limited channnels. In addition, high rate encoding can also be used to achieve an acceptable BER. (In a nonlinear bandlimited channel, the symbol error rate curve tends to flare out. That is, it will reach a point such that additional increase in EIRP can no longer reduce the SER. This is the system irreducible error due to signal related noise and distortions. If this noise floor occurs at an unacceptable SER, coding can be used to provide an acceptable BER e.g. if the SER floor is

10^{-4} a code that can provide "3 db" coding gain can bring the BER to 10^{-7}.) Hence, realistically, one can expect to be able to find modulation/coding combinations that can support between 2 to 3 bits/Hz. This may involve higher degradations than ideal BPSK transmission. However, it appears that size, weight and power are less of a problem with the SS so that these degradations can be overcome by increasing the SS transmit EIRP and G/T.

3.0 IF service

Although the IF service has been eliminated in the Rev. 5 Users' Guide, the capability still exists. In order to provide high data throughput using the existing TDRSS facilities, what is needed for the SS is a better characterization of the IF service, perhaps through negotiation with the GSFC. According to [3, p.2-19, 2-14], the SA users are provided with a 300 Mhz IF for the forward service and a 370 Mhz IF for the return service. The salient characteristics for the IF services are tabulated in table 3-3 and 3-6 of Rev. 4 of the Users' Guide.

4.0 Domsat links

The current capability to provide return link transmission form the NASA ground terminal at White Sands to user ground facilities at the JSC and GSFC is 48 Mbps through a DOMSAT transponder. To support the SS, this capability must be expanded. This may come in the form of more leased transponder. However, because of the expense involved with these transponders, it may be worthwhile to either move the SS POCC to White Sands or allow the TDRS to provide space ground links to either the JSC or GSFC.

REFERENCE

[1] TDRSS System Design Report, Vol.III: Space Segement. TRW 29000-200-003-001, 1 August 1978.

[2] V. K. Prabhu," Error Probability of M-ary CPSK Systems with Inter-symbol Interference," IEEE Trans. Comm. Tech., Feb. 1973, p.97-109.

[3] TDRSS System Design Report, Vol.II: Ground Segement. TRW 29000-200-003-001, 1 August 1978.

Table I. Maximum Data Rates (TDRSS Operational Format Per Rev. 5).

SERVICE	FORWARD	RETURN
MA	10 Kbps[1]	50 Kbps[2]
SSA	300 Kbps[3]	3.15 Mbps[4]
KSA	25 Mbps[5]	300 Mbps[6]

Notes:

[1]Uncoded

[2]Rate 1/2 Coded

[3]Uncoded

[4]Rate 1/2 Coded

[5]Uncoded

[6]Uncoded

Table II. Channel Bandwidths (3 dB).

SERVICE	FORWARD	RETURN*
MA	6 MHz	--
SSA	20 MHz	10
KSA	50 MHz	225

*Rev. 4