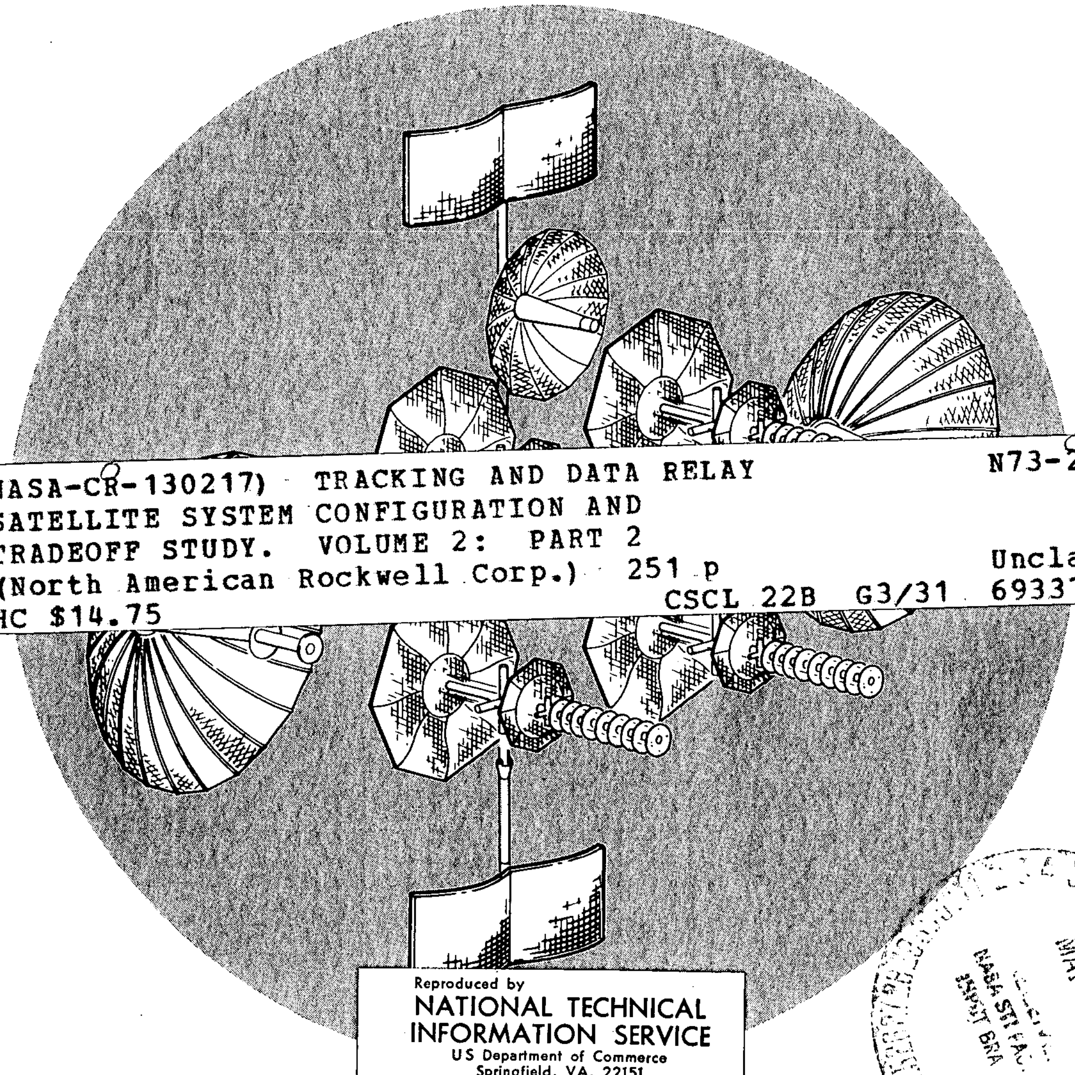
SD 73-SA-0018-2

PART II FINAL REPORT

TRACKING & DATA RELAY SATELLITE SYSTEM CONFIGURATION & TRADEOFF STUDY

VOLUME II TELECOMMUNICATIONS DESIGN (PART II)



APRIL 1973

SUBMITTED TO
GODDARD SPACE FLIGHT CENTER
NATIONAL AERONAUTICS & SPACE ADMINISTRATION

Space Division
Rockwell International

IN ACCORDANCE WITH
CONTRACT NAS5-21705

12214 Lakewood Boulevard, Downey, California 90241

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TRACKING & DATA RELAY SATELLITE SYSTEM CONFIGURATION & TRADEOFF STUDY

VOLUME II
PART II TELECOMMUNICATIONS DESIGN

T. E. Hill
TDRS STUDY MANAGER

Details of illustrations in this document may be better studied on microfiche

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FOREWORD

This report summarizes the results of the study conducted under Contract NAS5-21705, Tracking and Data Relay Satellite Configuration and Systems Trade-Off Study--3-Axis Stabilized Configuration. The study was conducted by the Space Division of North American Rockwell Corporation for the Goddard Space Flight Center of the National Aeronautics and Space Administration.

To ensure that the NASA would receive the most comprehensive and creative treatment of the problems associated with the definition of an optimum TDRS system concept, North American Rockwell entered into subcontracting agreements with the AIL Division of Cutler-Hammer and the Advanced Systems Analysis office of Magnavox. In this teaming relationship NR performed as the prime contractor with responsibility for study management, overall system engineering, TDR spacecraft and subsystem design, network operations and control, reliability engineering, and cost estimating. AIL was responsible for RF link analysis, the on-board telecommunications subsystem design and ground station RF equipment design. Magnavox was responsible for telecommunications system analysis, user spacecraft terminal design, and ground station signal processing.

The study was in two parts. Part I of the study considered all elements of the TDRS system but emphasized the design of a 3-axis stabilized satellite and a telecommunications system optimized for support of low and medium data rate user spacecraft constrained to be launched on a Delta 2914. Part II emphasized upgrading the spacecraft design to provide telecommunications support to low and high, or low, medium and high data rate users, considering launches with the Delta 2914, the Atlas/Centaur, and the Space Shuttle.

The reporting for both parts of the study is as follows:

Part I SD 72-SA-0133	Part II SD 73-SA-0018
1. Part I, Summary (-1)	1. Study Summary (-1)
2. System Engineering (-2)	2. Telecommunications Design (-2)
3. Telecommunications Serivce System (-3)	3. Spacecraft Design (-3)
4. Spacecraft and Subsystem Design (-4)	4. Cost Estimates (-4)
5. User Impact and Ground Station Design (-5)	
6. Cost Estimates (-6)	
7. Telecommunications System Summary (-7)	

This report consists of four volumes: Volume I, Study Summary: Volume II, Part II Telecommunications Design; Volume III, Part II Spacecraft Design, and Volume IV, Study Cost Analysis; This volume summarizes the activities and results of both Part I and Part II, as does Volume IV. The detailed technical material developed during Part I was extensively reported in the Part I Final Report and is not repeated in this report. Volumes II and III are technical reports covering only Part II. The reader is referred to the Part I Final Reports for detailed considerations of mission analysis, network operations and control, telecommunications system analysis, telecommunications subsystem design (baseline), spacecraft mechanical and structural design (baseline), spacecraft subsystem design and analysis, reliability, user spacecraft impact (baseline), and ground station design: except as they were influenced by Part II design and analysis activities.

Acknowledgement is given to the following individuals for their participation in and contributions to the conduct of this study:

North American Rockwell

M. A. Cantor	System Engineering and Spacecraft Design--Project Engineer
A. A. Nussberger	Electrical Power
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R. E. Oglevie	Stabilization and Control
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D. M. DeVito	Telecommunication System Analysis
D. Cartier	Ground Station Design
R. H. French	Operations Analysis
G. Shaushanian	User Transponder Design

In addition we acknowledge significant support and contributions from Mr. Peter Sielman of AIL and Dr. Neil Birch of Magnavox.

TABLE OF CONTENTS

LIST OF ILLUSTRATIONS

LIST OF TABLES

1.0 INTRODUCTION AND SUMMARY

This Part II phase of the study provides preliminary designs for alternate TDRS Telecommunication Systems for implementation on the Delta 2914, Atlas Centaur and Space Shuttle launch vehicles. In general, this phase includes a requirement to support a new category of spacecraft users identified herein as the High Data Rate (HDR) user, and as the name implies, requires a much higher return data rates (H = 100 Mbps) than its Part I predecessor, viz. the Low Data Rate (LDR) (1-10 kbps) and Medium Data Rate (MDR) (10 kbps to 1 Mbps) users. This addition plus modifications to the basic design requirements of the initial Part I phase has resulted in considerable improved link performance, service, and support capabilities that is offered with the Part II - Alternate TDRS Telecommunication System Configurations.

Part II design approach reflects two fundamental system configurations; viz:

- Uprated Delta 2914* Configuration which provides a multiple access capability to simultaneously support 20 LDR + 1 MDR + 1 HDR user.
- High Performance Atlas Centaur/Space Shuttle Configuration which provide multiple access capability to simultaneously support 20 LDR + 3 MDR + 1 HDR users with considerably greater link performance, operational flexibility, and overall system reliability.

In addition, 1 and 3 Uprated Delta 2914 Configurations can be launched simultaneously by the Atlas Centaur and Space Shuttle, respectively, to provide an alternative launch configuration for the larger boost vehicles.

1.1 UPRATED DELTA CONFIGURATION: TDRS SPACECRAFT

The Uprated Delta configuration is functionally similar to the Part I base line design, but employs larger MDR/HDR antenna (3.8 meter) and TDRS/GS antenna (1.8 meter), thereby considerably increasing the support capabilities to the MDR and HDR users. In addition, since the forward voice requirement was eliminated in this Part II phase for the LDR space-to-space link, it permitted this premium size, weight and prime power demands to be used to enhance the MDR and HDR user support capabilities. This configuration provides service and support as follows:

*Uprated Delta 2914 term as used herein refers to the Uprated TDRS Telecommunication System (as compared to Part I Baseline) for launch on the Delta 2914.

- LDR: Sequential commands to 20 users in the forward link; and multiple access to 20 users simultaneously in the return link.

- MDR: Multiple access to 1 or 2 (if no HDR) simultaneous users in forward and return link.

- HDR: Forward and return link support to 1 user (Assume only 1 MDR.)

In the LDR forward link, the prime mode is F-FOV mode which is used to simultaneously illuminate the entire 31° FOV with a fixed beam. In this mode, a single transmit channel excites a high gain disc-on-rod antenna to provide an EIRP of +30, +27 or +24 dBW. The reduced EIRP's are used to conserve the prime power demands during eclipse and other periods when prime power must be reduced.

This F-FOV beam is used to sequentially transmit commands to the 20 LDR users, but also to provide a continuous signal to all LDR users within the FOV to lock-up and synchronize their on-board PN code generator and frequency source. This lock-up is required to enable range and range rate measurements to be made to each LDR user. Four identical transmit channels are used to provide quadruple functional redundancy in the F-FOV mode or the 4 channels can be used as a phase array to provide an Emergency Steered Beam Mode with EIRP of +36, +39 or +42 dBW. This higher EIRP of up to +12 dBW can be effectively used to aid command transmissions to a user in emergency condition (e.g., tumbling state) or when the user is in the midst of high RFI environment.

In the LDR return link, an adaptive signal processing technique coined Jr. AGIPA (small aperture Adaptive Ground Implemented Phased Array) is utilized that employs adaptive spatial and polarization discrimination to reject interference signals, including unintentional RFI emitters as well as intentional in-band other user signals. As the name AGIPA implies the usual complexities associated with a multiple access phase array is located on the ground; thereby minimizing the size, weight and prime power demands as well as the system complexities on the TDRS relay spacecraft. RFI model syntheses conducted in Part I has shown the Jr. AGIPA provides upwards of 5 to 18 dB improvement of signal-to-interference ratio over a F-FOV design, in the presence of an interference environment. The LDR return link can also operate in the F-FOV backup mode by employing two orthogonally polarized receiver channels, or by using the 8 receive channels in an unadapted mode; viz. with the beam pointed along the spacecraft local vertical. In this backup mode however, the link performance is limited since all RFI and in-band intentional interference signals are received without discrimination.

The MDR and HDR functions have been combined into one dual frequency transponder as in Part I and is referred to herein as the MDR/HDR Transponder. Two such MDR/HDR Transponders employing 3.8 meter deployable reflector antennas are on-board, and can be used to simultaneously support:

- Return link: 2 MDR and 1 HDR user (assumes common MDR/HDR user on one transponder).
- Forward link: 2 MDR or 1 MDR and 1 HDR user.

Each transponder operates at S-band and K_u-band to support an MDR and HDR user, respectively. The MDR/HDR transponder can support:

	Data Rates	
Link	MDR	HDR
Forward		
• Unmanned	3 kbps (1)	1 kbps
• Manned	150 kbps (2)	Video
Return		
• Unmanned	1 Mbps	100 Mbps (4)
• Manned	250 kbps (3)	100 Mbps (4)

Each MDR/HDR Transponder is also designed to provide a functional backup redundancy to the TDRS/GS Transponder. In addition, one of the 3.8-meter antennas can be switched in lieu of the 1.8-meter TDRS/GS antenna to provide an additional 6.5-dB rain margin for the space-to-ground link to provide a total of 24-dB system margin for operation in rain.

The 3.8-meter MDR/HDR antenna provides an HPBW of approximately 0.4 degree at K_u-band and 2.6 degrees at S-band; consequently a closed loop pseudo-monopulse system is employed for acquisition and track of the HDR user spacecraft; and open-loop tracking via ground commands is used for acquisition and track of MDR users.

---

1. With 0 dBi antenna gain at user.
2. With FEC Coding gain of 3.8-dB (54 kbps required).
3. With user EIRP of 16 dBW (192 kbps required).
4. With user antenna gain of 41 dBi (0.91 meter aperture).

The TDRS/GS Transponder uses a 1.8-meter fixed parabolic reflector antenna with a HPBW of approximately 0.8 degree. This antenna is designed with a closed loop auto-track system with a 4 horn pseudo-monopulse K_u-band feed. The TDRS/GS Transmitter is an FDM/FM/FDM system, employing an HDR channel for the 100 Mbps HDR data and a FDM/FM channel to transmit the remaining 8 LDR + 2 MDR + TDRS Tracking + Order Wire + Telemetry data. Both channels have been designed with a 17.5-dB margin for operation in rain; however, the output power from the HDR channel can be reduced by 10-dB to conserve prime power during normal operation in clear weather. The HDR channel can also be turned off to further conserve prime power demands. The HDR channel employs dual mode TWT amplifier in its final power output stage, providing RF power outputs of approximately 10 and 1 watts to meet the 17.5-dB and 7.5-dB system margin, respectively. Since the established failure rates for TWT amplifiers are quite high ($\approx 6 \times 10^{-6}$), each TWT amplifier has been designed for increased reliability by using two parallel TWT's hard wired to a common high voltage power supply system (the appropriate heater will be energized to excite the desired TWT). In addition, the HDR channel has been designed with full block redundancy (2 TWT amplifiers such that quadruple redundancy exists for the TWT which constitutes the major failure component (Failure R $\approx 5 \times 10^{-6}$) in the TWT amplifier. The FDM/FM channel contains a solid-state K_u-band amplifier operating in the saturated mode driven by a solid-state VCO frequency modulated by the FDM signal.

Associated components of the TDRS Telecommunication System are the:

- TDRS Tracking and Order Wire Transponder
- TT&C Transponder
- Frequency Source

1.2 UPRATED DELTA: USER IMPACT

There have been a minimum number of design changes in the LDR and MDR User terminal, and the HDR terminal is a new design (i.e., HDR service not available in the base line system).

LDR User transponder characteristics for the Uprated Delta configuration are as follows:

- A 668 Kchip/sec PN chip rate in the forward link is to serve a two fold purpose, namely: 1) to provide sufficient spectrum spreading to meet the IRAC flux density requirements, and 2) to maximize the forward link processing gain (= PN chip Rate/Data Rate) by utilizing entire bandwidth allocation

- A short PN code is used to provide a maximum code acquisition time of approximately 40 seconds.
- A 1 Mchip/sec code rate in the return link is necessary to permit the multiple accessing of 20 users with sufficient process gain to meet the required level of performance.
- Unambiguous ranging is achieved by transmitting a coded word over the Command/Telemetry link whose duration is equivalent to the two way range uncertainty of 40,000 kilometers (approx. 133 m sec).

MDR User Terminal characteristics have changed only slightly from the base line configuration, namely:

1) A 5 Mchip/sec and 20 Mchip/sec PN chip rate for the MDR and Shuttle respectively in the forward link is necessary to distribute the signal energy to conform to the IRAC requirements.

2) The short code used during the code acquisition phase limits the maximum acquisition time to 40 sec. or less.

3) A coded word is switched in after code acquisition to provide unambiguous ranging over a two-way range uncertainty of 40,000 Km.

4) The PN modulation in the return link is a 5 Mchip/sec PN code which employed in both the spec. MDR user and the space shuttle.

The HDR user transponder design is functionally analogous to the spec MDR user transponder design. Doppler rates at K_u-band are approximately seven times that at S-band. The Doppler processor therefore must be designed to search over a greater range of frequency uncertainty.

The PN rate in the HDR forward link consists of a 50 Kchip-sec code of length 2047 in the normal mode. To meet the IRAC in the high power mode (i.e., 53.6 dBw to support forward link video) a chip rate of 60 Mchips/sec is required.

Range ambiguity resolution is achieved by transmitting a unique data word of 133 msec duration. This is equivalent in length to approximately 8000 PN chips on the forward link and 668,000 chips on the return link.

In addition to normal support of greater than 1 MBps, the return link may be required to support a 100 MBps HDR user. Performance of this link can be enhanced by approximately 5 dB with application of forward error control

1.3 UPRATED DELTA: GROUND STATION IMPACT

The TDRS ground station design for the Uprated Delta Launch configuration differs from the base line in the method of demodulation of the TDRS/GS signal. That is to say,

> The TDRS/GS link for the Uprated Delta Configuration employs FDM/FM for all the serious except the HDR return link which is in a separate channel.
>
> Correspondingly the Ground Station frequency selectively separates the HDR channel and all others are demodulated by the Phase Lock Loop FM Discriminator.

The other essential differences between the two ground stations is in the numberof MDR users the system can support simultaneously (four as opposed to two previously).

1.4 ATLAS CENTAUR/SPACE SHUTTLE CONFIGURATION: TDRS SPACECRAFT

The high performance Atlas Centaur/Space Shuttle TDRS Telecommunication System provides increased performance, service and support to meet the planned spaceborne user population as indicated in the NASA Mission Models. (1) This configuration uses:

(1) Notes on Preliminary 1978, 1979 and 1980 Mission Model by NASA-GSFC.

- Sr. AGIPA in the LDR return link to improve its performance in high interference signal environment.

- 4 - MDR/HDR dual frequency transponders to service and support the planned MDR and HDR user population.

In the LDR return link, Sr. AGIPA uses a 5 element (short backfire design) ring array with a diameter of 5λ to provide an effective half power beamwidth (HPBW) of approximately 8 degrees as compared to 24 degrees for Jr. AGIPA. RFI model analysis conducted in Part I has shown that Sr. AGIPA can more effectively provide interference discrimination, with improvements in signal-to-interference ratio (Δ SIR) of 9 to 15-dB as compared to a F-FOV approach. This Δ SIR provides the necessary improvement to meet the required data rates even in the presence of a high RFI environment.

It is to be noted that since AGIPA (both Jr. and Sr.) is an adaptive signal processing system that continuously computes and adjusts the amplitude and phase weighting factors (at the ground station), the system is immune to the normal structural flexure and antenna pointing variations experienced with conventional phased array. Consequently, considerable flexibility can be exercised in the transponder design and spacecraft implementation of the AGIPA systems.

The LDR return link can also be used in the F-FOV backup mode by using any two orthogonally polarized antenna channels.

In the LDR forward link, 5 identical transmit channels feeding a high gain disc-on-rod antenna element are used. Only one channel is required in the prime F-FOV mode as in the Uprated Delta Configuration providing an EIRP of +30, +27 or +24 dBW. However, in the Emergency Steered Beam Mode, all 5 elements are used as a phased array to provide an EIRP of +38, +41, or +44 dBW. Consequently, in this Steered Beam Mode the EIRP has been increased by +1 dB over the Uprated Delta design, due to the 1-dB increase in array gain.

In the 4 MDR/HDR Transponders, each transponder design and performance is identical to that described for the Uprated Delta Configuration. However, the 2 additional MDR/HDR channels can provide multiple user access to 4 MDR users and to 1 HDR user simultaneously. Since all 4 MDR/HDR Transponders are designed identically, quadruple functional redundancy exist for the support of MDR as well as HDR user. In addition, 2 of the MDR/HDR channels are each designed with a capability to backup the TDRS/GS Transponder, thereby providing triple redundancy to the TDRS/GS Transponder.

The TDRS/GS Transponder is an all FDM system in both return and forward direction. In the return link, two channels are used; viz. an HDR channel and an FDM channel (which combines the 10 LDR + 4 MDR + TDRS Tracking + Order Wire + Telemetry Data) - this channel replaces the FDM/FM channel used in the Uprated Delta. The HDR channel is 150-MHz wide to support the 100 Mbps return data and uses a solid-state K_u-band amplifier operating in a saturated mode. It is to be noted that since a 6-dB higher gain antenna is used in this link as compared to the Uprated Delta Configuration, a TWT amplifier is no longer required, thereby reducing the size and weight and increasing the overall system reliability. The FDM channel occupies a spectrum of 84-MHz and uses a solid-state K_u-band amplifier operating in the linear mode with approximately 10-dB backoff from saturation. Both channels are designed to operate at all times with a system margin of 17.5-dB.

The TDRS/GS forward link data is an FDM signal, occupying a spectrum of approximately 380-MHz. Each MDR/HDR channel is designed broadband (75-MHz wide) such that no tuning is required at the TDRS relay spacecraft. All tuning (frequency selection) is determined at the ground station, as in the Part I and Uprated Delta Configuration.

The TDRS/GS 3.8-meter K_u-band antenna provides a HPBW of approximately 0.4 degree; therefore, closed-loop tracking using a 4 horn pseudo-monopulse feed is employed on the TDRS relay spacecraft.

The associated components of the Alternate Atlas Centaur/Space Shuttle Telecommunication System are:

- TDRS Tracking and Order Wire Transponder
- TT&C Transponder
- Frequency Source

1.5 ATLAS CENTAUR/SPACE SHUTTLE: USER IMPACT

The impact of the Atlas Centaur Shuttle configuration on the user spacecraft transponder (regardless of the type of service) is virtually nil. With the exception of increased forward link EIRP in the LDRU emergency mode, the forward link support on a per user basis is unchanged.

1.6 ATLAS CENTAUR/SPACE SHUTTLE: GROUND STATION IMPACT

The impact of this maximum capability service on the TDRS ground station is in two areas. First the PLL/FM discriminator used in the Uprated Delta system has been eliminated. The downconverted signal (at S-band) is applied directly to an FDM demultiplexer, whose output is channeled to the demodulation/tracking units for each of the users. In the MDR/HDR telemetry channels the ground station can support simultaneously in the telemetry link either four MDR users or three MDR plus one HDR users.

In the forward link the various user command channels are combined in FDM in a manner similar to that in the ground station designed to support the up-rated Delta system. The maximum capability system up links nine channels in FDM. There are four MDR channels and one each for HDR, LDR, order wire, TT&C, and the pilot

1.7 S-BAND MULTIPLE ACCESS AGIPA ARRAY

In addition to the previously described Uprated Telecommunication Systems for implementation on the Delta 2914, an alternate approach employing an S-band AGIPA array similar to the VHF LDR AGIPA approach has been examined to support LDR users with a return data rate up to 20 kbps. The data included herein is brief, and is limited to comments on the modulation format and to the hardware implementation impact on the Delta payload, based on design requirements provided by NASA-GSFC TDRS Program Office. This data is included in Appendix E.

2.0 DESIGN REQUIREMENTS

The minimum service support requirements as revised by NASA-GSFC(1) for the Part II phase are summarized in Table 2-1. These requirements are used to establish the minimum performances for the TDRS Telecommunication System when launched by the Delta 2914, the Atlas/Centaur, and the Shuttle/Agena vehicles.

1. NASA and North American Rockwell meeting on 16 November 1972 at GSFC.

TABLE 2-1. SUMMARY OF MINIMUM SERVICE SUPPORT REQUIREMENT

A. LDR Services

1. Return Link (AGIPA VHF - Multiple Access)

- o 20 LDR users spacecraft simultaneously
- o Data rate User 1 to 10 kbps @ 10^{-5} BER

2. Forward Link (UHF)

- o Command rate 100 to 1000 bps @ 10^{-5} BER
- o One command link spacecraft time shared with all LDR users

The specifications for the command link (as given in the SOW) are as follows:

-30 dbw @ 26° beamwidth
-28 dbw @ 30° beamwidth

This specification can be met or exceeded by using a fixed FOV or beam steering approach. The fixed FOV approach is by far the preferred approach. A high power mode for the fixed FOV approach will be considered as a method to meet this specification.

- o Acq. (PN lock-up) - 20 LDR users simultaneously (Fixed FOV)

A fixed FOV approach must be used to satisfy this requirement. The EIRP shall be established by the contractor to insure a user acquisition time of 60 seconds or less. If a phased array (beam steering) approach is used for the command link then one of the elements must be used as a fixed FOV beam for acquisition.

- o No LDR Forward Voice (can be considered as an option - if possible only after the minimum service support model is satisfied. Forward link voice (if treated) should be treated as an extreme emergency capability - i.e. part or all of the TDRS service capability can be shut down in order to provide sufficient power for UHF voice).
- o Fixed FOV and beam steering approach can be considered - from operational standpoint. Fixed FOV approach is preferred

B. High Gain Antenna Service Capability

1. General

- o Two 3.8 meter parabolic reflector antenna
- o Dual Feed, S and X or Ku-band

S-band to service MDR user. X or Ku band to service HDR users (selection and rationale for X or Ku-band will be provided by contractor).

2. Return Link

- o Must support two users simultaneously from each TDRS S/C i.e.

a) 2 MDR users simultaneously or
b) 1 MDR and 1 HDR user simultaneously

(It is not required to support 2 HDR users simultaneously. See Note 1)

- o MDR user - up to 1 Mbps @ 10^{-5} BER
- o Shuttle (S-band only)

Data -128 kbps
plus
2 voice - 32 kbps voice channel
(Total Data Rate 192 kbps @ 10^{-4} BER - TDM

- o HDR User - 100 Mbps @ 10^{-5} BER

3. Forward Link

Must support two users simultaneously

a) 2 MDR user or
b) 1 MDR and 1 HDR

Forward Link Command Data Rates

a) Unmanned Users - 1 kbps @ 10^{-5} BER for both MDR and HDR users
b) Shuttle (S-band only)

Data 6 kbps (2 kbps info - rate 1/3 encoding)
plus
2 voice channels -24 kbps voice channel (total data rate - 54 kbps @ 10^{-3} BER - TDM
push to talk voice - 25% duty cycle - duty cycle defined on short time basis
(1 hour per channel)

C. Other Services

1. 1 beacon to assist the MDR and HDR user acquisition(3)

- o Frequencies compatible with MDR User
- o Frequencies compatible with HDR User

2. 1 S-band order wire antenna

- o Continuous operation
- o 30° BW minimum
- o Frequency compatible with user

3. 1 S-band transponder for trilateration tracking

- o Intermittant Operation (on the order of 1 or 2 hours/day)
- o Frequencies compatible with STDN

NOTES

1. The objective is to provide a 17.5 db rain margin in the TDRS ground links on a continuous basis to support the minimum service support capability outlined above. If this objective cannot be met then alternative operational modes should be considered to obtain this objective. If these alternatives prove ineffective then the 17.5 db rain margin can be reduced for the HDR signal only, for continuous operation during fair weather. The capability to increase to 17.5 db rain margin for HDR during rain will be provided. Contractor will define the maximum continuous operating time for this mode.

2. It is not required to support the HDR user during eclipse.

3. Beacon requirement has been removed. All MDR and HDR users will with S-band Command Receiver which can be used to access the desired user.

3.0 DELTA 2914 CONFIGURATION

The overall physical layout of the Part II alternate Uprated Delta 2914 Configuration as deployed in space is shown in Figure 3-1. The layout is very similar to the Part I Baseline design with the exception that larger (3.8 meter diameter) deployable parabolic reflectors have been used for the two MDR/HDR space-to-space links and a 1.8 meter fixed parabolic reflector (relocated onto one of the solar panel support booms) has been incorporated for the space-to-ground links. These changes and a NASA requirement to support only one forward LDR link has made it possible to improve Part I link performance capabilities to better service and support the MDR and HDR users, both unmanned and manned.

This design meets the requirement to service and support a minimum of 20 LDR + 1 MDR + 1 HDR users from one TDRS satellite.

3.1 SYSTEM DESCRIPTION

The overall functional block diagram of the Telecommunication System for the Part II - Uprated Delta 2914 Configuration is shown in Figure 3-2 which shows the major components of the system, viz:

- LDR transponder
- MDR/HDR #1 transponder
- MDR/HDR #2 transponder
- TDRS/GS transponder
- TT&C transponder
- TDRS tracking and order wire transponder
- Frequency source

Since this Part II design is functionally similar to the Part I base line design, a comparison of the transponder characteristics for Part II system versus the Part I base-line design is summarized in Table 3-1.

This Section 3 provides a description of the overall modes of service, frequency plan, link performance summary, detailed system block diagram, weight, and power summary; and a detailed description of the major components of Telecommunication System for the Part II: Uprated Delta 2914 Configuration.

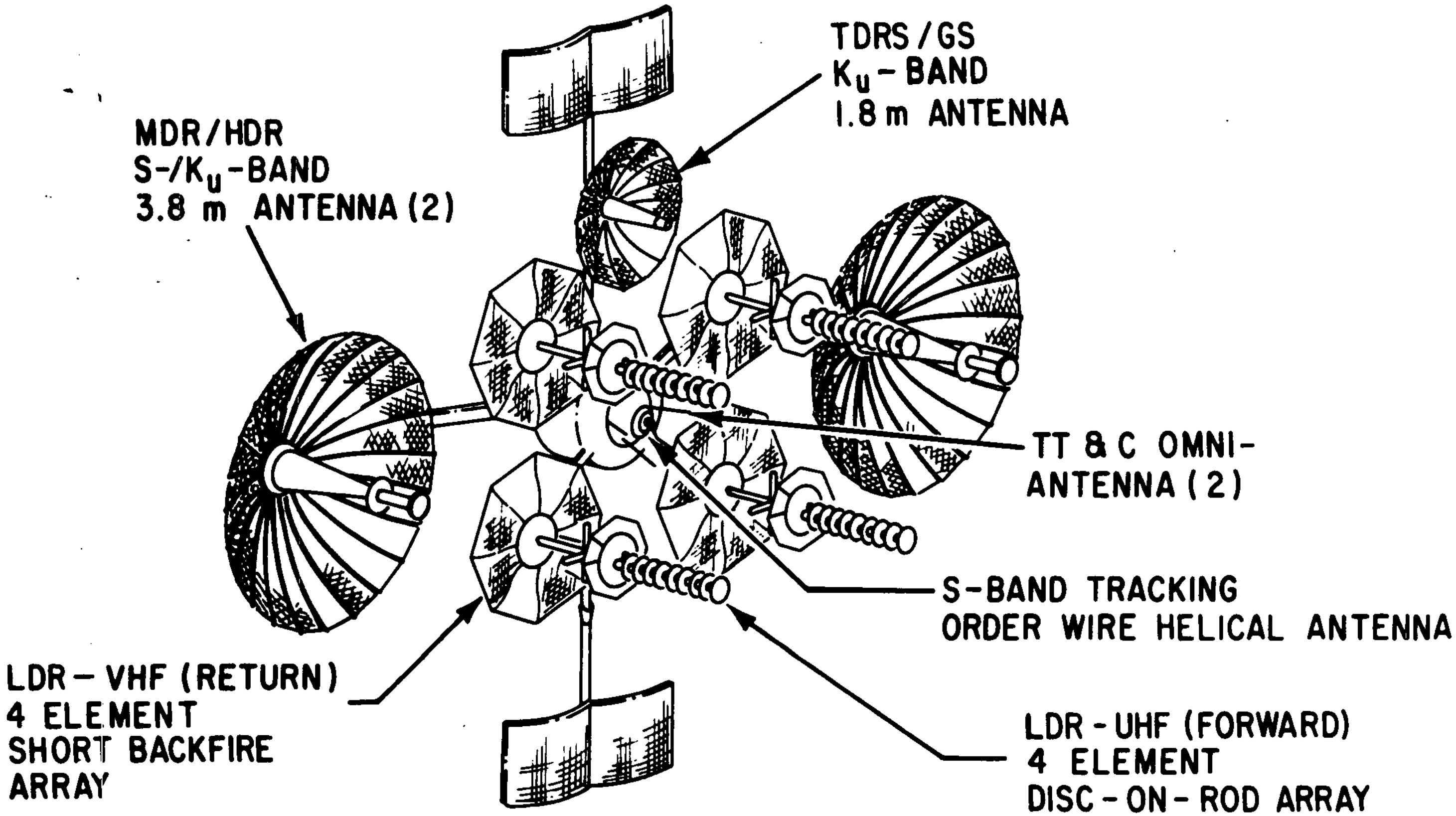
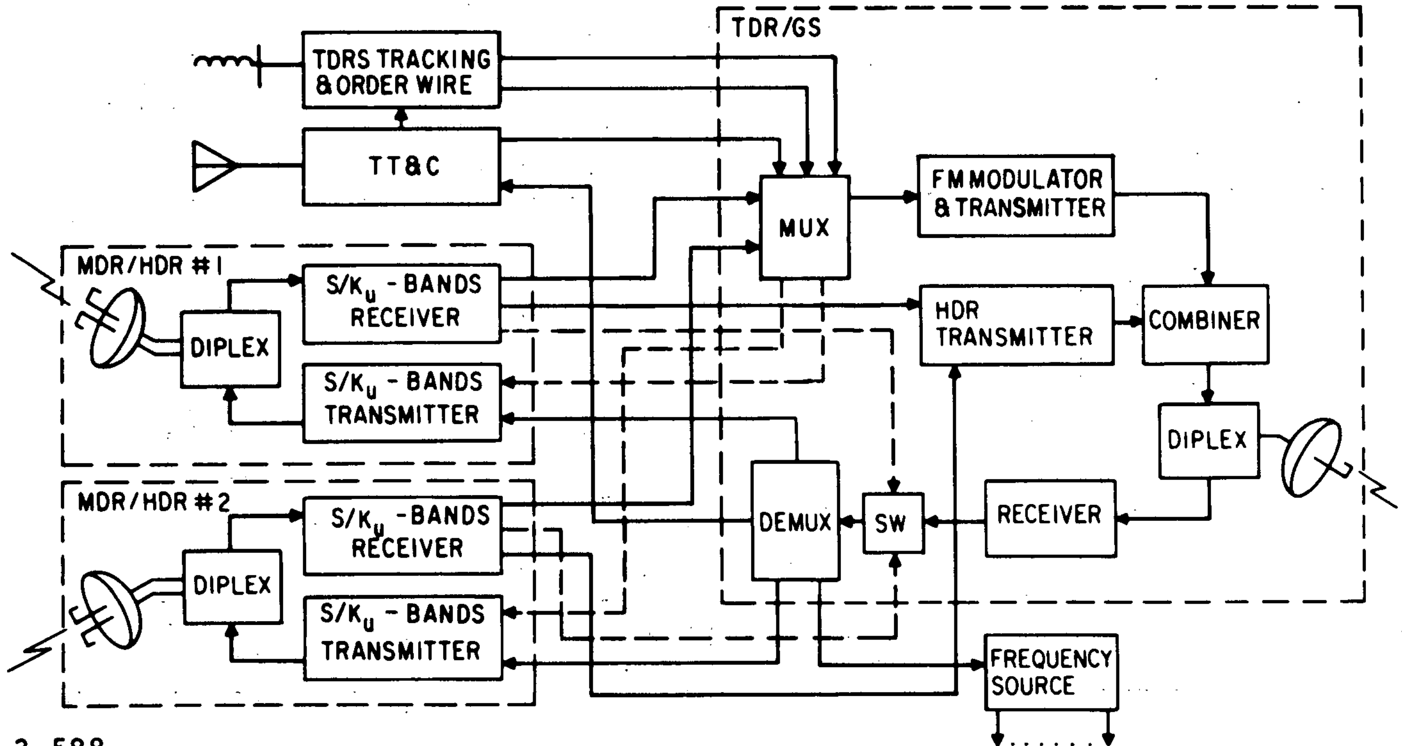


FIGURE 3-1. ANTENNA FARM CONFIGURATION: UPRATED DELTA 2914



3-588

FIGURE 3-2. TDRS TELECOMMUNICATION SYSTEM BLOCK DIAGRAM: UPRATED DELTA 2914

TABLE 3-1. TDRS TELECOMMUNCATION SYSTEM DELTA CONFIGURATION: PART I VS PART II MODIFICATIONS

Component	Part I	Part II
LDR Transponder • Forward	• 4-1 MHz channels to support 2 forward links (Chip rate = 167 kcps/channel): - prime mode: 2 satellite steered beams at +30 dBw at 31° FOV - backup mode: 1 F-FOV at +24 dBw at 31° FOV	• 4-1 MHz channels to support 1 forward link (Chip rate = 668 kcps/channel): - prime mode F-FOV with +30, +27 or +24 dBw at 26° FOV - emergency mode: satellite steered beam with +36, +39, or +42 dBw at 26° FOV
• Return	• 8-2 MHz channels Jr. AGIPA receiver to simultaneously support 20 users; G_{array} = 16.8 dBi	• Same except G_{array} = 17.4 dBi
MDR/HDR Transponder • Forward	• dual frequency: S or Ku • RF bandwidth: -S-band: 10 MHz tunable at GS over 95 MHz -Ku-band: 4-100 MHz band tunable at TDRS • EIRP: (Duty cycle) - S-band: +41(100%) or 47(25%) dBw - Ku-band: +45.6(100%) dBw	• Same • RF bandwidth: -S-band: 10 or 32 MHz tunable at GS over 75 MHz - Ku-band: fixed 100 MHz • EIRP: - S-band: +41(100%) or 47(100%) dBw - Ku-band: +23.6(100%) or +53.6 (≤10%) dBw
• Return	• dual frequency: S or Ku-band • RF bandwidth: - S-band: 10 MHz tunable in 5 MHz step over 100 MHz band -Ku-band: 4-100 MHz band tunable at TDRS • G/T_s : -S-band: +3.9 dB/°K -Ku-band: +20.4 dB/°K	• dual frequency: S and Ku-band • RF bandwidth: -S-band: Same -Ku-band: fixed 150 MHz • G/T_s: -S-band: +10.0 dB/°K -Ku-band: +25.9 dB/°K
• Backup	• Mode: FDM • Frequency: Ku or S-band • EIRP: -S-band: +47 dBw -Ku-band: +45.6 dBw	• Same • Same • EIRP: -S-band: +47 dBw -Ku-band: +53.6 dBw
TDRS/GS Transponder • Forward	• RF bandwidth: 240 MHz • G/T_s = +6.0 dB/°K	• Same • G/T_s = +12.8 dB/°K
• Return	• Mode: FDM/FM (600 MHz) • EIRP: - High power = +44.6 dBw - Low Power = +34.6 dBw	• Mode (dual channel): -HDR channel (150 MHz) -FDM/FM channel (200 MHz) • EIRP: - High power = +56.7 dBw - Low power = +49.0 dBw - FDM/FM channel only = +41.7 dBw

3.1.1 MODES OF SERVICE

The Telecommunication System has been designed with considerable flexibility and versatility to optimally service and support the mixture of LDR, MDR, and HDR users during the operational lifetime of the spaceborne relay platform. The flexibility has been incorporated to optimally meet the highly variable environment and user needs anticipated for the Telecommunication System, viz:

- Daylight and eclipse operation
- Decaying solar panel and battery output with life
- Number of LDR, MDR, and HDR users anticipated during the planned 5 year lifetime
- Versatility to improve link capacity for short time periods, either to meet emergency command requirements under high RFI environment or to send video information to manned user(s)
- To provide the required system margin of 17.5 dB under adverse rain condition, but include the flexibility to use this prime power to improve other link performance or user support capabilities when the rain margin is not required.

To meet these needs, each of the TDRS transponders(1) has been designed with several modes of operation which is summarized in Table 3-2. These various modes can be used in any combination, limited only by the available prime power at the time of usage. The available prime power is a function of daylight versus eclipse, life, and user support and is described in depth in the subsequent Operations and Capabilities Section.

The primary mode of operation for the LDR UHF forward link is the fixed-field of view (F-FOV) mode which generates an EIRP of +30 dBW under normal daylight operation. In this mode all 20 LDR users are simultaneously illuminated to provide a signal for coherently locking their local PN and frequency references. To conserve primary power during eclipse, each LDR transmitter is designed to provide a reduced output of +27 and +24 dBW. The LDR transmitter is designed with quadruple redundancy in the

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(1) Transponder is used herein to include the antenna, receiver, and transmitter, whereas transceiver includes only the receiver and transmitter.

TABLE 3-2. TDRS TELECOMMUNICATION: MODES OF SERVICE: UPRATED DELTA

Modes	Transmit EIRP at 26° FOV	Receive G/T_S at 26° FOV
LDR Link		
• Primary Mode	• F-FOV = +30/+27/+24 dBW (command 1 at a time)	• Jr. AGIPA = -13.7 dB/°K (support 20 simultaneously)
• Backup Mode	• Steered Beam = +42/+39/+36 dBW	• F-FOV = -16.9 dB/°K
MDR/HDR #1 & #2		
• Primary Mode	• Support 2 MDR or 1 MDR + 1 HDR	• Support 2 MDR or 1 MDR + 1 HDR
	S-band: Unmanned = -41 dBW Manned = +47 dBW	} +10.0 dB/°K
	K_u-band: Unmanned = +23.6 dBW Manned = +53.6 dBW	} +25.9 dB/°K
• Backup Mode	• K_u-band = +53.6 dBW	+25.9 dB/°K
	S-band = +47.0 dBW	+10.0 dB/°K
TDRS/GS LINK		
• Primary Mode	• FDM/FM/FDM	
	High Power Mode = +54.7 dBW Low Power Mode = 45.7 dBW	+12.8 dB/°K +12.8 dB/°K
	• FDM/FM only = +39.0 dBW	
• Backup Mode		
K_u-band	+53.6 dBW	+25.9 dB/°K
S-band	+47.0 dBW	+10.0 dB/°K
VHF	7.7 dBW	-30.2 dB/°K

F-FOV mode, which can also be used as a 4-element phased array to provide an EIRP of +36, +39, or +42 dBW. This additional EIRP of up to +12 dBW provides a critical need to send command to a user in an emergency condition, or to an LDR user who is exposed to a high RFI environment.

In the LDR VHF return link, the primary mode is Jr. AGIPA (Adaptive Ground Implemented Phased Array) which adaptively optimizes the desired signal-to-interference ratio (SIR) by employing spatial and polarization discrimination of the undesired interference signal including intentional interferences from other in-band LDR users. In this primary mode, the G/T_s (ratio of antenna gain-to-system noise temperature) is -13.7 dB/^{o}K. The return link can also be operated in a F-FOV backup mode by either using the quad-array in a fixed beam mode (antenna beam boresighted along the spacecraft vertical) or by using only one antenna element. The G/T_s is 16.9 dB/^{o}K and -19.4 dB/^{o}K for the fixed array and single element cases, respectively.

MDR/HDR #1 and #2 transponders are designed identically, and operate at either S- or K_u-band to support MDR and HDR users, respectively. Both transponders can be operated in the S-band mode, or in the K_u-band mode; or one S-band and one K_u-band. In the S-band mode, an EIRP of +41 dBW is generated to support unmanned user requirements, and an EIRP of +47 dBW to meet the manned space shuttle requirements. The return link G/T_s is +10 dB/^{o}K. In the K_u-band mode, the transmitter has also been designed with the flexibility to provide an EIRP of +23.6 dBW to send 1 kbps to unmanned users, and +53.6 dBW to send video information to manned users. The receiver K_u-band G/T_s is +25.9 dB/^{o}K.

Both MDR/HDR transponders are also designed to provide a functional backup mode to the TDRS/GS transponder at either K_u-band or S-band. The data is transmitted to the ground station (GS) in an FDM format at a EIRP of +53.6 dBW or +47.0 dBW at K_u-band and S-band, respectively. An additional backup capability has also been included on MDR/HDR #1 transponder such that its 3.8 meter antenna can be used in lieu of the 1.8 meter TDRS/GS antenna to provide an additional 6.5 dB for a total of 24 dB of system margin for operation in heavy rainfall.

The TDRS/GS transponder is a channelized system; the 100 Mbps HDR data is sent on one channel and all other data (8 channel LDR, 2 MDR, TDRS telemetry, TDRS tracking, and Order Wire data) are FDM and frequency modulated onto a separate channel, referred to herein as the FDM/FM channel. The combined HDR and FDM/FM channels are referred to as the FDM/FM/FDM system. The total FDM/FM/FDM system is designed to operate with a EIRP of +54.7 dBW to provide the required 17.5 dB rain margin, but the HDR channel can also be reduced to operate with a system margin of +7.5 dB for operation under normal clear weather. This will allow the excess prime power to be used to enhance other link performances or support capabilities when the 17.5 dB rain margin is not required. Furthermore, the HDR channel can be turned off completely such that only the FDM/FM channel data is returned to the GS. This FDM/FM channel always operates with a rain margin of +17.5 dB and requires a EIRP of +39.0 dBW. In the GS/TDRS forward link, the receiver operates with a G/T_S of +12.8 dB/^oK.

3.1.2 FREQUENCY PLAN

The overall TDRS telecommunication system Frequency Plan is shown in Figure 3-3 for each of the space-to-space and space-to-ground links.

The LDR space-to-space link operates in the UHF band (400.5 to 401.5 MHz) in the forward link and in the VHF band (136 to 138 MHz) in the return link.

The MDR space-to-space link opeates at S-band. In the forward link a 10 MHz (low power mode: EIRP = +41 dBW) or a 32 MHz (high power mode: EIRP = +47 dBW) wideband transmitted. This band is unable at the ground station to operate anywhere within the 75 MHz total transponder bandwidth between 2.025 to 2.100 GHz. The onboard MDR transponder is designed with an RF bandwidth of 75 MHz such that no tuning is required on the relay platform. In the return link, receiver front end is designed essentially, wide open with an RF bandwidth of 100 MHz between 2.2 to 2.3 GHz in order to provide service and support to all potential MDR users, including Department of Commerce satellites at the lower spectral range and JPL Deep Space Satellites at the higher spectral range. Since this band must be shared with the TDRS tracking transmitter and the order wire receiver, these functions were located at 2.285 GHz and 2.218 GHz, respectively, in order to minimize its impact to support the currently assigned Commerce and Deep Space Frequency allocation bands while maintaining support to other MDR users. The actual operating frequency can be selected by the user and/or ground control center anywhere within this 100 MHz bandwidth with the exception of approximately ±5 MHz of the TDRS tracking transmitter and order wire frequencies. Frequency tuning in 20-5 MHz steps is provided onboard the relay satellite. The TDRS tracking receiver is located at 2.12 GHz.

The HDR space-to-space link shares the allocated K_u-band with the space-to-ground link. The HDR forward link occupies 100 MHz in the 14.81 to 14.91 GHz band while the HDR return link occupies 150 MHz in the 13.85 to 14.0 GHz band. The GS/TDRS ground-to-space link uses 240 MHz in the 13.4 to 13.64 GHz band. The TDRS/GS space-to-ground link is an

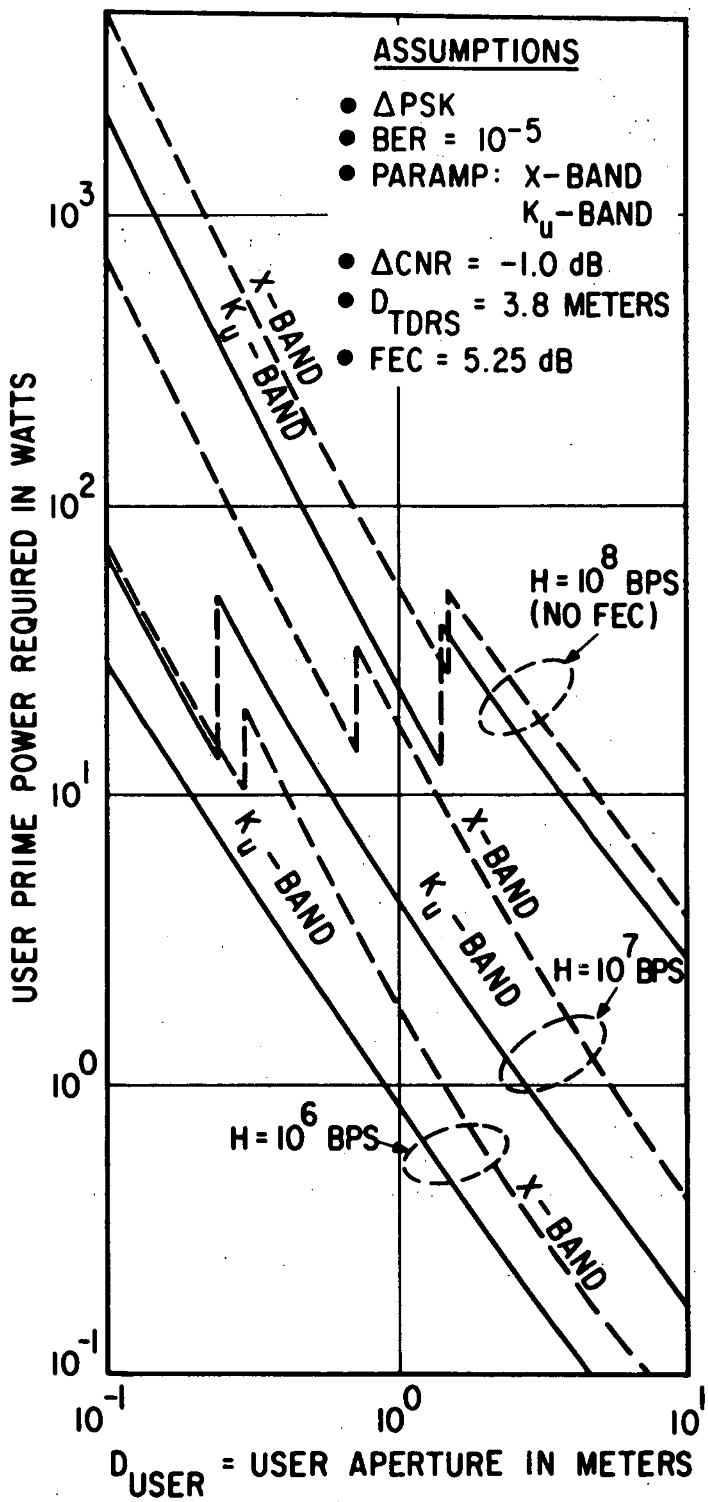

FIGURE 3-3. TDRS TELECOMMUNICATION SYSTEM FREQUENCY PLAN

FDM/FM/FDM channelized system with the HDR channel requiring 150 MHz in the 14.6 to 15.75 GHz range and the FDM/FM channel requiring 200 MHz in the 15.0 to 15.2 GHz range.

Contrary to the Part I base line approach there is no spectral overlap between the space-to-space and space-to-ground K_u-band links. However, both TDRS #1 and #2 satellites operates in the same frequency band. In the space-to-ground link, ground antenna designs with ultra-high front-to-side lobe levels (>60 dB) and good ground site preparation as described in detail in the Part I Final Report can be used to eliminate any cross interferences between the down links. In the MDR space-to-space links unique operating frequency within the allocated frequency bands must be selected to eliminate cross interferences. In the HDR space-to-space link, only one HDR user support is required by the two TDRS satellites; therefore, no cross interference exist. In the LDR space-to-space link, a unique frequency is used at each TDRS satellite in the forward link; and in the return link, a unique PN code is assigned to each of 40 users. Of these 40 users, each TDRS satellite can support a maximum of 20 users.

The Tracking Telemetry and Command (TT&C) Subsystem operates in the VHF band (136 to 138 MHz) in the return telemetry direction and in the 148 to 150 MHz in the forward command direction. This VHF TT&C Subsystem is the prime communication link during the inflight transfer phase but becomes a secondary backup system on-station at geosynchronous altitude. Consequently, the overlap in the 136 to 138 MHz band with the LDR space-to-space return link is not expected to cause interferences under normal operation.

3.1.3 BANDWIDTH SPREADING REQUIREMENT

The Interdepartment Radio Advisory Committee (IRAC) of the Office of Telecommunications Policy has established guidelines for the maximum power flux density that can illuminate the earth's surface. These guidelines in conjunction with the required EIRP for each of the space-to-space and space-to-ground links are used to establish the minimum bandwidth spreading that is required to stay within the IRAC guidelines.

The carrier flux density at the earth's surface from an emitter with antenna gain G_T can be approximated by

$$\text{Flux density} = \frac{P_T \cdot G_T}{4\pi R^2} \text{ watts/m}^2 \tag{3-1}$$

where:

P_T = transmit power in watts

G_T = antenna gain

R = distance from the emitter to earth in meters

To conform with the IRAC measurement, which is made in a 4 kHz bandwidth, and to take into account any spectrum spreading (BW) of the signal, the following relationship is used:

$$\text{Flux density } (\text{dBw/m}^2/4\text{ kHz}) = \frac{G_T \cdot P_T \cdot 4 \times 10^3 \text{ Hz}}{(4\pi R^2)\text{ BW}}$$

With the TDRS at synchronous altitude the expression, when solved for spread bandwidth required to meet the flux density specification is,

$$\text{Spread bandwidth (BW)} = \frac{\text{EIRP} \cdot 4 \times 10^3}{4\pi R^2 \cdot [\text{Flux Density}]}$$

or in dB notation

$$BW_{dB} = \text{EIRP} + 36 - 163 - \text{Flux Density}$$

Table 3-3 summarizes the EIRP, IRAC requirements, the minimum spread bandwidth, and the RF bandwidth used in the design of the TDRS transponders. All links meet the required minimum spreading bandwidth requirement with the exception of the LDR Emergency Steered Beam Mode when an EIRP of +39 and +42 dBW is emitted at UHF. Since this mode is to be used primarily for emergency burst (short duration) condition to overcome high RFI, and since no firm IRAC guideline has been established in the UHF band, this exception was considered acceptable.

TABLE 3-3. BANDWIDTH SPREADING REQUIRED TO MEET IRAC SPECIFICATIONS

MODE	EIRP (dBw)	IRAC SPEC (dBw/m^2/4 kHz	BANDWIDTH (MHz)	
			MIN SPREADING REQ'D	TRANSPONDER DESIGN
LDR				
• F-FOV	+30	−150	• ≥ 250 kHz	1 MHz
• EMERGENCY STEERED BEAM	+36/+39/+42		• ≥ 1/2/4 MHz	
MDR				
• UNMANNED	+41	−154	• ≥ 8 MHz	75 MHz
• MANNED	+47		• ≥ 32 MHz	
HDR				
• UNMANNED	+23.6	−152	• ≥ 91 kHz	100 MHz
• MANNED	+53.6		• ≥ 91 MHz	
TDRS/GS (HDR CHANNEL)				
• HIGH POWER	+56.6	−152	• ≥ 137 MHz	150 MHz
• LOW POWER	+47.1		• ≥ 15 MHz	

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3.1.4 LINK PERFORMANCE SUMMARY

In order to compute the link performance in a tandem link, the following equation is used to compute the carrier-to-noise ratio (CNR) required in each leg of the overall tandem link:

$$CNR_0 = \frac{CNR_1 \times CNR_2}{CNR_1 + CNR_2 + 1}$$

where:

CNR_0 = effective output CNR required at the ground or user terminal

CNR_1 = CNR required in the first leg of the tandem link

CNR_2 = CNR required in the second leg of the tandem link

The above equation can be rewritten as:

$$\Delta CNR = \frac{CNR_2}{CNR_1 + CNR_2 + 1} \qquad (3\text{-}2)$$

where ΔCNR is the allowable CNR degradation or loss ($\Delta CNR = CNR_0 / CNR_1$) through the TDRS relay transponder. This expression is plotted in Figure 3-4 for several values of ΔCNR.

In the return link a ΔCNR of -1 dB has been used in order to minimize the impact on the user; and in the forward link, a ΔCNR of -0.25 dB is used since the Ground Station has the greater flexibility and capacity to increase its effective instantaneous radiated power (EIRP) to minimize the requirements on the space-to-space link.

The link power budget and performances has been computed for each of the space-to-space and space-to-ground links.

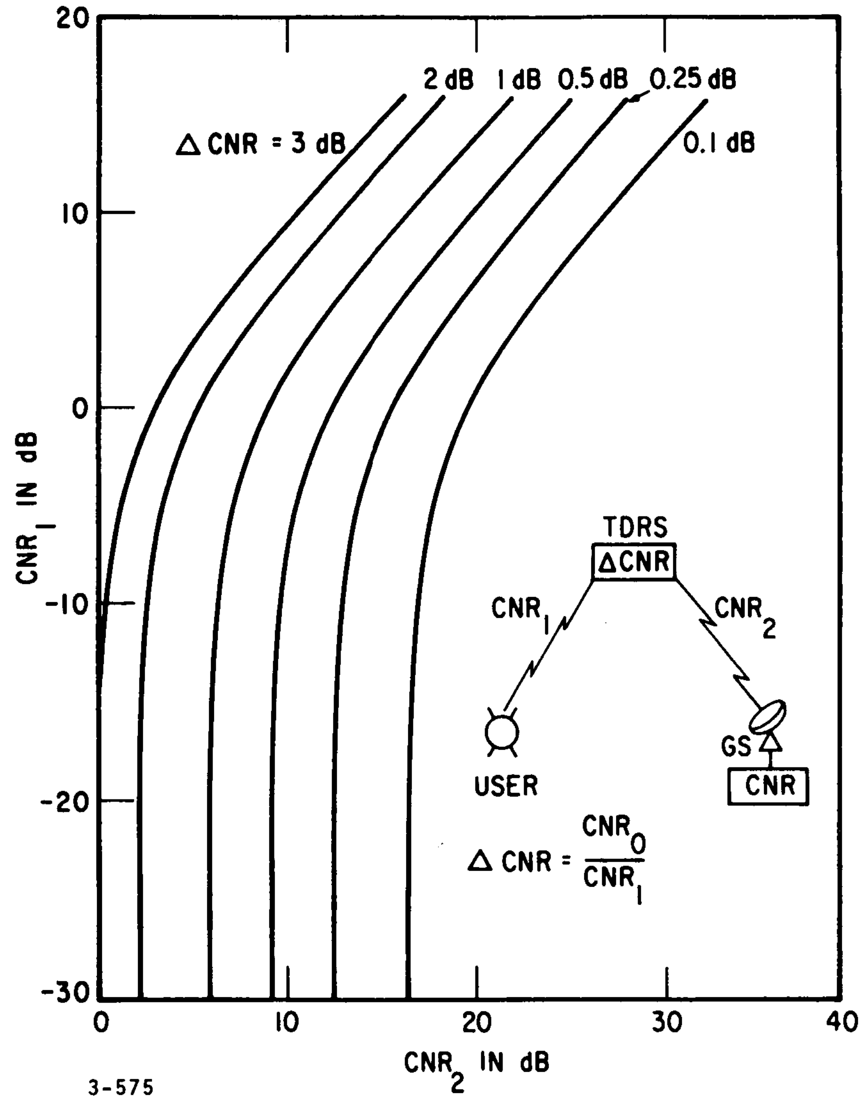

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FIGURE 3-4. CNR REQUIREMENT IN TANDEM LINK AS A FUNCTION OF CNR DEGRADATION THROUGH TDRS

3.1.4.1 LDR LINK PERFORMANCE

The LDR Forward Link operates in the 400.5 to 401.5 MHz UHF band, and provides a forward channel for data. In the prime mode the link operates in the F-FOV mode, but also has an Emergency Steered Beam Mode where higher EIRP is generated for operation when the user is engulfed in the midst of high RFI environment.

Sample link calculations have been made,as shown in Table 3-4 for both F-FOV and Emergency Steered Beam Modes, using the following expression for command data rate (H):

$$H = \frac{P_u}{E_b/N_o \left\{ N_i + RFI_o + \frac{1}{CR} \left[\frac{4 P_u}{\alpha} \right] \right\}} \tag{3-3}$$

where:

P_u = received signal power at user

E_b/N_o signal-to-noise power density per bit

N_i = input noise power density

RFI_o = effective RFI power density at receiver input

CR = chip rate

Multipath = $4 P_u/\alpha$ (worst case)

$\alpha = 2\,h/300$

h = user altitude (300 to 5000 km)

In the calculation, the worst case combination of the space loss and scan loss has been used, since the individual worst cases do not occur simultaneously. The link calculation shows typically that in the presence of -160 dBm/Hz RFI environment that peak command data of 176 bps and up to 2780 bps can be supported in the primary and backup modes, respectively. This expression has been plotted as a function of RFI power density in Figure 3-5 for the

TABLE 3-4. LDR FORWARD LINK BUDGET

	F-FOV Mode	Emergency Steered Beam Mode
Modulation	ΔPSK	ΔPSK
EIRP at 26^{o} FOV, dBw	+ 30.0	+36/ +39 / + 42
Losses-space + scan(1), dB	- 178.1	- 178.1
- polarization, dB	- 3.0	- 3.0
User Antenna Gain, dBi	- 3.0	- 3.0
Received Power, dBw	- 154.1	-148.1 / - 145.1 / - 142.1
System Noise Temp. (2), dB	26.8	26.8
Noise Density, dBw/Hz		
• Thermal	- 201.8	- 201.8
• Multipath (3)	-209.3	-203.3/-200.3/-197.3
• RFI. (4)	- 190.0	- 190.0
Total Noise Density, dBw/Hz	- 189.7	-189.5/-189.4/-189.0
TDRS ΔCNR Loss(5), dB	- 0.25	- 0.25
Available C/N_o, dB-Hz	35.35	41.15 / 44.05/ 46.65
E_b/N_o (6), dB	9.9	9.9
System Margin, dB	3.0	3.0
Achievable Bit Rate, dB	22.45	28.25 / 31.15 / 33.7
bps	176	660 /1300 / 2340

NOTES:

1. Worst case combination of the space loss and scan loss since they do not maximize at the same spacecraft aspect angle. Note also that EIRP at 26° FOV includes 2 dB scan loss.
2. Assume transistor front end (Te ≈ 100 K) at user.
3. CR = (Chip Rate) = 668 kcps
4. RFI_o of -160 dBm/Hz assumed for link budget calculation.
5. A CNR degradation (ΔCNR) of -0.35 dB assumed for the TDRS relay transponder.
6. E_b/N_o = 9.9 for ΔPSK at BER = 10^{-5}.

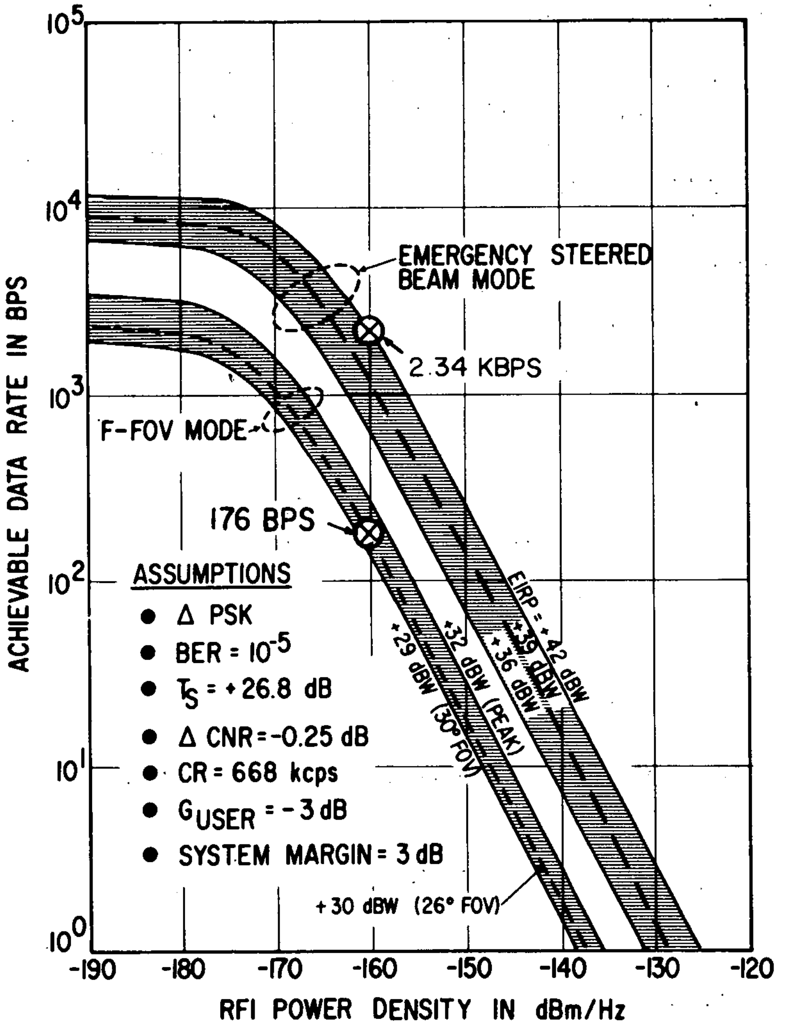

FIGURE 3-5. LDR FORWARD LINK: DATA RATE VS POWER DENSITY (PART II - DELTA 2914 CONFIGURATION)

primary F-FOV mode at the peak of the beam, 26° FOV and 30° FOV points. The assumptions used in the plot are shown in Figure 3-5 where E_b/N_o for ΔPSK (delta PSK) at a bit error rate probability (BER) of 10^{-5} is 9.9 dB. The figure also shows the data rate performance for the Emergency Steered Beam Mode with an EIRP of +36, +39, and +42 dBW. Most significantly the emergency mode provides a capability to send emergency commands to a user who could be engulfed in the midst of high RFI environment (+12 dB greater than F-FOV Mode). The +36 dBW can be sent continuously but the +39 dBW and +42 dBW is limited to short duration burst (e.g. 10 minutes per day) since it uses battery power to meet its peak power requirement.

The LDR return link operates in the 136 to 138 MHz VHF band, and uses an adaptive process called Jr. AGIPA in its primary mode, and F-FOV as a backup mode. The link performance was calculated as shown in Table 3-5 using the following expression to compute the achievable data rate (H):

$$H = \frac{P_u\ (FEC)}{E_b/N_o \left[N_i + \frac{1}{\Delta SIR} \left(RFI_o + M_o + D_o \right) \right]} \qquad (3\text{-}4)$$

where:

E_b/N_o = required signal-to-noise power density per bit

N_i = input thermal noise density at TDRS

P_u = $(EIRP)_{User} \times G_{TDRS} \times \alpha_T$ = received user power at TDRS

α_T = all losses including space, polarization, and scan

FEC = forward error control coding gain

ΔSIR = improvement in signal-to-interference ratio provided by AGIPA

RFI_o = unintentional interference signal power density

M_o = intentional multipath interference signal from desired and other in-band LDR users

$$= \frac{1}{CR}\left[K\left(\frac{4\,P_u}{\alpha} + \frac{n-1}{20} \times R \times P_u\right)\right]$$

CR = PN chip rate

K = an attenuation factor proportional to the refeléction coefficient = 1.0

n = the number of users/channel = 20

R = the ratio of average multipath contribution of other users to the desired signal power (≈ 7 dB for the 1976 user s/c distribution)

α = 2h/300; h = user altitude = 300 km (worst case for multipath)

D_o = intentional direct interference signal from other in-band LDR users

$$= (n-1)\,P_u/CR$$

The sample link calculation was computed for an RFI_o of -160 dBm/Hz, and user EIRP of +4 dBW. The ΔSIR improvement is a minimum value based on results of an RFI model analysis conducted during the study. (RFI Model analysis has shown improvements varying from approximately 5 dB to 18 dB, depending on desired-to-interference signal spatial and polarization relationship). The calculation shows that even in the relatively large RFI environment of -160 dBm/Hz that AGIPA with minimum adaptive processing gain provides over 5 kbps of data and 16 kbps with maximum improvement. The F-FOV mode can support approximately 1 kbps. Equation 3-4 has been plotted in Figure 3-6 for both the prime Jr. AGIPA mode and the F-FOV backup mode. The assumptions used in the computation are shown in the same figure where E_b/N_o for ΔPSK with a BER of 10^{-5} is 9.9 dB. Notable in Figure 3-6 at the low RFI_o levels is the increase in data rate with Jr. AGIPA due to its capability to reject intentional in-band other user signals, and due to the reduced antenna scan loss at scan limits due to the use of an electronically scanned array. Typically at a RFI_o of -190 dBm/Hz the

TABLE 3-5. LDR - RETURN LINK BUDGET

		AGIPA Mode	F-FOV Mode
		Data	Data
Modulation		ΔPSK	ΔPSK
User EIRP	dBw	4.0	4.0
TDRS antenna gain (peak)	dBi	17.4	17.4
Losses - space + scan (1)	dB	- 169.3	-173.1
- polarization	dB	- 0.5	- 0.5
Received power	dBw	- 148.4	- 152.8
System noise temperature (2)	dB	30.2	30.2
Noise density - thermal	dBw/Hz	- 198.4	- 198.4
multipath (3)	dBw/Hz	- 200.1	- 200.7
Other direct signal (3)	dBw/Hz	- 195.6	- 196.7
RFI_o (4)	dBw/Hz	- 190.0	- 190.0
Total noise density	dBw/Hz	- 188.2	- 188.4
TDRS Δ CNR degradation (5)	dB	- 1.0	- 1.0
C/N_o	dB-Hz	38.8	34.6
FEC coding gain (6)	dB	5.25	5.25
Available C/N_o (w/o AGIPA process gain)	dB-Hz	44.05	39.85
E_b/N_o (7)		9.9	9.9
AGIPA process gain (8)	dB	5.0	--
Achievable bit rate	dB	39.32	--
	bps	≈ 8500	≈ 1000

NOTES:

1. Worst case combination of space and scan losses since they do not maximize at the same spacecraft aspect angle.
2. Transistor front end ($T_e \approx$ 100 K) assumed at TDRS, includes line loss of 0.2 dB.
3. Chip rate (CR) = 10^6 bps.
4. RFI_o of -160 dBm/Hz assumed for link budget calculation.
5. A CNR degradation (ΔCNR) of -1.0 dB assumed for the TDRS transponder
6. Coding gain achieved with Forward Error Control (FEC) using Rate 1/2 constraint length 7 with a Viterbi decoder.
7. E_b/N_o = 9.9 dB for ΔPSK at 10^{-5}.
8. Minimum AGIPA processing gain included. Actual AGIPA performance in RFI model has shown 5 to 18 dB signal-to-interference ratio improvement over F-FOV approach.

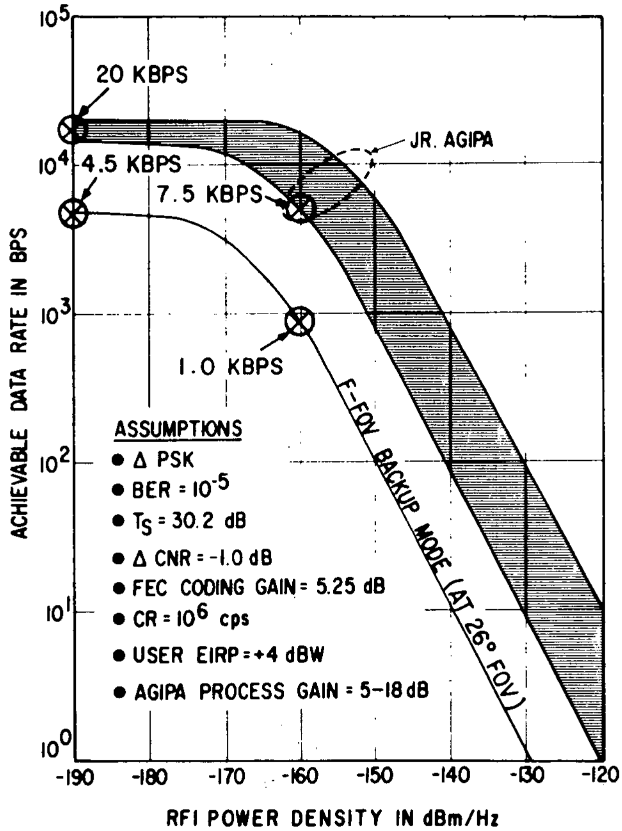

FIGURE 3-6. LDR FORWARD LINK: DATA RATE VS RFI POWER DENSITY

F-FOV is limited to approximately 4.5 kbps whereas the Jr. AGIPA can support approximately 20 kbps, using the same quad-array.

3.1.4.2 MDR AND HDR LINK PERFORMANCE

The performance for the MDR and HDR links can be computed using the following expression for the achievable data rate (H):

$$H = \frac{EIRP\ G_R\ \alpha_T\ \Delta CNR\ (FEC)}{(E_b/N_o)\ (K\ T_s)\ Margin} \qquad (3\text{-}5)$$

where

EIRP = effective radiated power above an isotropic radiator and is the product of transmitted output power (P_t) and antenna gain (G_t)

G_R = receive antenna gain

α_T = total losses including space, antenna pointing and polarization

KT_s = thermal noise power density where T_s is the system noise, temperature

E_b/N_o = signal-to-noise power density per bit

ΔCNR = carrier-to-noise ratio degradation through TDRS relay transponder

FEC = forward error control coding gain

Margin = system design margin

A tradeoff was performed to determine the optimum operating frequency (X-band versus K_u-band) for the HDR space-to-space link as described in Appendix A. The results indicate that optimum link performance and minimum hardware impact is obtained by operating this link in the K_u-band spectral range.

The link power budget calculations for the MDR and HDR forward and return space-to-space links are shown in Tables 3-6 and 3-7, respectively, inclduing the assumptions used. The results of Table 3-6 are plotted parametrically for the unmanned and manned MDR users as a function of the achievable data rate (H) versus the user antenna gain for the forward link in Figure 3-7. It is to be noted in particular that the required data rate of 54 kbps for the manned space shuttle can be readily achieved without employing forward error control (FEC): and that with FEC coding gain of +3.8 dB (Rate 1/2 constraint length with Viterbi decoder) that a data rate of over 150 kbps can be achieved with a EIRP of +47 dBW. At a reduced EIRP of +41 dBW, approximately 40 and 16 kbps can be supported to the manned user with and without FEC, respectively; and approximately 3 kbps to an unmanned user with an isotropic (G = 0 dBi) antenna).

The MDR return link performance is shown in Figure 3-8 as a function of achievable data rate (H) versus user EIRP. It is seen that the required manned user requirement of 192 kbps can be met with a user EIRP of +14 dBW or 25 watts which is well within the 40 watts available on the manned user (Space Shuttle).

The HDR link performances is highly dependent on the size of the user antenna aperture. Since the return link performance primarily dictates the HDR user's antenna requirements, the return link performance as shown in Figure 3-8 is described initially. Typically, to achieve a data rate of 100 Mbps Figure 3-8 shows that the user must generate an EIRP of +48 dBW without FEC coding gain. If we assume than a typical 100 Mbps user could generate an RF power output (P_t) at K_u-band of 5 watts or +7 dBW, the user requires an antenna gain of +41 dBW, or a relatively small antenna aperture with a diameter of 0.91 meter. Since this EIRP of +48 dBW was considered readily achievable for the typical 100 Mbps user, the more complex FEC processor operating in real time at 100 Mbps (although considered practical)[1] was not deemed necessary and not included in this analysis. However for the typical 10 Mbps user, FEC was used and Figure 3-8 shows that such a user requires an EIRP of +33 dBW. For this analysis we assumed that the 10 Mbps user could generate an RF power of +2 dBW and carry a 0.38 meter antenna with a gain of +31 dBi.

These typical antenna sizes for the 100 Mbps and 10 Mbps HDR users were used in Figure 3-9 to determine the achievable data rate in the

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(1) Private conversation between Magnavox and Linkabit Corp. on 15 November 1972.

TABLE 3-6. MDR/HDR FORWARD LINK BUDGET

Parameter	MDR (S-Band) Unmanned User	MDR (S-Band) Manned User	HDR (Ku-band) Unmanned User	HDR (Ku-band) Manned User
EIRP , dBw	41.0	41.0/47.0	23.6	53.6
• P_T , dBw	5.5	5.5/11.5	-30	0
• G_T , dBw	35.5	35.5	53.6	53.6
G_{User} , dBi	G_u	G_u	G_u	G_u
a_T = Space , dBi	-192.0	-192.0	-208.1	-208.1
Others , dB	-0.6	-0.6	-0.7	-0.7
T_s , dB	29.1 (1.1)	27.3 (1.2)	29.3 (1.3)	29.3 (1.3)
KT_s, dBW/K-Hz	-199.5	-201.3	-199.3	-199.3
CNR (2) , dB	-0.25	-0.25	-0.25	-0.25
System Margin, , dB	3.0	3.0	3.0	3.0
Available C/N_o, dB-Hz	$G_u + 44.65$	$(46.45/52.45) + G_u$	$G_u + 10.85$	$G_u + 40.85$
BER	10^{-5}	10^{-3}	10^{-5}	10^{-5}
E_b/N_o (3) , dB	9.9	7.4	9.9	9.9
Achievable H , dB	$G_u + 34.75$	$G_u + (39.05/45.05)$ (4)	$G_u + 0.95$	$G_u + 30.95$
	$D^2 + 58.87$	$D^2 + (63.17/69.17)$ (4)	$D^2 + 42.23$	$D^2 + 72.23$

NOTES: 1. $T_s = T_{ant} + (L_1 - 1)\ T_o + T_{e_1}\ L_1 + \frac{T_{e_2}\ L_1}{G_1}$ ....

	T_{ant} - °K	L_1 - dB	T_{e_1} - °K	T_{e_2} - °K	G_1 - dB	T_s - dB
(1.1)	65	2.1	350	-	15	29.1
(1.2)	123	3.0	50	350	15	27.3
(1.3)	100	1.5	450	-		29.3

2. CNR loss through TDRS relay transponder.
3. Δ PSK
4. With FEC coding (Rate 1/2 constraint length 7 with a Viterbi decoder) gain; H can be increased by 3.8 dB.

TABLE 3-7. MDR/HDR RETURN LINK BUDGET

PARAMETER	MDR (S-BAND)		HDR (K_u-BAND)
	UNMANNED USER	MANNED USER	ALL USER
Bit Error Rate	10^{-5}	10^{-4}	10^{-5}
E_b/N_o , dB	9.9	8.7	9.9
EIRP , dBw	EIRP	EIRP	EIRP
G_{TDRS} , dBi	36.2	36.2	52.8
α_T = SPACE(1)	-191.1	-191.1	-208.1
OTHERS	- 0.7	- 0.7	- 0.7
Δ CNR(2) , dB	- 1.0	- 1.0	- 1.0
T_s , dB	26.2(3.1)	26.2(3.1)	26.9(3.2)
KT_s dBW-Hz	- 202.4	- 202.4	- 201.7
SYSTEM MARGIN , dB	3.0	3.0	3.0
FEC CODING GAIN(4), dB	5.25	4.7	5.25
ACHIEVABLE H, dB	EIRP + 38.15	EIRP + 38.8	EIRP + 37.05

NOTES:

1. Value taken for worst case space loss + system temperature.
2. CNR loss through relay transponder
3. $T_s = T_{ant} + (L_1 - 1) T_o + Te_1 L_. + \frac{Te_2 L_1}{G_1} + \ldots\ldots\ldots$

	T_{ant} - °K	L_1 - dB	Te_1 - °K	Te_2 - °K	G_1 - dB	T_s - dB
(3.1)	234	1.45	50 (5)	350	15	26.2
(3.2)	253	1.30	100	-	15	26.9

4. Rate 1/2 constraint length 7 with a Viterbi decoder.
5. Te_1 = 200°K is adequate to meet Space Shuttle requirement for 192 kbps.

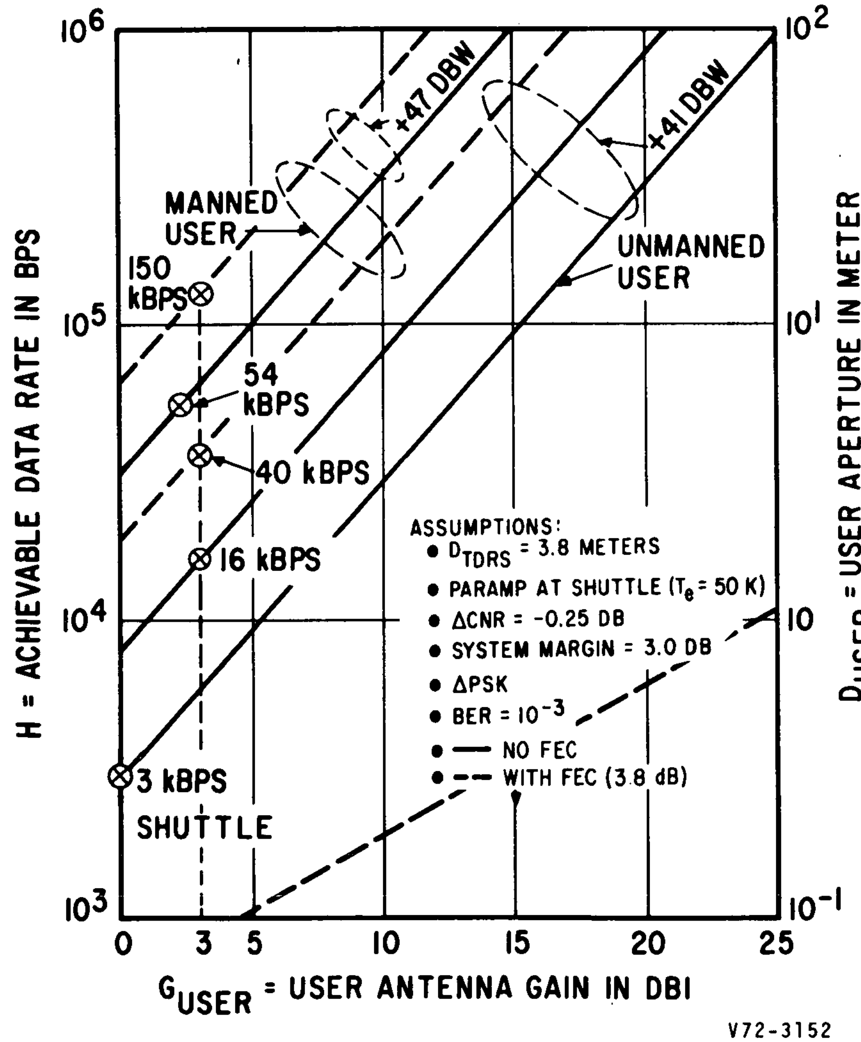


FIGURE 3-7. MDR FORWARD LINK: DATA RATE VS G_{USER}

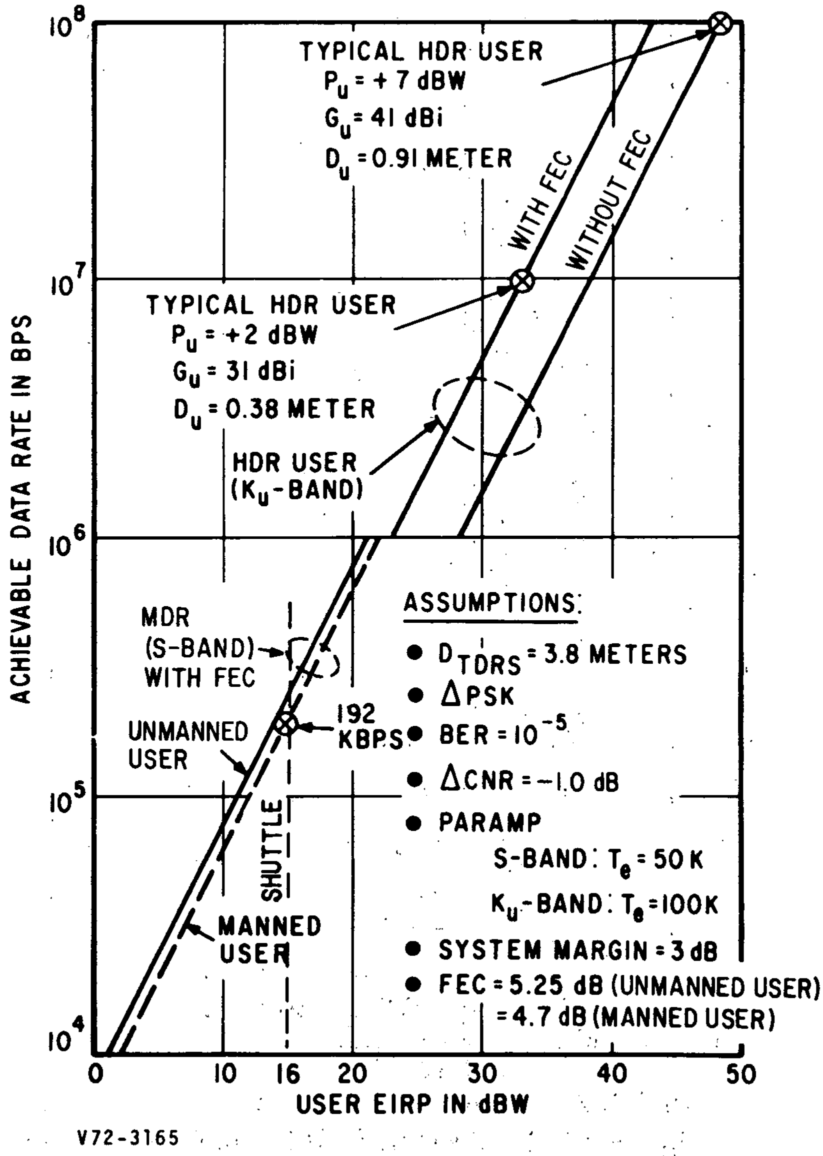


FIGURE 3-8. MDR/HDR RETURN LINK: EIRP VS DATA RATE

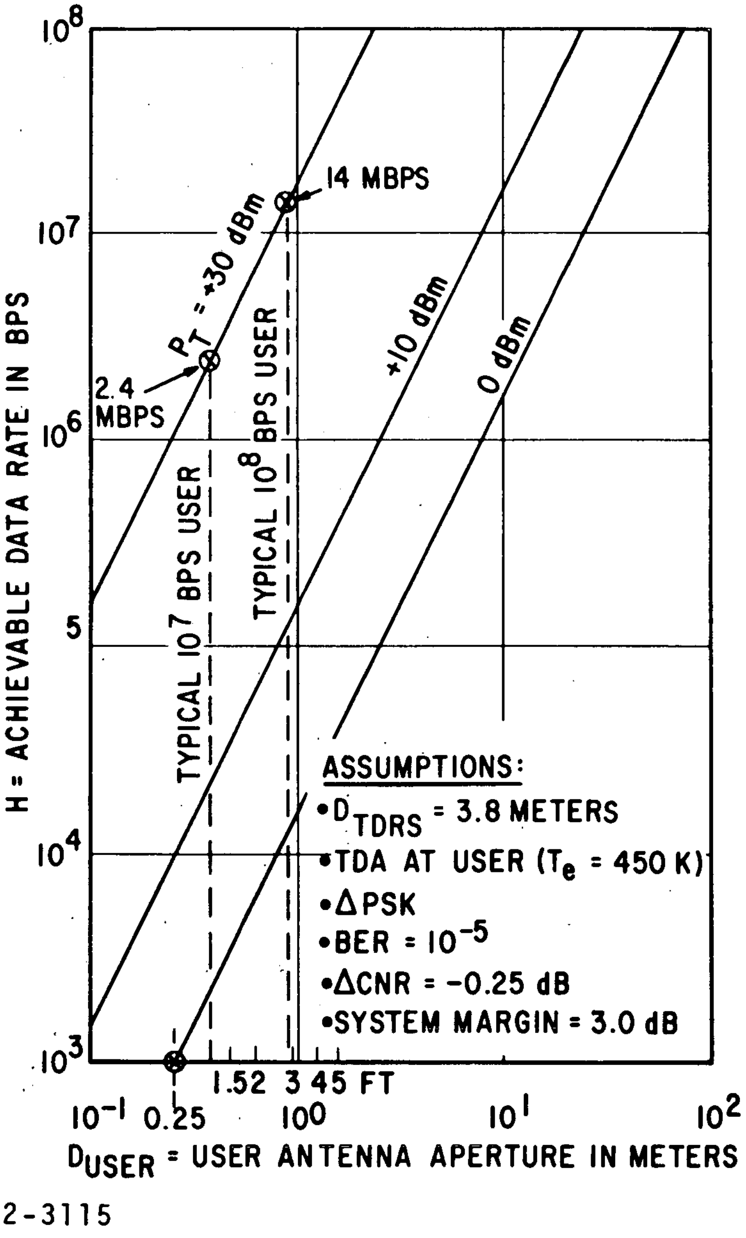


FIGURE 3-9. HDR FORWARD LINK: DATA RATE VS D_{USER}

forward HDR link. With one watt or +30 dBm radiated from the TDRS, a data rate of 14 Mbps and 2.4 Mbps can be transmitted to the typical 100 Mbps and 10 Mbps users, respectively, at a BER of 10^{-5}. The 14 Mbps at a BER of 10^{-5} is equivalent to 25 Mbps at 10^{-3} which is adequate to send commercial quality black and white television to a manned user with an antenna aperture of 0.91 meter in diameter. The required 1 kbps of data can be sent to an HDR user having an antenna aperture of approximately 0.25 meter diameter (antenna gain of 29 dBi). The HDR transceiver is designed to provide a multilevel radiated power (P_t) output of +30 dBm and 0 dBm in order to provide the flexibility to service HDR users with commercial quality video as well as the required 1 kbps command data.

MDR/HDR AS TDRS/GS BACKUP - Capability is designed into the MDR/HDR Transceiver to permit it to function as the TDRS/GS Transceiver. Considering the forward link, the Ground Station normally transmits a spectrum depicted in Figure 3-10. This must be modified, however, because in this backup mode only one of the MDR/HDR Transceivers is servicing a user. If we assume that MDR/HDR #2 is in the backup mode, then MDR/HDR #2 slot (Figure 3-10) is not required and the HDR #2 Receiver, with its 150 MHz bandwidth, can receive the Ground Station's reduced bandwidth transmission. Table 3-8 presents a typical HDR channel power budget for this mode of operation. It can be seen that only 1.8 dBW (1.5 watts) is required for the HDR data. This is in comparison to +14.8 dBW (30 watts) required for normal operation. The reduced power requirement is due to the larger MDR/HDR antenna size and the more sensitive receiver (paramp front end versus mixer front end). In this mode, the return link to the Ground Station is transmitted as an FDM signal without frequency modulation. Figure 3-10 shows the baseband configuration. The HDR channel bandwidth of 150 MHz is not available, but is limited to 10 MHz and inserted in the baseband spectrum in place of the MDR #2 data. The transmitted spectrum is that shown in Figure 3-10B, suitably translated to K_u-band. Table 3-9 contains the Return Link Power Budget for this mode of operation. It is seen that the total power required is only 0.075 watts. However, the signal is in FDM format and must be amplified linearly to prevent intermodulation distortion. This distortion is reduced to tolerable levels by operating the power amplifier at least 10 dB below saturation. The K_u-band amplifier saturation is at 1.6 watts which is 13 dB above 0.075 watts.

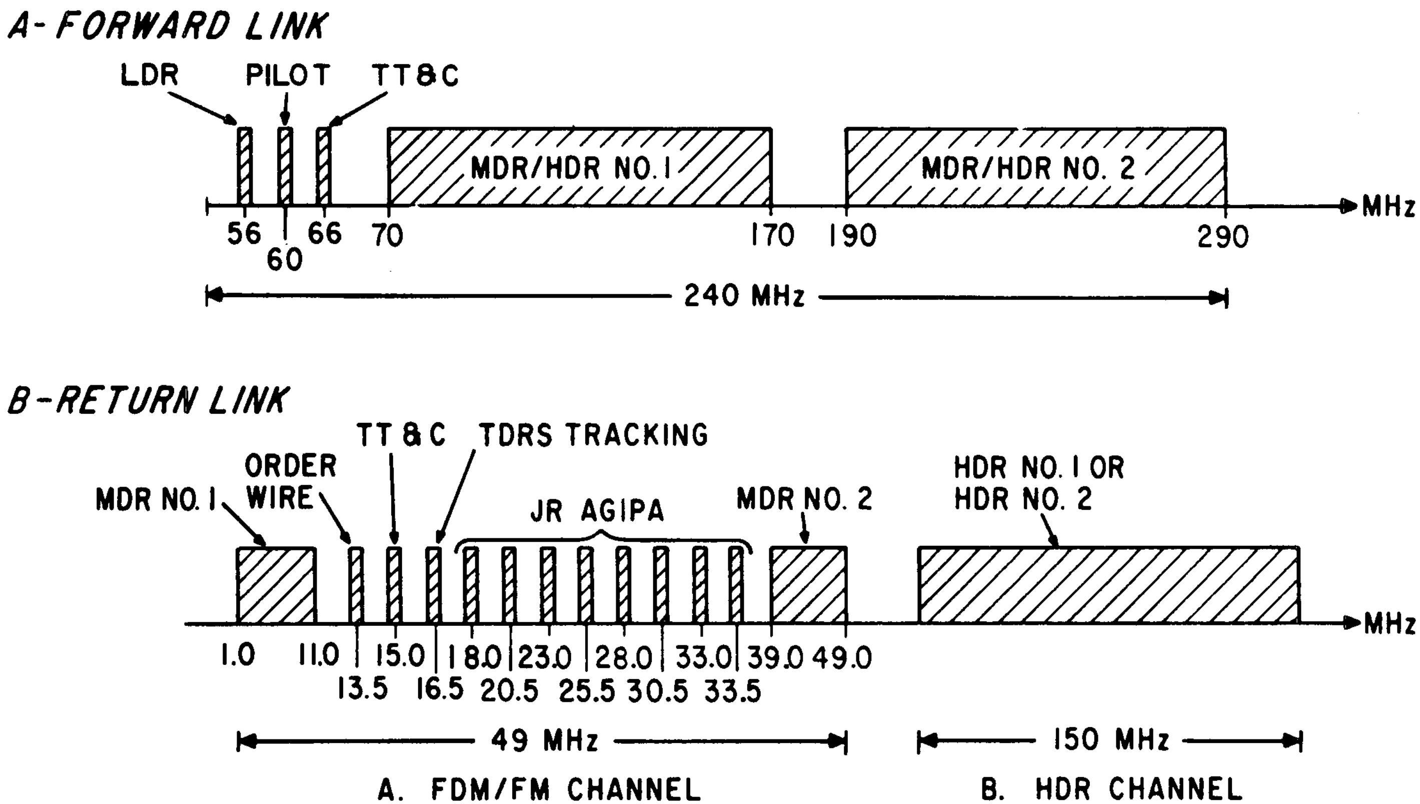


FIGURE 3-10. SPACE TO GS BASEBAND FREQUENCY LAYOUT: PRIMARY LINK

TABLE 3-8. MDR/HDR AS TDRS/GS BACKUP: FORWARD LINK BUDGET

	Units	Forward Link
CNR Required,	dB	22.8
RF Bandwidth,	MHz	15
System Noise Temperature,	dB	26.9 (1)
Thermal Noise Power,	dBw	-129.9
GS Antenna Gain,	dBi	67.0
α_T = Space	dB	-207.2
Other	dB	- 4.0
MDR Antenna Gain,	dBi	52.8
Rain Margin,	dB	17.5
Transmit RF Power,	dBw	1.8

NOTES:

1. Paramp ($T_e = 100^{o}K$)
2. Computed for HDR Video ($H = 10^7$ bps)

TABLE 3-9. MDR/HDR AS TDRS/GS BACKUP: RETURN LINK BUDGET

Parameters	Units	MDR#1	Order Wire	TT&C	TDRS Tracking	LDR (per Channel)	MDR#2
System Noise Temperature(1)	dB	25.2	25.2	25.2	25.2	25.2	25.2
GS Antenna Gain	dBi	67.1	67.1	67.1	67.1	67.1	67.1
α T - Space	dB	-208.1	-208.1	-208.1	-208.1	-208.1	-208.1
Others(2)	dB	-3.5	-3.5	-3.5	-3.5	-3.5	-3.5
MDR Antenna Gain	dBi	53.6	53.6	53.6	53.6	53.6	53.6
Rain Margin	dB	17.5	17.5	17.5	17.5	17.5	17.5
Bandwidth Required	MHz	10	1	1	1	2	10
CNR Required	dB	10	10	10	10	6	10
Thermal Noise Power	dBw	-133.4	-143.4	-143.4	-143.4	-140.4	-133.4
Transmit Power RQD/Channel	dBw	-15.0	-25.0	-25.0	-25.0	-26.0	-15.0
Total Transmit Power, Linear Amp	Watts	.075					
Backoff From Saturation	dB	10					
Saturated Power Required	Watts	0.75					

NOTES: (1). Two uncooled paramps in cascade.

(2). Comprised of the following losses (dB):

- Pointing (1.0)
- Polarization (0.5)
- Switch (.6)
- Diplexer (.5)
- Waveguide (.9)

3.1.4.3 TDRS/GS LINK CALCULATION

The TDRS/GS space-to-ground link forms the second leg of the spaceborne user-to-TDRS-to-Ground tandem link. In the return link, the transmitter is channelized FDM/FM/FDM system, where the HDR data is returned on one channel, and the remaining LDR (Jr. AGIPA), 2 MDR, Order wire, TT&C, and TDRS Tracking data is frequency division multiplexed (FDM) at a low baseband frequency and frequency modulated (FM). This resultant FDM/FM signal is returned in the FDM/FM channel as shown in to baseband frequency layout in Figure 3-10. In the FDM/FM channel the 49 MHz of FDM baseband signal is frequency modulated with a modulation index (B) of 1.09 to occupy a total RF bandwidth of 209 MHz (see Section 3.2.3 for details). Using a PLL demodulator at the GS, a CNR of 0 dB is required for the FDM/FM channel.

The HDR channel requires a 150-MHz RF bandwidth to send the 100 Mbps of data. The CNR required in this can be determined by referring to the curves showing the CNR requirement in a tandem link as shown previously in Figure 3-4. For a BER of 10^{-5} and a Δ CNR degradation of -1 dB for the return link, the CNR required for the space-to-ground link is 17.1 dB (CNR of 10.9 dB assumed for HDR space-to-space link). The resultant link power budget calculations of the two return link channels are shown in Table 3-10. The HDR channel will be designed with two levels of transmit output power, viz; 1) +10.2 dBW for operation with 17.5 dB of margin for operation in rain, and 2) 0 dBW for operation with 7.5 dB margin for operation in clear weather. The HDR channel can also be turned off to conserve prime power, e.g., during eclipse and/or at the end of life when the available prime power is decreased.

The GS/TDRS forward link is an FDM system which occupies approximately a 240 MHz baseband spectrum as shown in the previous Figure 3-10. The link power budget is computed for each data channel, using the CNR requirement for a tandem link as shown previously in Figure 3-4. A Δ CNR loss of -0.25 dB is used for the forward link, and the burden of the CNR requirement for the tandem link has been assigned to the ground-to-space link since the GS has the greater flexibility and inherent capacity to generate a large EIRP. The calculation for CNR requirement for each data channel is shown in Table 3-11, and the results are used in equation 3-5 to compute the link power budget for each data channel as shown in Table 3-12.

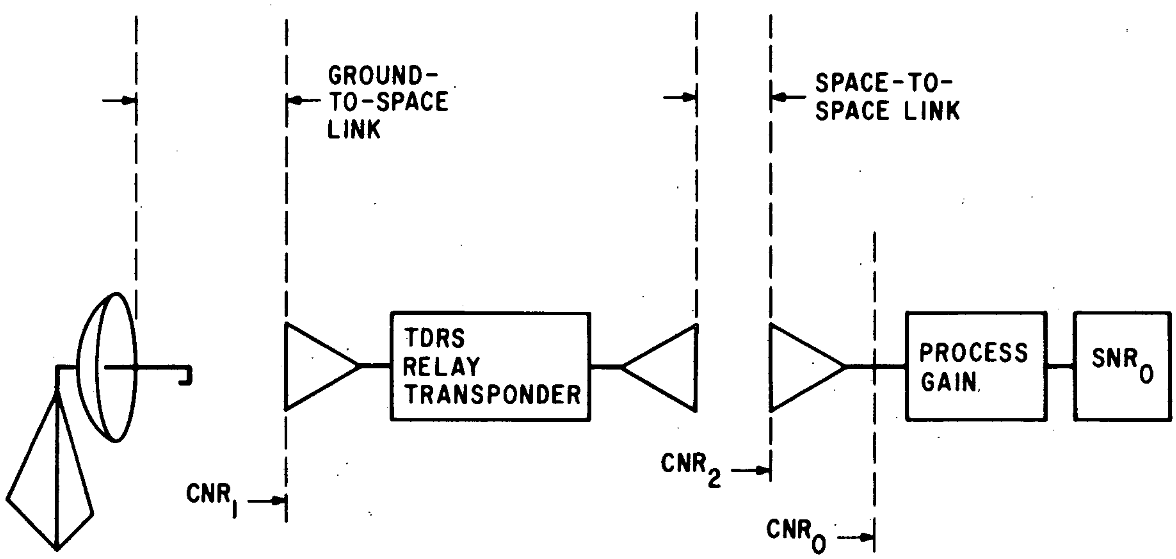
TABLE 3-10. TDRS/GS RETURN LINK BUDGET

Parameter		FDM/FM[1]	HDR
CNR required,	dB	0[4][6]	17.1
RF Bandwidth,	MHz	209	150
System Noise Temperature[2]	dB	25.2	25.2
Thermal Noise Power,	dBW	-120.2	-121.6
GS Antenna Gain[3]	dBi	67.9	67.9
α_T = SPACE	dB	-208.1	-208.1
Pointing and Polarization	dB	- 1.5	- 1.5
Rain Margin	dB	17.5	17.5
EIRP	dBW	39.0	54.7
TDRS Antenna Gain	dBi	47.1	47.1
TRANSMITTER RF POWER[5]			
• High Power Mode	dBW	- 5.5	10.2
• Low Power Mode	dBW	- 5.5	0.

NOTES:

1. Modulation Index = 1.09, PLL Demodulator
2. Two uncooled paramps in cascade
3. 18.3 meter diameter antenna
4. An experimental model of a wideband PLL demodulator was built and demonstrated to provide a threshold of 1.5 dB by Schwartz, L. and Chang, F.; "Wideband Low Threshold Phase Locked Demodulator," ADR 03-19-66.1 dated July 1966, Grumman Aerospace Co.
5. Includes internal losses (waveguide, diplexer, switches, etc.) of 2.6 dB
6. Schilling D. L. and Billig, J.; "A Comparison of the Threshold Performance of the Frequency Demodulator Using Feedback and the Phase Locked Loop," Polytechnic Institute of Brooklyn, 1965 Record of the International Space Electronics Symposium

TABLE 3-11. CNR REQUIREMENT FOR GS/TDRS FORWARD LINK



$$CNR_1 = \frac{CNR_0\,(CNR_2 + 1)}{CNR_2 - CNR_0}$$

DATA CHANNEL	SNR$_0$ (dB)	PROCESSING GAIN = $\frac{B}{1.5\,H}$ B-(MHz)	H-(bps)	PG-dB	CNR$_0$ (dB)	CNR$_2$ (dB)	CNR$_1$ (dB)
LDR	9.9	1.0	10^3	28.2	-18.3	-18.05	12.34
MDR							
• UNMANNED	9.9	10.0	10^3	38.2	-28.3	-28.05	12.27 (10 MHz) 3.52 (75 MHz)
• MANNED	7.4	32.0	54×10^3	25.9	-18.5	-18.25	12.34 (32 MHz) 8.64 (75 MHz)
HDR							
• UNMANNED	9.9	1.0	10^3	28.2	-18.3	-18.05	12.34 (1 MHz) -7.66 (100 MHz)
• MANNED	9.9	100.0	14×10^6	8.54	1.36	1.61	16.16

TABLE 3-12. GROUND STATION-TO-TDRS FORWARD LINK BUDGET

Parameter		LDR	MDR Unmanned	MDR Manned	HDR Unmanned	HDR Manned
Modulation		Δ PSK	Δ PSK	Δ PSK	Δ PSK	Δ PSK
Data Rate, kbps		1.0	1.0	54.0	1.0	10^4
CNR Required[1],	dB	12.34	3.52	8.64	-7.66	16.6
RF Bandwidth,		1.0	75.0	75.0	100.0	100.0
System Noise Temp.[2]	dB	33.6	33.6	33.6	33.6	33.6
Thermal Noise power,	dBW	-135.0	-116.2	-116.2	-115.0	-115.0
TDRS Antenna Gain,	dBi	46.4	46.4	46.4	46.4	46.4
Losses: Space,	dB	-207.2	-207.2	-207.2	-207.2	-207.2
Pointing,	dB	-1.0	-1.0	-1.0	-1.0	-1.0
Polarization,	dB	-0.5	-0.5	-0.5	-0.5	-0.5
Atmospheric,	dB	-0.4	-0.4	-0.4	-0.4	-0.4
Rain Margin[3],	dB	17.5	17.5	17.5	17.5	17.5
EIRP Required,	dBW	57.54	67.47	72.59	57.54	81.8
GS Antenna Gain[4],	dBi	67.0	67.0	67.0	67.0	67.0
GSTransmitter Power[5] Required,	dBW	-9.46	0.47	5.59	-9.26	14.8
	watts	0.11	1.11	3.62	0.12	30.2

NOTES:

1. See Figure 3-11 for the required CNR for the GS/TDRS link.
2. Uses mixer front end (NF - 7.5 dB); T_{ant} = 290K, Losses = 1.5 dB, T_s = 33.6 dB
3. Rain margin provides operation in 25 mm/hr rainfall rate.
4. 18.3 meter aperture with 75-percent efficiency
5. Includes internal losses (waveguide, diplexer, combiner, etc.) of 2.1 dB.

3.1.4.4 TDRS TRACKING/ORDER WIRE LINK CALCULATION

The link calculations for the TDRS Tracking Mode and the Order Wire Mode are shown in Table 3-13. The link calculation shows that the available C/N_0 in the tracking mode is approximately 70 dB-Hz, more than adequate to provide ranging accuracy within 0.05 meter.

The Order Wire provides support for the manned user (Space Shuttle) with a EIRP of +16 dBW. For this case, the link will support approximately 133 bps which is more than adequate to provide order wire service requests.

3.1.4.5 TRACKING, TELEMETRY AND COMMAND LINK CALCULATION

The TT&C operates in the 148-150 MHz for the forward command data transmission, and in the 136-138 MHz band for return telemetry data transmission. It is estimated that the command data rate will not exceed 128 bps and that the telemetry data will be approximately 16 kbps during the inflight transit phase but reduced to 1024 bps after the TDRS spacecraft is on-station at geosynchronous altitude.

Link calculations for this link is shown in Table 3-14, for PCM - ΔPSK modulation with a probability of bit error (BER) of 10^{-5}.

The VHF telemetry transmitter will be designed to provide a multilevel power capability of 6 and 1.0 watts to meet the real time burst requirement to send data rates up to 16 kbps from the sun sensor and thrust accelerometers, as well as to send 1024 bps during normal operation when on-station at geosynchronous altitude, respectively.

3.1.5 SYSTEM BLOCK DIAGRAM

A detailed block diagram of the overall telecommunication system for the Part II Alternate Uprated Delta 2914 Configuration is shown in Figure 3-11. This system provides simultaneous service and support for:

- 20 LDR users, and
- 1 or 2 MDR users, and
- 1 HDR user

TABLE 3-13. LINK BUDGET FOR TDRS TRACKING/ORDER WIRE TRANSPONDER

PARAMETER		TDRS TRACKING MODE		ORDER WIRE MODE
		Fwd	RETURN	
TRANSMITTER POWER,	dBW	3.0	10.0	16.0
TRANSMITTER ANTENNA GAIN,	dBi	14.5	43.3	3.0
TRANSMITTER LOSSES,	dB	-1.0	-1.0	-3.0
EIRP,	dBW	16.5	52.3	16.0
LOSSES – SPACE		-192.0	-191.3	192.4
OTHERS		-1.0	-1.0	-1.0
RECEIVE ANTENNA GAIN,	dBi	44.0 (2)	14.5	12.0
RECEIVED POWER,	dBW	-136.5	-125.5	-165.4
SYSTEM NOISE TEMPERATURE,	dB	21.8 (3)	27.5 (4)	27.5 (4)
THERMAL NOISE TEMPERATURE,	dBW/Hz	-206.8	-201.1	-201.1
TDRS Δ CNR LOSS (6),	dB	-0.5	-1.0	-1.0
DESIGN MARGIN,	dB	3.0	3.0	3.0
AVAILABLE C/N_0	dB-Hz	+70.8 (7)	71.6 (7)	31.7
Eb/No	dB-Hz			8.7
ACHIEVABLE DATA RATE,	bps			>100 (8)

NOTES:
1. AVAILABLE POWER FROM SPACE SHUTTLE
2. 9.1 METER ANTENNA
3. INCLUDES PARAMP (Te = 30K) AND 0.9 dB LOSSES
4. TRANSISTOR (Te = 360K)
5. INCLUDES PARAMP (Te = 50K) AND 4 dB TOTAL LOSSES
6. Δ CNR LOSS THROUGH TDRS RELAY TRANSPONDER
7. ADEQUATE C/No FOR RANGING ERROR (Δ R) OF 5×10^{-2} METER
8. ADEQUATE FOR PROVIDING ORDER WIRE SERVICE

{MSFN EQUIPMENT HANDBOOK MG - 403}

TABLE 3-14. LINK BUDGET FOR THE TRACKING TELEMETRY AND COMMAND TRANSPONDER

Parameter		Command	Telemetry	
			Transfer Phase	On-Station
H	, bps	128	16×10^3	1024
C/N_o required	, dB	31.0	51.9	40.0
T_s	, dB	30.2	30.9	30.9
KT_s	, dBw-Hz	- 198.4	-197.7	-197.7
Receive Antenna	, dBi	0.0	20.0	20.0
α_T = SPACE	, dB	- 168.3	$\leq 167.5^{(2)}$	-167.5
= OTHER	, dB	- 3.0	- 3.0	- 3.0
Margin	, dB	3.0	3.0	3.0
EIRP	, dBW	6.9	7.7	-4.2
	watts	4.9	5.9	0.38

NOTES:

1. Transistor ($T_{e_1} \approx 100$ K), $T_{ant} \approx 1000$ K
2. Inflight transit phase varies up to a maximum at geosynchronous altitude.

The LDR return link operates in the 136-138 MHz VHF band and utilizes a quad array of short backfire elements, providing a peak gain of 17.4 dBi. Both veritcally and horizontally polarized components from each antenna element are returned in a 2 MHz channel to the GS. In the prime LDR return link Jr. AGIPA mode, these eight channels are adaptively processed for each user to optimize the desired signal-to-interference ratio (SIR) using spatial and polarization discrimination of interference signals. Up to 20 LDR users are serviced by each TDRS satellite (or up to 40 LDR users for the entire TDRSS network). At the GS, unique PN Gold codes are used for each of 40 users to identify the desired signal. As a backup mode the signal from either: (1) two orthogonally polarized channels, or (2) all eight channels in an unadapted mode (beam pointed along spacecraft's local vertical) can be used to provide a F-FOV mode.

Referring to Figure 3-11, the signals received by the LDR-1 through LDR-4 receivers are code division multiplexed by a unique Gold code for each LDR user. Thus, the signal received over the eight channels consists of the 20 uniquely coded user signals, the user signal multipaths, RFI, and thermal noise. The receiver will be designed to handle RFI power levels of -92 dBm as shown in Figure 3-11. The eight receivers convert the composite antenna output signals to an IF frequency ranging from 48 to 65.5 MHz so that the eight channels can be frequency division multiplexed for transmission to the ground station. The 48 to 65.5 MHz IF frequencies were selected based upon obtaining an operating point where filter components which are small and stable could be employed to obtain the required phase linearity needed to pass the 1 MHz PN chip rate undistorted through the system. These IF filters only partially form the band pass response for the LDR channels which also include demultiplexing and processing at the GS. Therefore, to achieve the desired amplitude and phase linearity over a 1.5 MHz bandwidth, a bandwidth of 2 MHz is employed in the TDRS. The signals are linearly amplified in the four receiver modules and summed at LDR-9, the receiver summing network. The composite FDM signal is then further downconverted to a 19.5 to 37.0 MHz band which is designated for the eight LDR channels. AGC is applied to the total FDM/LDR signal to maintain a constant drive level into the TDRS/GS FM modulator, regardless of RFI or noise fluctuations.

Each mixer in the LDR receivers is controlled by a phase-locked crystal VCO which is locked to a reference from the frequency source. Therefore, all mixing is performed coherently; in addition, the VCO/PLL is designed so that each one can be commanded to operate open-loop as a noncoherent local oscillator source should a malfunction develop in the pilot channel or frequency source reference.

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The LDR forward link operates in the 400.5 to 401.5 MHz UHF band whose prime mode is the F-FOV mode which utilizes a single disc-on-rod surface wave structure to illuminate the entire 31° FOV. This signal consists of a common (to all user) PN code which is used to coherently synchronize all of the LDR users within the F-FOV for range and range rate measurement, and to synchronize each of their on-board unique PN code generators, as well as a unique PN code address when command data is to be sent to the addressee. Command data is sent sequentially to each of the users. This F-FOV antenna provides a gain of 15, 13, and 12 dBi at the peak, 26 and 31° FOV points. The RF output power is such that an EIRP of +30 dBW can be provided at 26° FOV under normal daylight operation, but also with a reduced capability to provide an EIRP of +27 and/or +24 dBW under eclipse or other periods when spacecraft prime power conservation is required.

The LDR forward link has also a Emergency Steered Beam Backup Mode which uses four antenna and four transmit channels identical to the one provided for the prime F-FOV (three additional antenna/transmit channels). Each of these four antenna/transmit channels are implemented with phase shifters such that a steerable phase array is formed. As a result, the additional 6 dB gain of the quad array is now available to increase the normal operational EIRP from +30 to +36 dBW without essentially increasing the prime power requirements (EIRP of 24 dBW from each transmit channel). If the EIRP per channel is increased to +27 and +30 dBW, the EIRP in the Steered Beam Mode is increased to +39 and +42 dBW, respectively. This additional EIRP of up to +12 dBW can be effectively used to increase the system gain; especially when command data must be transmitted to a user spacecraft in an emergency condition and/or in a highly dense RFI environment when the normal EIRP of +30 dBW in the F-FOV made is inadequate. Since this emergency Steered Beam Mode uses four identical antenna and transmit channels as the F-FOV mode, it provides an inherent functional quadruple redundance to the prime F-FOV mode.

Referring to Figure 3-11, the LDR forward signal to be transmitted is received from the TDRS/GS receiver by the LDR Transmitter Divider Network (LDR-10) at 56 MHz. The forward link signal is spread with a PN chip rate of 668 kcps to meet the IRAC flux density guidelines of -150 dBW/m^2/4 kHz when received on the earth's surface. The 56 MHz is upconverted to a transmit frequency of 401 MHz. The mixer output signal is filtered and divided four ways to feed each of the four identical transmit channels. Each channel includes a phase shifter (controlled via commands from the GS) which has been included to steer the beam in the emergency mode. In the F-FOV since only one channel is used, the setting of the phase shifter has no effect on the transmitted command signal. Each transmitter chain amplifies the signal to a maximum RF power output level of 57.5 watts or +17.6 dBW, or +17.0 dBW into the antenna. The final power amplifier and the driver amplifier are also designed for each to provide 3 dB less gain so that the effective RF power into the antenna is reduced to +14 and 11 dBW, respectively. Therefore, when combined with the antenna gain, each transmit channel provides an EIRP of +30, +27, or +24 dBW at 26^{o} FOV.

The Telecommunication System is also designed to service and support MDR and HDR users. Rather than to provide two separate transponders which are each designed to support either MDR or HDR users, the two transponders are designed almost identically such that either transponder can service an MDR or HDR user at S- or K_u-band, respectively. In addition, both transponders are designed to provide a functional backup to the TDRS/GS transponder at K_u-band as well as at S-band. Each transponder employs a 3.8 meter deployable antenna with a dual S-/K_u-band feed. The MDR/HDR #1, 3.8 meter antenna can also be switched in place of the 1.8 meter TDRS/GS antenna to provide an additional 6.5 dB margin for operation in rain, for a total rain margin of 17.5 +6.5 dB, or 24 dB.

Each transponder is designed such that both S- and K_u-band return link data can be received simultaneously from a common MDR/HDR user; however, forward command data can be sent only at S-band or K_u-band to the same user.

At S-band, antenna pointing is open loop since the spacecraft attitude stabilization and antenna pointing accuracies (0.5 to 1.0 degree) are well within the 2.5 degree half power beam width (HPBW) of the antenna. However, at K_u-band, the HPBW is approximately 0.4 degree; therefore, a closed loop auto-tracking pseudo-monopulse technique is employed. The acquisition and tracking procedure for both MDR and HDR users assumes that each user has an S-band command receiver with an isotropic (G = 0 dB) antenna on-board. This receiver is then used to command the MDR and HDR users to point his respective antenna toward the TDRS relay spacecraft. The step-by-step procedure for the MDR and HDR user is shown in Table 3-15.

Referring to Figure 3-11, the MDR/HDR #1 and MDR/HDR #2 transponders are seen to be essentially identical with the exception that MDR/HDR #1 has some additional circuitry which permit its 3.8 meter antenna to replace the TDRS/GS 1.8 meter antenna. Since the transceivers are essentially the same, the discussion herein applies to both although for discussion purposes reference is madeto MDR/HDR #1. The receiver can simultaneously receive signals at S- and K_u-band RF front end uses

TABLE 3-15. ACQUISITION AND LOCK-ON PROCEDURE FOR MDR AND HDR USER

S-Band MDR Link

- TDRS 3.8 meter antenna is pointed (via ground command) toward MDR user.
- The MDR user receives ground commands via the MDR space-to-space link and his on-board S-band command receiver to point his antenna toward the TDRS spacecraft.
- The MDR user is commanded to turn on his MDR space-to space link transceiver, and commence data transmission.

Ku-Band HDR Link

- TDRS 3.8 meter is pointed via ground command toward the approximate location of the HDR user.
- The HDR user is commanded via the S-band Command receiver and the TDRS S-band MDR space-to-space link to point his HDR antenna (spoiled beam mode) toward the TDRS spacecraft. It is assumed that the HDR user will have a dual-beam mode capability; initially a spoiled beam mode with sufficiently wide beamwidth ($\approx 5^{\circ}$) to illuminate the TDRS spacecraft, and subsequently a high gain pencil mode for normal data transmission.
- The HDR user shall be commanded to send a simple AM modulated tone (or carrier).
- The TDRS HDR antenna will be commanded via the ground to initiate an acquisition search scan. The total scan coverage will be designed to cover approximately 5 degree conical volume to insure including the known angular uncertainties; viz., antenna pointing, spacecraft altitude, and known user ephemeris. This 5 degree conical coverage can be increased if required.
- After TDRS has acquired the HDR user, the HDR user will be commanded to go into his narrow beam mode, and go into a simple acquisition search procedure until he acquires the beacon tone (or command data) transmitted from the TDRS spacecraft.
- After user lock-on, the HDR user will be commanded to initiate data transmission.

a parametric amplifier (Te_1 = 50 K) in cascade with a transistor amplifier (Te_2 = 350 K), resulting in overall system noise temperature (including antenna temperature ≈ 234 K) of 26.2 dB. The S-band signal which can be located anywhere in the 2200 to 2300 MHz range is downconverted to a 500 MHz IF by mixing with a tunable (in discrete 5 MHz step) 1725 to 1775 MHz local oscillator signal that is coherently locked to a reference from the Frequency Source. The resultant signal is filtered by a 10 MHz band pass filter and then down converted to either a center frequency of 6 or 45 MHz. The selection of the second downconversion is done via ground command, and is provided to enable either MDR/HDR #1 or #2 receivers to be fully interchangeable such that either receiver can occupy either of the two selected MDR baseband frequency slots (see Figure 3-10). The output of the receiver drives the Weighted Combiner Network in the TDRS/GS FDM/FM channel transmitter.

When the MDR/HDR #1 is used in S-band as a backup to the TDRS/GS transceiver, the output of the S-band channel 500 MHz IF is switched to the TDRS/GS receiver.

The K_u-band receiver is a fixed tuned 150 MHz wide receiver which uses a parametric amplifier with a noise temperature of 100 K, resulting in a system noise temperature (including 253 K antenna temperature) of 26.9 dB. The K_u-band signal is downconverted to 500 MHz by a fixed tuned (14.425 GHz) VCO/PLL, and filtered in a 150 MHz wide band pass filter. The output of the filter is amplified and sent to the TDRS/GS transmitter for subsequent upconversion to 14.675 GHz and power amplification of a TWT amplifier.

When the MDR/HDR #1 is used in K_u-band as a backup to the TDRS/GS Transceiver, the output of the K_u-band channel 500 MHz band pass filter is switched to the TDRS/GS receiver. In addition, the MDR/HDR transponder which is to be used for the space-to-space link is designed such that its HDR channel can also be used simultaneously with the MDR channel. In this mode the amplified 500 MHz signal is downconverted to 45 MHz to fill the frequency slot normally occupied by the other MDR channel (now used as a backup to the TDRS/GS Transceiver). This 45 MHz signal is amplified and sent to the Weighted Combiner Network in the TDRS/GS transmitter where it is subsequently combined with the other data and routed to the backup MDR/HDR transmitter.

On command transmission at S-band, the MDR signal is received from the TDRS/GS receiver at 120 MHz (240 MHz for MDR/HDR #2) and upconverted to S-band. The signal is spread at the Ground Station to 10 MHz for an unmanned user, and 32 MHz for the manned user; however, the MDR transmitter is designed such that this signal can be located anywhere within a 75 MHz bandwidth (2.025 to 2.1 GHz), such that no tuning is required on the TDRS spacecraft. The operating frequency is determined and selected completely at the ground station. This S-band signal is amplified in a two stage solid state amplifier to +43.8 dBm which when combined with the antenna gain provides an EIRP of +47 dBW to support the manned user. The S-band transmitter is also designed to provide an EIRP of +41 dBW by bypassing the final solid-state power amplifier.

For command data transmission to HDR users, 100 MHz of instantaneous bandwidth is available. A common S-band MDR channel is shared up to the final S-band solid-state power amplifier. This can be done since both MDR and HDR data transmission to a common MDR/HDR user are not required simultaneously, as in the return links. Therefore, the forward link HDR data received from the TDRS/GS receiver at 120 MHz (or 240 MHz for MDR/HDR #2) is upconverted to S-band through the same converter-amplifier chain as the MDR command data. The signal is power divided prior to the solid state S-band amplifier, upconverted again to K_u-band (14.860 GHz) and fed to a K_u-band solid-state amplifier. This amplifier is a dual-mode amplifier, providing an RF power output of +2 dBm or +32 dBm, or when combined with the antenna gain, an EIRP of +23.6 or +53.6 dBW. This EIRP is adequate to provide the required 1 kbps of command data to an HDR user with a 0.25 meter aperture; or if desired, video to a manned user with an aperture of 0.91 meter in diameter. The selection of the power level is accomplished via ground commands.

The TDRS/GS transmitter is a channelized system comprised of an FDM/FM channel and an HDR channel. The outputs of the LDR, 2 MDR, TDRS telemetry, TDRS Tracking and Order Wire receivers are all summed and transmitted to the ground in the FDM/FM channel. These signals are initially combined and pre-emphasis weighted in the Weighted Combiner Network, then amplified, and used to frequency modulate a K_u-band VCO with a modulation index of 1.0^o, the resulting spectrum occupying total RF bandwidth of approximately 200 MHz. This FDM/FM signal is amplified to an RF power level of 0.27 watts or -5.7 dBW in a K_u-band solid-state amplifier. This provides a 17.5 dB margin for operation in approximately 25 mm/hr of rain.

The HDR transmit channel selects either the HDR input from MDR/HDR #1 or #2, the upconverts the signal to 14.675 GHz. This signal is then amplified in a dual mode traveling wave tube amplifier (TWTA) to RF output level of +40.2 dBm or +30.2 dBm. When combined with the 1.8 meter antenna, this provides an EIRP of 57.3 and 47.3 dBW for operation in rain and clear weather, respectively. This dual mode operation is obtained by changing the B+ voltages on the TWTA. When the overall operational plan necessitates, this HDR channel can also be turned off in order to conserve the prime power on spacecraft. Since TWT's have a relatively high failure rate ($\approx 6000 \times 10^{-9}$), each TWTA is designed with two TWT's sharing a common power supply system. The TWT's are hard wired such that no high voltage switching is required--only the heater of the desired TWT is energized. In addition, as with all active components of the Telecommunication System, the TWTA is fully redundant and quadruple redundance is obtained for the TWT. This provides a relatively high TWTA reliability such that an overall telecommunication goal of better than 90 percent is achieved.

The ground-to-space forward link signal is received in the TDRS/GS receiver. This signal is an FDM signal consisting of the LDR, 2 MDR/HDR, TDRS Command, TDRS Tracking and coherent pilot reference signals, and occupies an RF bandwidth of 240 MHz from 13.4 to 13.64 GHz. This signal is received in a 1.8 meter antenna that is two-axis gimbal mounted, and employs a pseudo-monopulse acquisition and tracking circuit to acquire and maintain lock to the ground station carrier.

The received signal from the GS is downconverted in a mixer front end to an 500 MHz IF. A mixer front end is used in lieu of a TDA or parametric amplifier, since the GS has the capability to generate a large EIRP to make up for the difference in system gain, and to minimize the weight, power and size complexity in the TDRS Spacecraft. The MDR/HDR channels have been designed 100 MHz wide such that the operational frequency selection can be made at the GS without tuning the TDRS/GS receiver. The 500 MHz first IF is filtered, amplified and downconverted to the VHF band. The VHF signal is then demultiplexed into the respective LDR, 2 MDR/HDR, TDRS command, and coherent pilot reference channels.

The 60 MHz pilot reference is used in the Frequency Source component in Figure 3-11 to generate all the coherent reference signals used on-board the TDRS spacecraft. A phase locked loop is used to frequency lock the crystal stabilized oscillator whose output is then used in the Frequency Multiplier section to generate the 10 reference signals to be used by the other components of the Telecommunication System. The reference signals as required are combined on discrete lines for transmission to the other subsystem components of the Telecommunication System, and are used to coherently lock its local VCO. In this manner, if the prime 60 MHz pilot reference from the GS is lost, the main VCO in the Frequency Source can free run and maintain coherence between the component of the on-board electronics. As a functional backup, since the S-band Tracking and Order Wire Transponder illuminates the earth, the GS can use this signal to coherently lock to the spaceborne reference, reversing the normal operational procedure.

The TDRS Tracking, Telemetry and Command subsystem shown in Figure 3-11 operates in the VHF band during the inflight transit phase, but operates at K_u-band through the main TDRS/GS space-to-ground transponder when on-station at geosynchronous altitude. At K_u-band, the TDRS command data from the GS is received from the TDRS/GS receiver at 66 MHz. This signal also contains the PN, code at 500 kcps which has been modulo 2 added to the command data and is subsequently modulo 2 added to a 15 MHz reference signal. This combined PN_1 X 15 MHz is sent ot the TDRS Tracking and Order Wire Transponder where it is sent to 2 remote ground transponders GS and returned with a new PN code, coherently related to PN_1. These returned signal are used subsequently for making location measurements of the TDRS spacecraft through trilaleration techniques.

The 66 MHz signal is received in the TT&C Modem unit where a Costas Loop is used to separate the command data as well as to coherently lock the on-board PN_1 generator. The command data is Δ PSK demodulated and sent to the TT&C Processor where it is decoded, demultiplexed and distributed to the approximately nine analogs (after digital-to-analog conversion) and 135 digital command points.

The TDRS housekeeping telemetry data from approximately 132 analog and 173 digital points are multiplexed in the TT&C Processor unit. The output of the Multiplexer is sent to the TT&C Modem unit where it is Δ PSK modulated and modulo 2 added to the PN_1 Code and 15 MHz reference signal, and sent to the TDRS/GS FDM/FM channel transmitter for transmission to the GS.

In the VHF TT&C Mode the command data and ranging signals from the TDRS or STDN Ground Stations are received in the VHF Transceiver. In the command data mode the signal is received in the 148 to 150 MHz band (typically at 148.26 MHz), downconverted to 12.15 MHz, and this output is sent to the TT&C Modem unit for demodulation in the STDN demodulator. In the ranging mode, the 12.15 MHz is downconverted, amplified, upconverted to the 136 to 138 MHz band, amplified, and returned to the Ground Station. The 22.685 MHz coherent local oscillator used in the initial receiver mixer stage and the transmit upconverter is derived from the received signal in the STDN VCO/PLL. On transmit, an RF power level of approximately six watts is generated for transmission of peak data rates of 16 kbps during the inflight transit phase, and approximately 0.5 watt for sending the normal 1024 bps on-station. It is to be noted that the receiver and transmitter uses separate antennas, thereby eliminating the need for a diplexer.

The TDRS Tracking and Order Wire transponder is the last component shown in Figure 3-11 and provides means for trilateration measurement of the TDRS spacecraft position; and an Order Wire receiver for establishing priority access to the MDR/HDR Transponder, as well as a pilot frequency reference to which the GS can coherently lock-on in the event that the normal reverse mode becomes unavailable. In the TDRS Tracking Mode, the TDRS Tracking and Order Wire Transponder receives a PN_1 Code Modulo 2 added to a 15 MHz signal from the TT&C Modem unit. This signal is upconverted twice to the S-band frequency of 2.285 GHz at a RF power level of +33.4 dBm, and feeds a helical antenna with a peak gain of 15 dBi, illuminating the entire 31° FOV. In the receive link, the two remote ground stations each return a unique PN code, coherently related to the PN_1. These two codes, referred to herein as PN_2 and PN_3, are both received at 2.120 GHz in a receive channel with a transistor front end ($T_e \approx 350$ K). This signal is doubly downconverted to a low VHF frequency of 16.5 MHz, and sent to the TDRS/GS FDM/FM transmit channel. The S-band Order Wire data is also received in this same transponder at 2.218 GHz. This signal is also received in a parallel transistor front end receive channel, doubly downconverted to 13.5 MHz and sent to the TDRS/GS FDM/FM transmit channel.

3.1.6 WEIGHT AND POWER SUMMARY

The weight and prime power summary for the TDRS telecommunication system for the Uprated Delta 2914 Configuration is tabulated in Table 3-16. The prime power requirements are shown for the various modes of operation, where applicable, for each component of the Telecommunication System. The total system weighs 123.6 kg including the supporting booms for the antennas. The total prime power requirement varies with the numerous modes of operation of each component, and is covered in Section 2 of Volume III.

TABLE 3-16. TDRS TELECOMMUNICATION SYSTEM: WEIGHT & POWER SUMMARY

(PART II--UPRATED DELTA 2914 : LDR + MDR + HDR)

COMPONENT	WEIGHT (KG)	PRIME POWER (Watts) PEAK	AVE
1. LDR			
• Receiver	4.1	9.1	9.1
• Transmitter	2.5	403.6/205.2/106.0 (1)	106.0/56.0/31.6 (2)
• Antenna & Support Booms	14.4		
2. MDR/HDR #1			
• Receiver	4.5	8.2	8.2
• Transmitter	6.4		
*S-band		66.0/15.0 (3)	66.0/15.0 (3)
*Ku-band		35.8/5.1 (4)	35.8/5.1 (4)
• Antenna & Support Booms	17.9	24.0	4.0
3. MDR/HDR #2			
• Receiver	4.5	8.2	8.2
• Transmitter	6.4		
*S-band		66.0/15.0 (3)	66.0/15.0 (3)
*Ku-band		35.8/5.1 (4)	35.8/5.1 (4)
• Antenna & Support Booms	17.9	24.0	4.0
4. TDRS/GS			
• Receiver	2.2	5.3	5.3
• Transmitter	9.6		
-HDR + FDM/FM Channel		50.6 (5)	20.8 (6)
-FDM/FM Channel only		6.0 (5)	6.0 (5)
• Antenna & Support Booms	7.5	24.0	0.
5. Frequency Source	3.5	8.0	8.0
6. TT & C			
• Processor	4.4	10.0	10.0
• Transceiver	1.8	13.5/4.5	0.5 (7)
• Antenna (2)	1.4		
7. TDRS Tracking/Order Wire			
• Transponder	2.5	7.9	2.0 (8)
• Antenna	0.1		
8. Cabling/Wires/Waveguide	12.0		
TOTAL	123.6		

NOTES:

1. Emergency steered Beam mode provides EIRP of +42, +39, or + 36 dBw.
2. F-FOV mode provides EIRP of +30, +27, or +24 dBw at 26° FOV.
3. S-band mode emits EIRP of +47 and +41 dBw to support manned and unmanned user, respectively.
4. Ku-band mode emits EIRP of +53.6 and +23.6 dBw to support video to manned user and 1 kbps to unmanned user, respectively.
5. With 17.5 dB rain margin.
6. With 7.5 dB clear weather margin.
7. Transmitter normally turned off "on-station".
8. Transmitter turned on only periodically; receiver is always on.

3.2 SUBSYSTEM DESCRIPTION

3.2.1 LDR TRANSPONDER

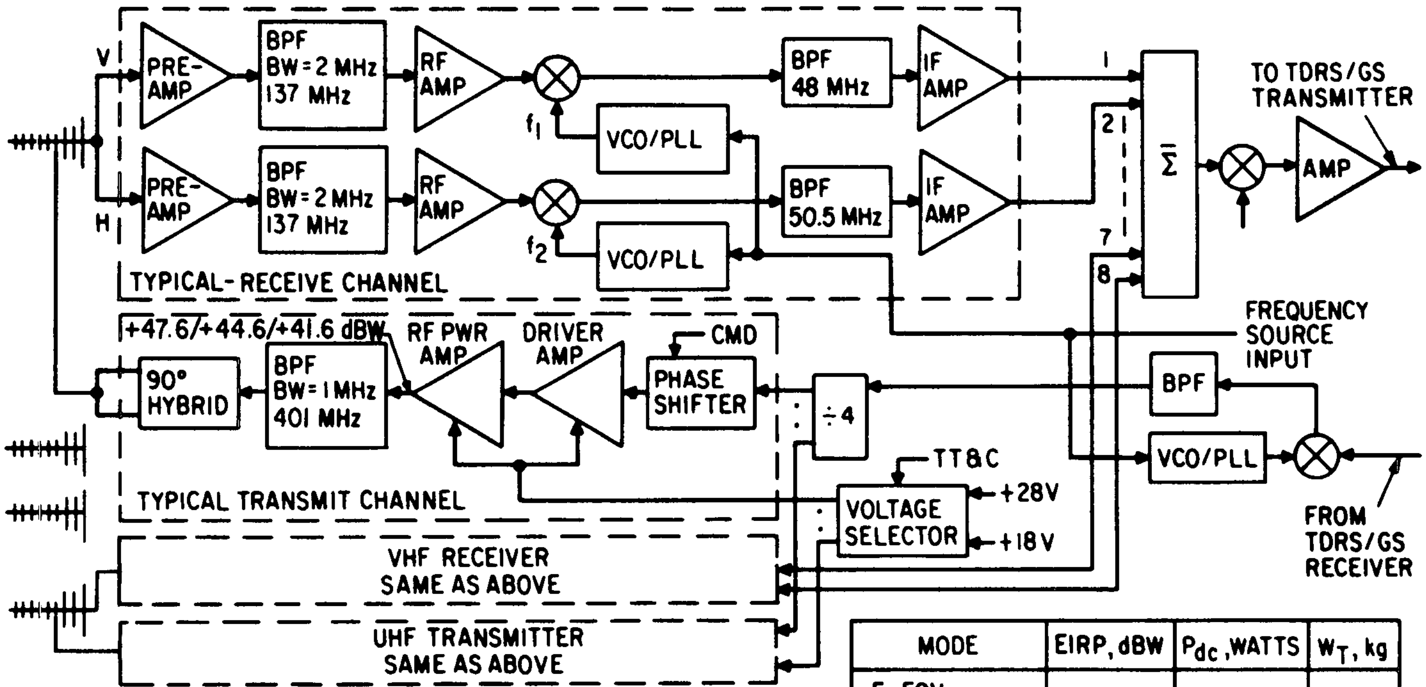
The LDR Transponder as shown in a functional block diagram in Figure 3-12 is a dual frequency (UHF-VHF) design, operating in the UHF (400.5 to 401.5 MHz) band in the forward transmit direction, and in the VHF (136 to 138 MHz) band in the return receive direction. The transponder characteristics are summarized in Table 3-17.

The UHF transmitter has four identical 1-MHz channels. In the prime F-FOV mode, it utilizes one of these channels to provide an EIRP (when combined with the 15 dBi disc-on-rod antenna element) of +30, +27 or +24 dBW. Four such identical channels provide quadruple redundancy for the F-FOV mode, or can be used in a phased array Emergency Steered Beam Mode to provide an EIRP of +42, +39 or +36 dBW.

The VHF receiver is an eight channel adaptive phased array system which utilizes adaptive spatial and polarization filtering techniques to discriminate against intentional and unintensional interference signals. This system is coined Jr. AGIPA for Adaptive Ground Implemented Phased Array. The term Jr. is used herein to describe a small aperture array ($\approx 1.1\lambda$ separation between elements) as distinguished from a Sr. AGIPA which employs a large aperture ring array with five elements on a 5λ diameter. Sr. AGIPA is used in the subsequent Atlas Centaur/Space Shuttle configuration. All of the beam steering and adaptive signal processing functions for Jr. AGIPA are conducted on the ground; thereby minimizing the size, weight, power penalty on the TDRS spacecraft. The return link can also be used in an unadapted F-FOV backup mode by using the outputs: (1) from two orthogonally polarized antenna channels, or (2) by using all eight channels with the beam pointed along the spacecraft local vertical.

3.2.1.1 LDR ANTENNA DESIGN

The LDR antenna is a quad-array of collinearly stacked UHF elements mounted on top of the VHF elements as shown in Figure 3-13. The UHF element is a surface wave disc-on-rod structure fed by independent crossed dipoles. These dipoles in turn are fed with a 90° hybrid to provide circular polarization. Each element is designed to illuminate the entire 31° FOV, and has a peak gain of 15 dBi. In the phased array Steered Beam Mode, the peak gain is 21 dBi. Other characteristics are shown in Figure 3-13.



MODE	EIRP, dBW	P_{dc}, WATTS	W_T, kg
F-FOV			
NORMAL	+30	115.1	21.0
ECLIPSE	+27 +24	65/35	
STEERED BEAM			
CONTINUOUS	+36	115.1	
BURST	+39/+42	214.3/412.7	

V72-3149

FIGURE 3-12. LDR TRANSPONDER DESIGN

TABLE 3-17. LDR TRANSPONDER CHARACTERISTICS

FORWARD LINK

- 1 - 1 MHz CHANNEL
- FREQUENCY TRANSLATION
- QUAD ARRAY OF HIGH-GAIN DISC-ON-RODS
- MODES:

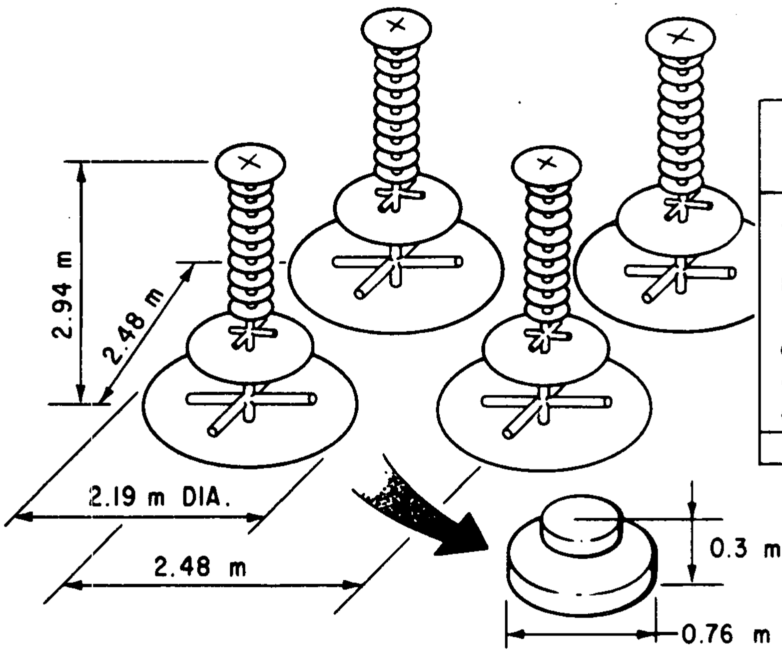
MODES:	F-FOV	EMERGENCY STEERED BEAM
GAIN AT 26°	+13.0 dBi	+19.0 dBi
EIRP AT 26°	+30, +27 OR +24 dBw	+42, +39, + 36 dBw

- QUADRUPLE REDUNDANCY

RETURN LINK

- 8 - 2 MHz CHANNELS
- FM REMODULATION
- QUAD ARRAY OF SHORT BACKFIRE
- MODES:

MODES:	JR AGIPA	F-FOV
GAIN AT 26°	+16.5 dBi	+13.3 dBi
G/T_S	-13.7 dB/K	-16.9 dB/K



A. *DEPLOYED*

B. *STOWED*

PARAMETERS	VHF		UHF	
	AGIPA	F-FOV	F-FOV	STEERED
ANTENNA TYPE	QUAD ARRAY OF SHORT BACK-FIRE ELEMENTS		QUAD ARRAY OF DISC-ON-RODS	
HPBW-ELEMENT	46°	46°	31°	31°
-ARRAY	23°	23°		9°
GAIN-PEAK	17.4 dBi	17.4 dBi	15.0 dBi	21.0 dBi
GAIN - 26° FOV	16.5 dBi	13.3 dBi	13.0 dBi	19.0 dBi
GAIN - 30° FOV	16.2 dBi	11.6 dBi	12.0 dBi	18.0 dBi
WEIGHT	14.4 kg INCLUDING BOOMS			

FIGURE 3-13. DUAL FREQUENCY (VHF-UHF) LDR QUAD ARRAY

The VHF element is a short backfire element fed by independent crossed dipoles, and provides a peak element gain of 11.4 dBi or peak array gain of 17.4 dBi. Other array characteristics are summarized in Figure 3-13.

These VHF short backfire characteristics are based on empirical test programs conducted initially by L. R. Dod[1,2] and more recently for the described VHF-UHF collinearly stacked configuration by L. Deerkoski, both of NASA-GSFC. In the collinearly stacked configuration, the UHF elements were mockup, only.

Each collinearly stacked VHF-UHF element is stowed in a compact package as shown in Figure 3-13. The surface wave structures are erected with a simple stepper motor design to the fully deployed configuration as described in detail in the Spacecraft Design section.

The total dual frequency LDR quad-array including the supporting boom structure weighs 14.4 kg.

3.2.1.2 LDR TRANSCEIVER

The receiving portion of this transceiver consists of eight almost identical superheterodyne channels that receive signals at 137 MHz and downconvert them to individual IF's with center frequencies from 48 to 65.5 MHz. These signals are combined, downconverted as a group to the 19.5 to 37 MHz region, and sent to the TDRS/GS Transmitter for multiplexing with other signals before transmission to the Ground Station.

The LDR transmitter divider accepts the IF output of the TDRS/GS receiver, frequency source reference outputs and beam steering data. It provides four output signals at 401 MHz phase locked to the reference, with the proper phase shifts between the output signals to form a steerable antenna pattern when amplified and applied to the UHF antenna array. Amplification takes place in a solid-state UHF transmitter. Four such units are provided, but only one is used for normal mode operation. All four units are used only in the Burst mode which requires +23 dBW (200 watts) of UHF power to be developed. The transmitter is straightforward in design.

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1. L. R. Dod, Experimental Measurements of the Short Backfire Antenna, NASA-GSFC X-525-66-480, October 1966.
2. L. R. Dod, Backfire Yagi Antenna Measurements, NASA-GSFC X-525-67-604, December 1967.

The output of each transmitter is split in a hybrid junction to feed the horizontal and vertical antenna feeds.

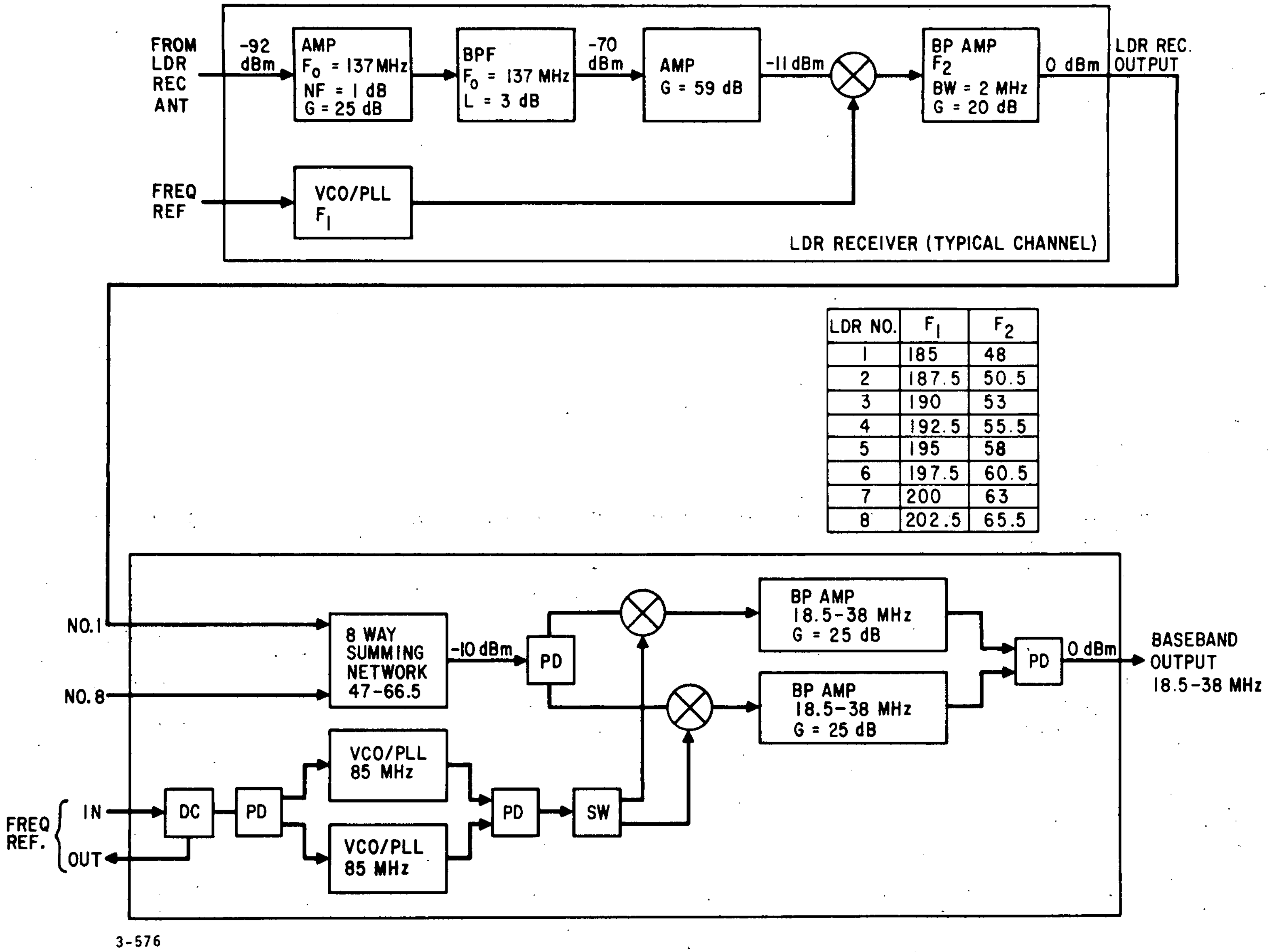
Redundancy in the receiver is not provided because of the unique nature of the AGIPA concept. AGIPA redundancy is inherent in that the eight channels are handling essentially the same information. Loss of a channel merely reduces the resolution achievable in beam forming and constitutes graceful degradation rather than catastrophic failure. The LDR summing network and second downconversion, common to all eight channels, will be 100 percent redundant. Similarly, the transmitter will not have redundancy for each of its four channels. Since only one transmitter is normally on at a time, quadruple redundancy is inherent. In the Burst mode, failure of one transmitter again does not cause catastrophic failure but gradual reduction in power output.

3.2.1.2.1 LDR RECEIVER AND IF SUMMING NETWORK

One channel of the LDR receiver eight-channel superheterodyne receiver is shown in Figure 3-14. Eight individual signals received at 137 MHz are downconverted to individual IF's ranging between 48 and 65.5 MHz. The overall IF spectrum of 19.5 MHz is then downconverted to a baseband spectrum of 18.5 to 38 MHz. This spectrum then is used to modulate the TDRS/GS transmitter and is shown in Figure 3-25.

Signals are received at a power level of -92 dBm at the input to the RF preamplifier. The design features a two-stage device using discrete transistors such as the 2N5650. The amplifier will be tuned to 137 MHz to optimize noise figure (1 dB) and gain (25 dB) as well as relieve the rejection requirement of the band-pass filter following the preamp. This design provides an overall device noise figure of approximately 1.0 dB with a total gain of 25 dB. The effective system noise figure is less than 2.0 dB, including second stage N.F. contributions and cable losses to the antenna. The maximum preamp output level is -55 dBm on 4σ RFI peaks. The equivalent receiver noise input level is -109 dBm. A low-pass filter is employed to keep the LDR transmitter power leak through in the linear region.

The signal is then (137 MHz) band-pass filtered to eliminate out-of-band signals and noise. Since the RF preamp is well within its linear region for all signals and noise, the filtering is performed after the preamp without loss of effectiveness. This configuration permits higher filter loss without sacrifice of noise figure and also simplifies the filter design with corresponding savings in size and weight. A six-pole lumped constant filter design having a linear phase characteristic and approximately a 3-dB insertion loss is used. Its bandwidth together with the other filters will produce a total link bandwidth of 2.0 MHz.



LDR NO.	F_1	F_2
1	185	48
2	187.5	50.5
3	190	53
4	192.5	55.5
5	195	58
6	197.5	60.5
7	200	63
8	202.5	65.5

FIGURE 3-14. LDR JR. AGIPA RECEIVER BLOCK DIAGRAM

The RF amplifier will consist of two MIC amplifiers and deliver most of the overall receiver gain. The 1-dB compression point is approximately +2 dBm (maximum RFI level) which minimizes intermodulation products. The noise figure of this amplifier is not critical since system noise is established by the RF preamp.

The mixer will be a double-balanced plug-in type suitable for integrated operation in hybrid circuits. The mixer operates with +10 dBm LO drive to minimize conversion loss, signal compression, and intermodulation distortion on RFI levels up to +2 dBm. Each mixer in the eight channels operates with a different LO frequency to produce different IF's for each channel.

From the mixer the signal is fed to the band-pass filter. Each channel will have a separate IF ranging between 48 and 65.5 MHz, a range selected for minimum filter size. This, in effect, takes the eight channels of information and places them into a spectrum 19.5 MHz wide. The channels are 2 MHz wide with 2.5 MHz center frequency separation. Each filter exhibits 6-dB insertion loss and provides 30-dB rejection at adjacent channel center frequencies. Lumped constant designs are used with components mounted directly on the same printed circuit board as the rest of the receiver. Compartmentalized construction is used to assure adequate RF isolation.

After filtering, the eight IF signals are amplified and summed together as shown in Figure 3-14. The function of this circuit is to combine the individual IF spectrums into a single output. Each IF signal receives a single stage of amplification (15 dB) and is resistively added into a common load. After summing, the IF signals are fed to the second mixer.

The combined IF spectrum is downconverted to a baseband of 18.5 to 38 MHz. This is achieved by mixing with a 85.0 MHz LO at +10 dBm producing the desired spectrum while maintaining linearity. After downconverting to baseband the signal is filtered and amplified. The gain is sufficient to bring the RFI level to approximately -10 dBm before being fed to the TDRS/GS transmitter.

Since the receiver output level can change 26 dB, the last amplifier provides 26 dB of AGC range. The actual device gain is controlled from +14 to -12 dB using pin diode attenuators. The amplifier is a linear integrated type with a 1-dB compression point of about +5 dBm. The second IF band-pass filter is lumped constant to reduce out-of-band noise. This portion of the LDR receiver is common to all eight receiver channels, and in order to maintain high reliability a redundant mixer amplifier and filter is used.

3.2.1.2.2 LDR TRANSMITTER

This transmitter is comprised of a divider unit and four power amplifiers. It is identical to that of the Phase One configuration except that only one frequency is transmitted and each of the four transmitters have three output power levels available. These levels are +41, +44, and +47 dBm. A block diagram of the transmitter and its divider unit is shown in Figure 3-15. For a complete description refer to the Part I Final Report.

3.2.2 MDR/HDR TRANSPONDER

Both MDR/HDR #1 and MDR/HDR #2 are nearly identical as shown in the simplified block diagram in Figure 3-16. The details and the differences between the two transponders is described in the subsequent section 3.2.2.2. Both transponders are dual frequency, S-band to service the MDR users, and K_u-band to service the HDR user. S- and K_u-band receive channels are on simultaneously to support a common MDR/HDR user; however, only S- or K_u-band can be used on transmit. Each transponder uses a 3.8 meter deployable antenna. Transponder characteristics are summarized in Table 3-18.

3.2.2.1 MDR/HDR ANTENNA

The MDR/HDR Antenna is a dual frequency S-/K_u-band 3.8 meter deployable parabolic reflector as illustrated in Figure 3-17, and is of the type currently in development by Radiation, Inc. under NASA Advanced Aerospace Flight Experiment (AAFE) Development Program. This design uses an umbrella type supporting structure with 12 ribs to support a reflector surface. Due to the high surface tolerance requirements for K_u-band operation the double mesh design is utilized for the reflector. The double mesh technique utilizes two mesh surfaces which are separated by the rib thickness and connected to one another by tensioned metallic ties. The front reflector surface is contoured to a precision parabolic shape by properly tensioning the tie wires. This technique essentially eliminates surface tolerance dependency on the number of ribs. The entire structure stows into a compact package for launch.

The dual frequency feed is shown in Figure 3-18; and employs a prime focus feed for S-band, and a cassegrain four-horn pseudo-monopulse feed for K_u-band operation. The K_u-band cassegrain uses a frequency sensitive subreflector (FSS) which is highly reflective at K_u-band, but transparent at S-band.

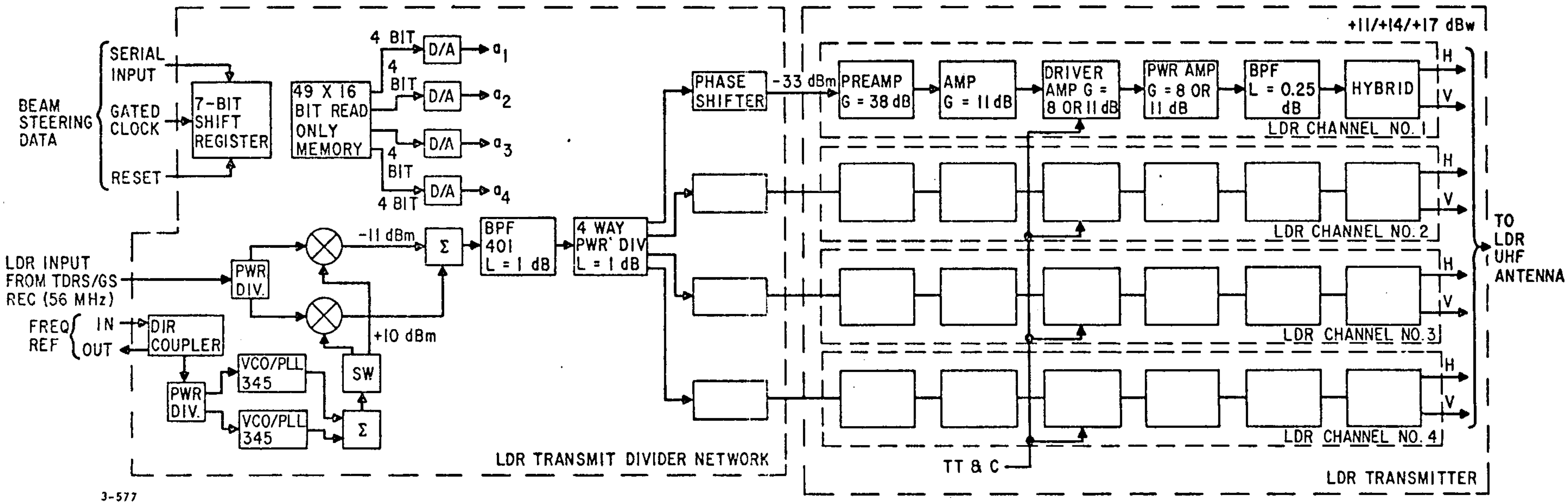
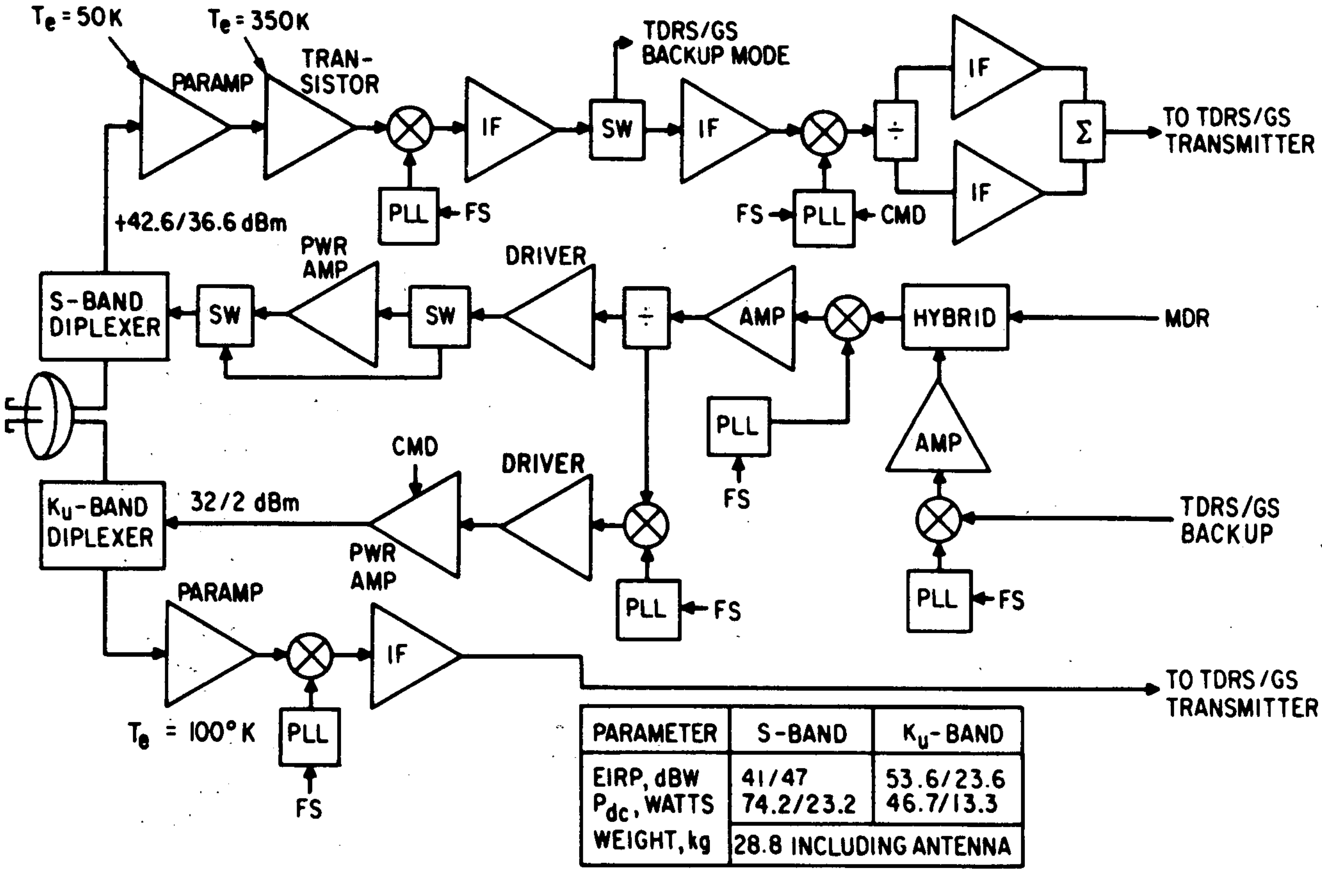


FIGURE 3-15. LDR UHF TRANSMITTER BLOCK DIAGRAM

FIGURE 3-15



PARAMETER	S-BAND	K_u-BAND
EIRP, dBW	41/47	53.6/23.6
P_{dc}, WATTS	74.2/23.2	46.7/13.3
WEIGHT, kg	28.8 INCLUDING ANTENNA	

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FIGURE 3-16. MDR/HDR TRANSPONDER BLOCK DIAGRAM

TABLE 3-18. MDR/HDR TRANSPONDER CHARACTERISTICS

	S-BAND		K_U - BAND
FORWARD LINK			
• FREQUENCY	S-BAND	OR	K_U - BAND
• RF BANDWIDTH	10 OR 32 MHz TUNABLE OVER 75 MHz BAND		100 MHz FIXED
• EIRP			
UNMANNED	+41 dBW		23.6 dBW
MANNED	+47 dBW		53.6 dBW
• REDUNDANCY	100% (ACTIVE COMPONENTS)		100% (ACTIVE COMPONENTS)
RETURN LINK			
• FREQUENCY	S-BAND	AND	K_U - BAND (SIMULTANEOUSLY)
• RF BANDWIDTH	10 MHz TUNABLE OVER 100 MHz		150 MHz
• G/T_S	+10.0 dB/K		+25.9 dB/K
• REDUNDANCY	100% (ACTIVE COMPONENTS)		100% (ACTIVE COMPONENTS)
BACKUP MODE (MDR/HDR NO. 1 AND NO. 2)			
• FREQUENCY	S-BAND	OR	K_U - BAND
• MODE	FDM		FDM
• COMPONENTS	ANTENNA AND TRANSCEIVER		ANTENNA AND/OR TRANSCEIVER (MDR/HDR NO. 1 ONLY)

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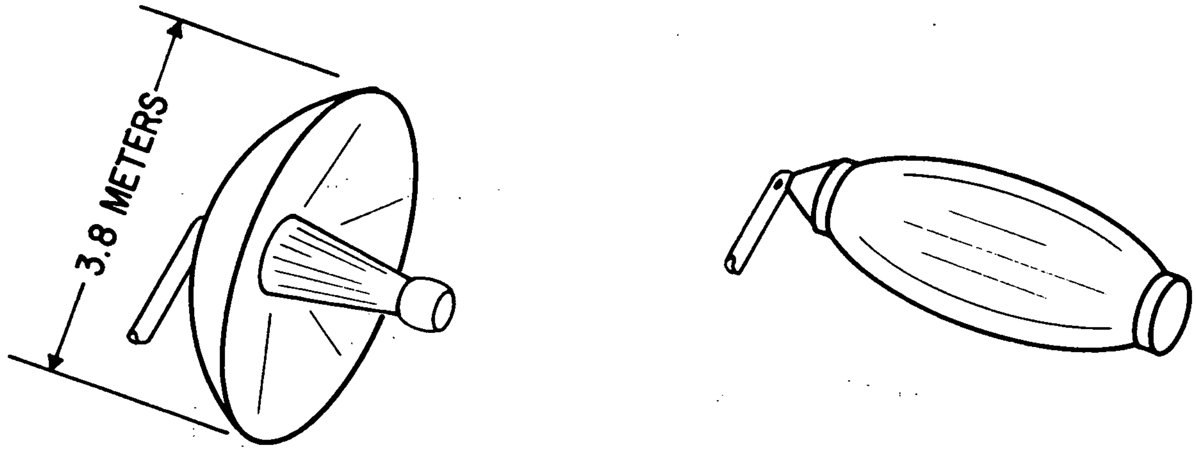

A. DEPLOYED

B. STOWED

PARAMETER	S-BAND	Ku-BAND
• REFLECTOR TYPE	• DEPLOYABLE DOUBLE MESH	
• FEED DESIGN	• HELIX AT FOCAL POINT	• 4 HORN PSEUDOMONOPULSE CASSEGRAIN FEED
• HPBW	≈ 2.6°	≈ 0.4°
• GAIN – FORWARD – RETURN	35.5 dBi 36.2 dBi	53.6 dBi 52.8 dBi
• WEIGHT	17.9 kg INCLUDING SUPPORTING BOOMS	

FIGURE 3-17. MDR/HDR DUAL FREQUENCY ANTENNA CHARACTERISTICS

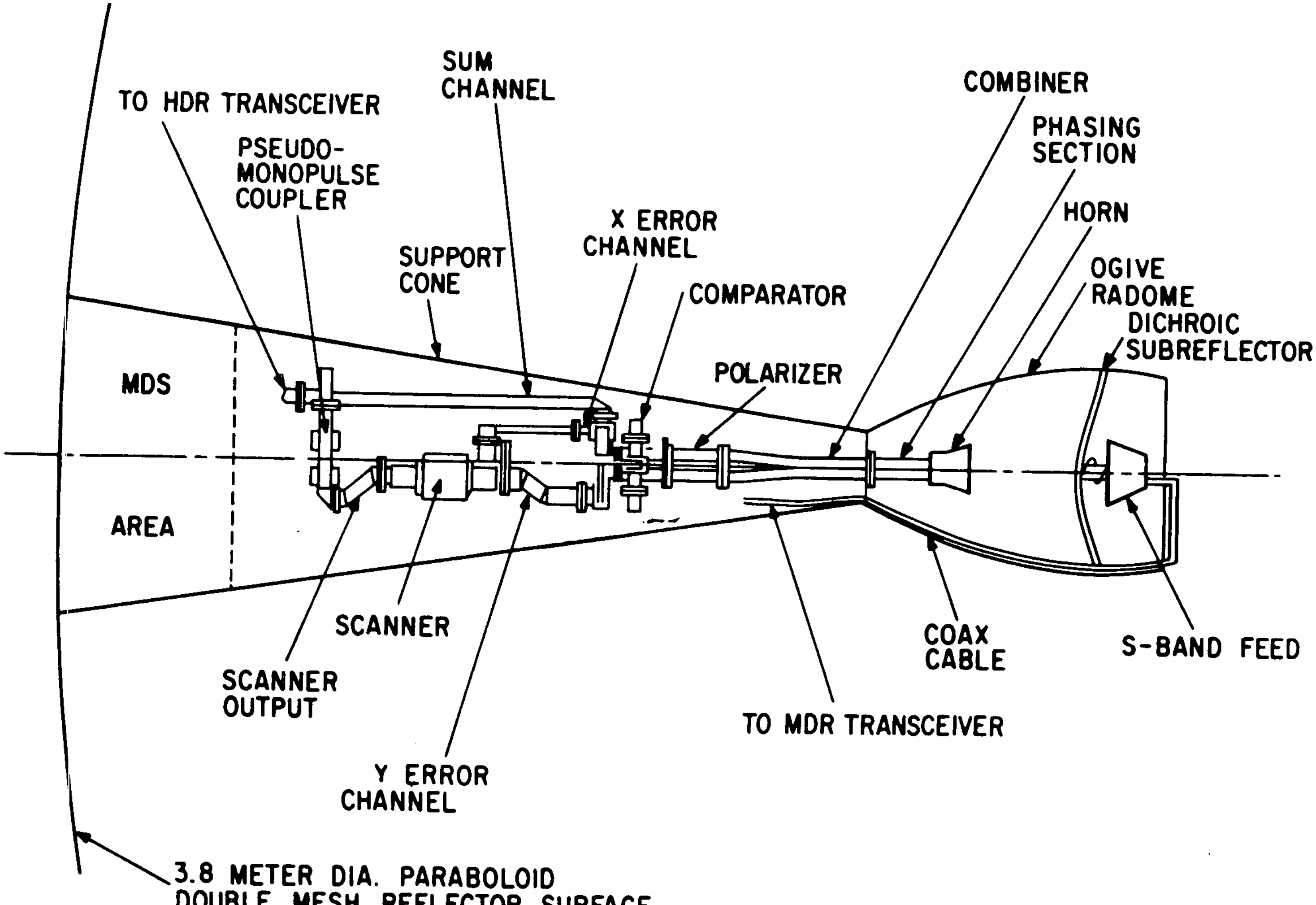

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FIGURE 3-18. MDR/HDR DUAL FREQUENCY FEED

Table 3-19 summarizes the antenna efficiency calculation for the S- and K_u-band modes; and is used to compute the transmit and receive gains for each frequency as tabulated in Table 3-20.

The weight of the dual frequency MDR/HDR antenna is summarized in Table 3-21. The total antenna system including the gimbal mount, monopulse drive electronics, and support booms weighs 17.9 kg.

3.2.2.2 MDR/HDR TRANSCEIVER

The MDR/HDR Transceiver handles transmissions at S- and K_u-band between the TDRS and user spacecrafts. Reception can occur simultaneously in each band but transmission occurs in one band at a time. Two transceivers are provided to independently service two users. MDR operation occurs at S-band and HDR at K_u-band. In addition to this primary mode of operation, either transceiver can function as a backup for the TDRS/GS transceiver. 100-percent redundancy is provided in each unit.

3.2.2.2.1 HDR RECEIVER, PRIMARY MODE

The HDR receiver uses a parametric amplifier front end whose noise temperature is 100 K. Such low noise devices permit maximum performance on low-level signals. Reception is at a frequency of 13,925 MHz with a bandwidth of 150 MHz. Single conversion is used with the IF at 500 MHz. This frequency is low enough to permit low noise transistors to be used and high enough to provide a wide separation to the RF image response, easing the front-end design. The use of single conversion is justified because the data rate of up to 100 Mbps requires a wide bandwidth of 150 MHz.

The antenna output signal (see Figure 3-19) is fed through a diplexer to a low-loss waveguide switch. The position of the switch, selected at the ground station, determines whether the primary or redundant system is connected to the antenna. The diplexer permits the antenna to be accessed simultaneously by the receiver and the transmitter. The signal at -78 dBm is amplified by a paramp to a level of -63 dB and fed to the mixer where it is downconverted to 500 MHz. The local oscillator used is phase locked to the Frequency Source. Redundancy is also provided in the LO and further flexibility achieved by cross strapping the output so either mixer can be operated from either LO. The mixer conversion loss is kept low and spurious responses reduced by employing double balanced mixers and operating at +10 dBm LO level. Following the mixer is a low noise 500 MHz preamplifier that provides 20 dB of gain and drives a band-pass filter that determines the receiver spectral shape. The main IF amplifier has a gain of 64 dB with an output capability of +8 dBm. Its output feeds a

TABLE 3-19. DUAL FREQUENCY MDR/HDR
ANTENNA EFFICIENCY BUDGET

	Parameter	S-Band	K_u-Band RCV	K_u-Band XMIT
A.	Illumination Efficiency			
	• Spillover/Amplitude Taper	65.0	85.0	85.0
	• Primary Phase	97.0	99.0	99.0
	• Blockage	95.7	98.1	98.1
	• Primary Cross Polarization	99.8	99.0	99.0
	• Secondary Cross Polarization	97.8	99.9	99.9
	• Subreflector	98.0	94.0	94.0
	• Radome	98.0	96.5	96.5
	Sub-Total	56.5	74.0	74.0
B.	Reflector Efficiency			
	• Surface Tolerance (0.018)	99.6	93.0	92.0
	• Mesh Reflectivity	99.5	99.0	99.0
	Sub-Total	99.1	92.0	91.0
C.	Components Efficiency			
	• S-Band Components/Cable (0.24 dB)	94.5		
	• K_u-Band Horn, Phasing Polarizer/Combiner		97.8	97.8
	• Diplexer		98.5	98.5
	• Power Divider		---	95.5
	• Comparator		93.5	---
	Sub-Total	94.5	0.90	0.919
	Total Overall Efficiency (AXBXC)	52.9	61.3	62.0

TABLE 3-20. SUMMARY OF MDR/HDR ANTENNA GAIN

Mode		Transmit	Receive
S-Band			
• Frequency,	GHz	2.025 to 2.100	2.200 to 2.300
• Gain,	dBi	35.5	36.2
K_u-Band			
• Frequency,	GHz	14.810 to 14.910	13.85 to 14.0
• Gain,	dBi	53.6	52.8

hybrid junction whose other input is fed from the redundant channel. The hybrid output is split into two paths, one driving the pseudo-monopulse antenna steering circuits, the other going to the TDRS/GS transmitter for transmission to the Ground Station. The other components of the HDR receiver are utilized for the TDRS/GS Backup Mode and will be described in a subsequent subsection.

3.2.2.2.2 MDR RECEIVER, PRIMARY MODE

The front end of this receiver uses a parametric amplifier whose noise temperature is 50 K and whose gain is 15 dB. Reception is in S-band between 2200 and 2300 MHz. Following the paramp is an S-band transistor amplifier whose low noise and high gain assure the lowest possible overall noise figure for this receiver. A double conversion receiver is employed with the first IF at 500 MHz, permitting commonality with the HDR IF amplifiers. The second conversion brings the signal to a 10 MHz band centered either at 6 or 45 MHz depending upon the frequency of the second LO selected. Broadband operation (100 MHz) is maintained past the first mixer to permit the receiver to function as a backup for the GS/TDRS Receiver. Subsequently, the bandwidth is narrowed to the required 10 MHz.

Referring to Figure 3-19, the antenna signal is fed through a diplexer to a switch which determines whether the primary or redundant system is to be used. The diplexer provides sufficient isolation between the Transmitter (2025 to 2100 MHz) and Receiver (2200 to 2300 MHz) to prevent saturation of the receiver. Another potential source of receiver

TABLE 3-21. DUAL FREQUENCY MDR/HDR
ANTENNA WEIGHT BUDGET

Component	Weight (kg)
• Reflective Surface	3.68
• Deployment System	0.95
• Restraint System	0.77
• Feed Support Structure	1.50
• S-Band Feed, Components, Cable	1.05
• K_u-Band Feed, Components, Cable	0.95
• Scanner	1.13
• Coupler	0.11
• S-/K_u-Band Rotary Joint (2)	0.68
• K_u-Band Rotary Joint (2)	0.45
• Gimbal Mount	1.48
• Drive Electronics	2.27
• Antenna Boom Support	2.90
Total	17.9

saturation is the TDRS Tracking signal at 2285 MHz. To prevent this signal from being space-coupled between antennas and saturating the MDR Receiver, the diplexer contains a notch at this frequency. From the switch a low noise paramp providing 15 dB gain amplifies the signal and feeds the low noise transistor amplifier. The resulting signal is fed to the first mixer together with a local oscillator at +10 dBm level. The frequency of the LO is controlled by ground command to any frequency between 1705 and 1795 MHz in 5-MHz steps. The selected frequency corresponds to that transmitted by

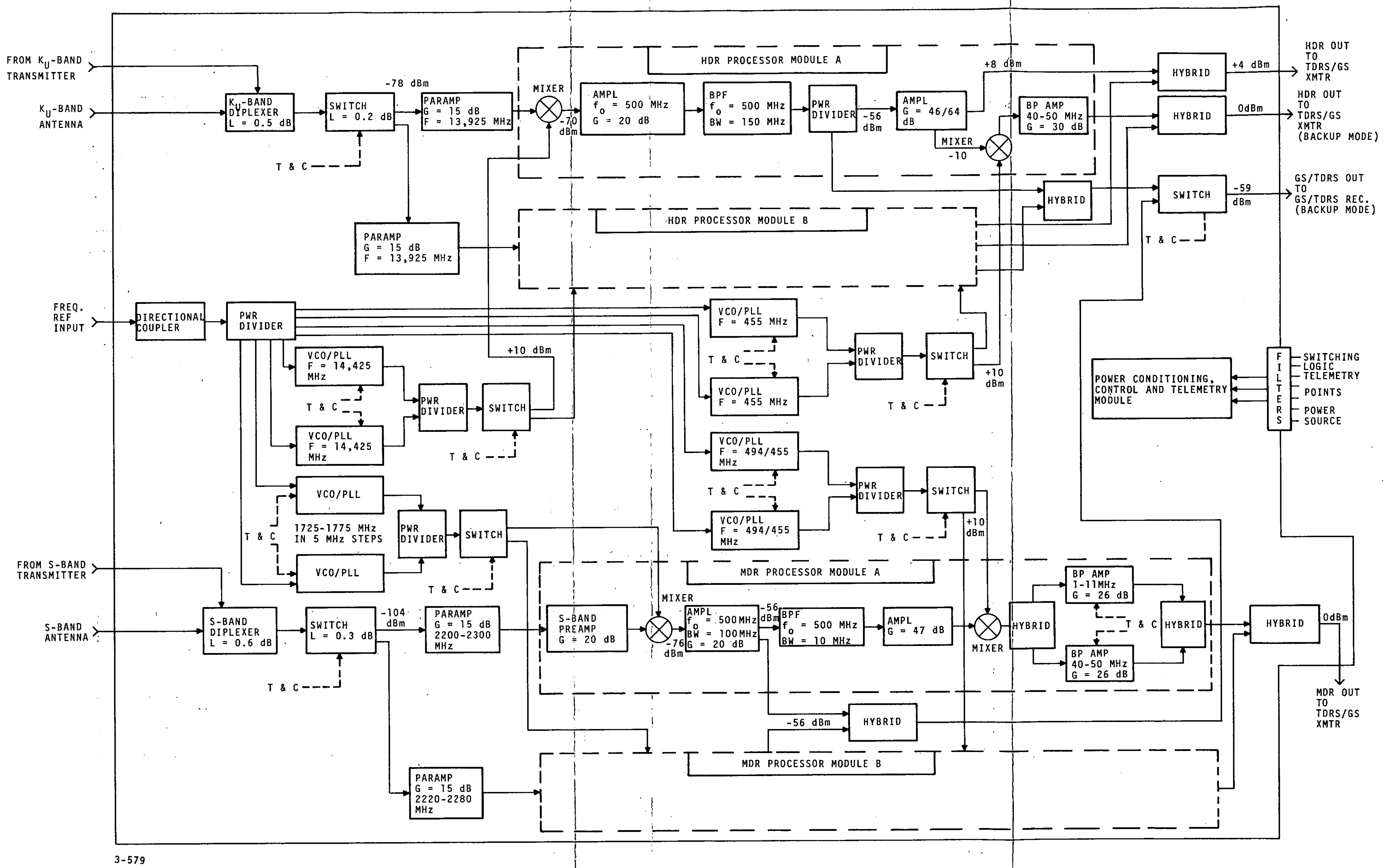


FIGURE 3-19. MDR/HDR RECEIVER BLOCK DIAGRAM

FIGURE 3-19

the MDR user being serviced. Following the mixer is a 100-MHz bandwidth 20-dB gain IF amplifier that provides the output signal for the GS/TDRS backup mode and drives a narrowband (10 MHz) 500 MHz filter that efines the receiver shape. The band-pass filter is immediately followed by a 47 dB gain IF amplifier which drives the second mixer. The second LO is ground selectable to either 494 or 455 MHz and converts the first IF signal to either 6 or 45 MHz. This provides an additional degree of flexibility in that either MDR receiver can supply its output signal at either of these two frequencies. A hybrid junction following the mixer splits its input signal into two paths, one containing a 6-MHz band-pass amplifier, the other containing a 45 MHz band-pass amplifier. Only one of these amplifiers is energized at a time. The selected one corresponds to the second LO used. A second broadband hybrid combines the outputs of the latter amplifiers to a single line and feeds a third hybrid fed also from the redundant channel. Hybrids are used for splitting and combining instead of switches because they are passive and more reliable. Each hybrid is characterized by a 0.5 dB insertion above the 3-dB split inherent in the device which must be made up by amplifier gain.

3.2.2.2.3 MDR/HDR TRANSMITTERS

The MDR/HDR Transmitter accepts an input at 120 MHz (MDR #1) or 240 MHz (MDR #2), translates it to either S- or K_u-band, and amplifies the resulting signal for transmission to the MDR/HDR user. The bandwidth available is 75 MHz in S-band and 100 MHz in K_u-band. In S-band, the frequency range between 2025 and 2100 MHz is used; in K_u-band, the frequency range is from 14,810 to 14,910 MHz. The instantaneous signal bandwidth of 32 MHz can be positioned anywhere within either transmission band by adjustment by the Ground Station transmitted frequency. All frequency translations within the MDR/HDR Transmitter are accomplished with phase-locked oscillators resulting in a completely coherent transmitter. Provision is made to permit either band to function as a backup to the TDRS/GS transmitter.

3.2.2.2.4 MDR/HDR #1 TRANSMITTER

Figure 3-20 is a detailed block diagram of the MDR/HDR #1 Transmitter. The input signal at 120 MHz is applied to a hybrid that drives both the primary and redundant mixers used to upconvert the input to a frequency in S-band. A phase locked local oscillator is used whose frequency is 1942.5 MHz. Redundancy is provided in the LO and further flexibility achieved by cross-strapping the output so either LO can drive either mixer. Following the mixer, a 100-MHz bandwidth filter selects the upper sideband at 2062.5 MHz and feeds a 17 dB gain solid-state amplifier. A 3 dB power divider splits the amplified signal into two 0 dBm signals, one each to drive the S- and K_u-band amplifiers.

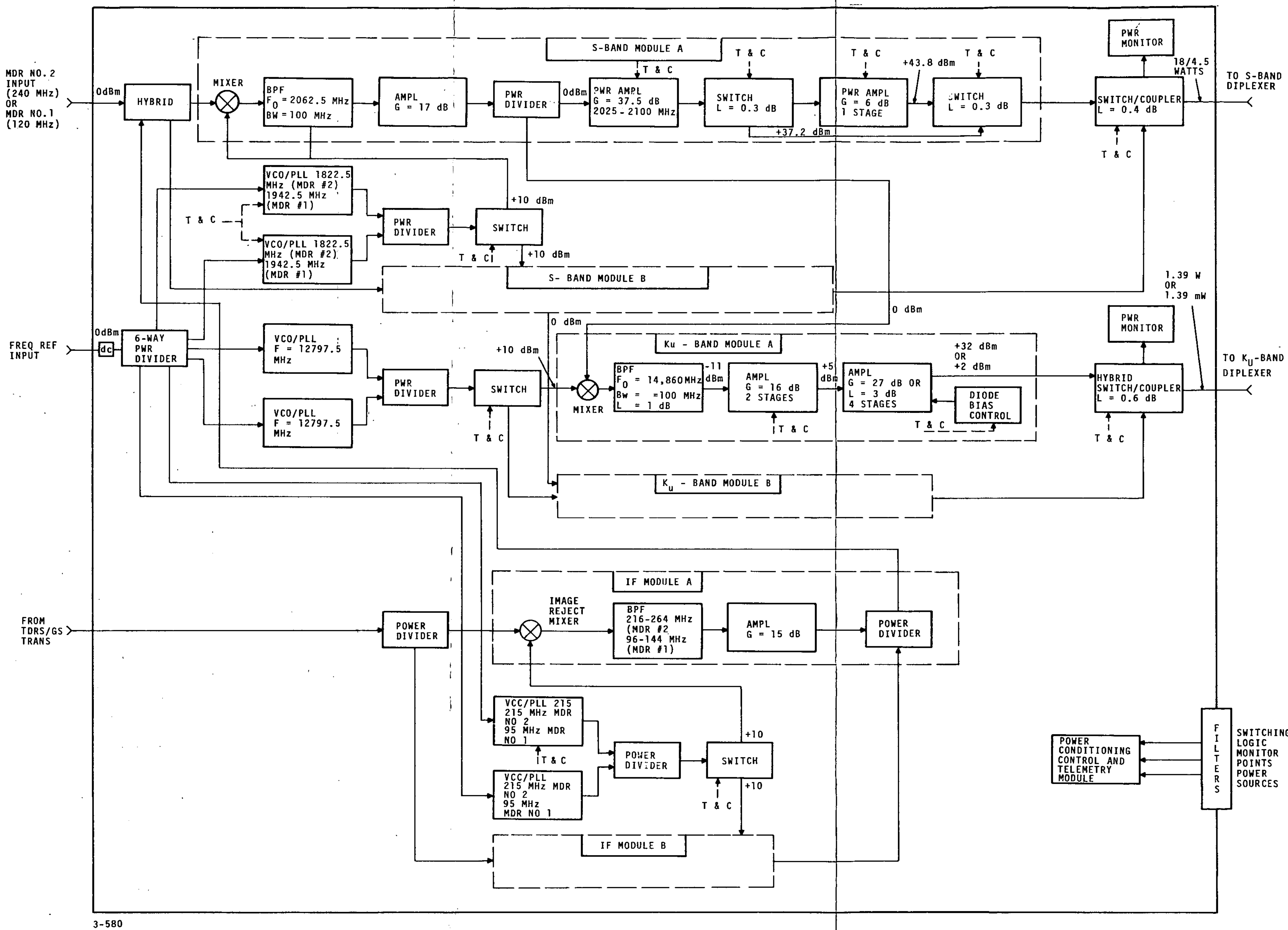


FIGURE 3-20. MDR/HDR TRANSMITTER BLOCK DIAGRAM

FIGURE 3-20

In S-band, power levels of 5.6 and 21 watts can be developed which when combined with the antenna, provide an EIRP of +41 and +47 dBW, respectively. The higher power level is used to support a manned user. The low-level preamplifier stages are available today as complete integrated assemblies suitable for direct incorporation into strip line. The high-level power stages are constructed from discrete components and are configured to be class C amplifiers to obtain maximum efficiency.

The K_u-band module accepts its 2062.5 MHz input from the S-band module and upconverts it to K_u-band (14,860 MHz) with an internally generated local oscillator phase locked to the reference frequency. A band-pass filter selects the upper sideband from the mixer and drives a six-stage circulator-coupled avalanche diode amplifier with an overall gain of 43 dB. The gain of the last four stages is controlled by a diode bias network to provide either 27 dB gain or a loss of 3 dB. The radiated power is then 1 watt or 1 milliwatt. The lower power mode is available for eclipse operation when power conservation is mandatory. This amplifier is constructed from strip line with all stages utilizing a common substrate.

3.2.2.2.5 MDR/HDR #2 TRANSMITTER

The above description of the MDR/HDR #1 Transmitter also applies to #2 Transmitter with the exception that the input signal is at 240 MHz instead of 120 MHz and therefore requires a local oscillator whose frequency is 1822.5 MHz.

3.2.2.2.6 MDR/HDR AS BACKUP TO TDRS/GS TRANSCEIVER

In the following description, the MDR/HDR #1 Transceiver is presumed to service a user and the MDR/HDR #2 Transceiver to serve as the TDRS/GS Transceiver. This latter link is capable of operation at either S- or K_u-band (with suitable Ground Station equipment) although K_u-band operation is described. It is noteworthy that complete interchangeability exists between MDR/HDR Transceivers so either one can assume either role. In the forward link, the K_u-band signal received by MDR/HDR #2 Receiver is downconverted to 500 MHz (see Figure 3-18). The output of the HDR Processor Module is at -56 dBm and feeds a hybrid junction whose other input is from the redundant module. The hybrid output is fed to a switch that selects either the HDR Processor Module output or a corresponding output from the MDR Processor Module (used for S-band operation). The switch output at a level of -59 dBm is routed to the TDRS/GS Receiver where demultiplexing occurs. In this receiver, a second downconversion takes place in the IF Module. The corresponding LO frequency is changed from 670 to 600 MHz to compensate for the narrower transmitted bandwidth and fixed frequency filters following the mixer. From the demultiplexer, MDR #1 output (120 MHz) is sent to the MDR/HDR #1 Transmitter for normal forward direction transmission at either S- and K_u-bands.

In the return link, both the user-to-TDRS and TDRS-to-Ground Station is considered. The MDR/HDR #1 Receiver, operating simultaneously at S- and K_u-bands, receives MDR data and HDR data, respectively. MDR data is downconverted to the 1 to 11 MHz slot and HDR data is downconverted to the 40 to 50 MHz slot (see Figure 3-10). Both signals are sent to the Multiplexing section of the TDRS/GS Transmitter. After being combined here with the other Return Link signals, it is routed to the MDR/HDR #2 Transmitter for transmission to the ground. This signal occupies the frequency band from 1 to 50 MHz (see Figure 3-10) and must be upconverted to 240 MHz to be compatible with the normal MDR/HDR #2 Transmitter input. It is seen in Figure 3-19 that this is accomplished with an image reject mixer and LO at 215 MHz. The image reject mixer is required because a band-pass filter will not provide sufficient rejection to the unwanted sideband. A 240-MHz amplifier increases the signal to the proper insertion level for the S-band module. This level is 10 dB below the normal mode signal to insure amplifier linearity and tolerable intermodulation distortion. The balance of the MDR/HDR #2 Transmitter operates as previously described.

3.2.3 TDRS/GS TRANSPONDER

The TDRS/GS Transponder as shown in the simplified diagram of Figure 3-21 provides the interface with the Ground Station, and operates in the 13.4 to 13.64 GHz and 14.6 to 15.2 GHz K_u-band range for the forward and return links, respectively. Its characteristics are summarized in Table 3-22.

The return link transmitter is a dual channel system, viz. FDM/FM and HDR channels. The FDM/FM channel combines the 8 LDR, 2 MDR, TT&C, TDRS Tracking and Order Wire input data at a low UHF frequency in the Combiner and Pre-emphasis Weighting Network and frequency modulates the combined FDM spectrum. The HDR channel selects either the HDR from MDR/HDR #1 or MDR/HDR #2, upconverts it to K_u-band and amplifies the signal in a TWT amplifier.

The forward link data is a 240 MHz FDM channel, consisting of LDR, 2 MDR/HDR, TT&C, and pilot reference signal as shown in Figure 3-10.

The antenna is a 1.8 meter fixed (fully erected) antenna with a double mesh reflector surface as in the MDR/HDR 3.8 meter antenna. The antenna has auto track circuits and 2 axis gimbal mounts for acquisition and tracking of the ground station.

3.2.3.1 TDRS/GS ANTENNA

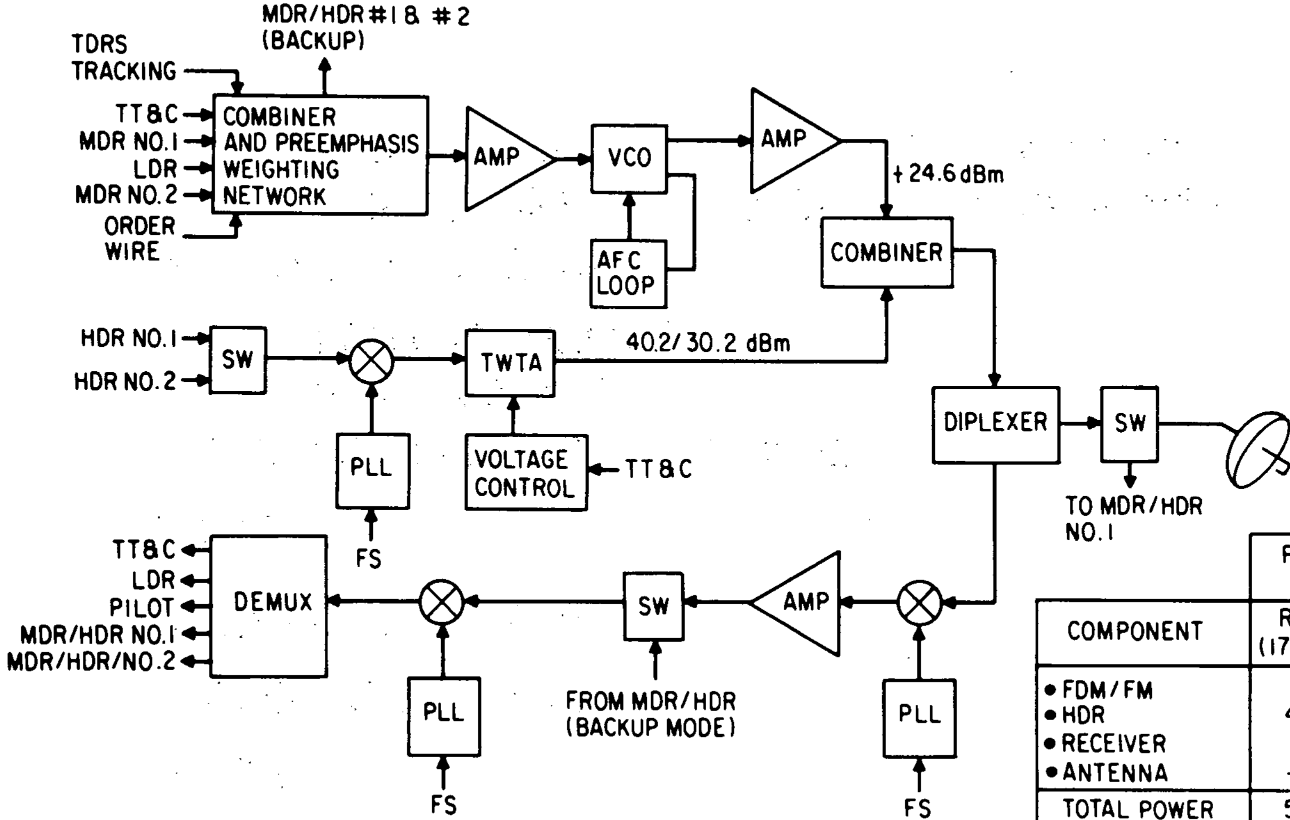
The TDRS/GS Antenna is a fully erected 1.8 meter parabolic reflector with a double mesh surface as described previously for the MDR/HDR antenna. The antenna is a Cassegrain design which employs a 4 horn monopulse feed for auto-tracking in the ground station. Its characteristics are summarized in Figure 3-22.

Table 3-23 tabulates the antenna efficiency budget, as well as the resultant antenna gains for the transmit and receive frequency bands.

The antenna weight summary is shown in Table 3-24.

3.2.3.2 TRANSCEIVER

The TDRS/GS Transceiver provides all communications between the TDRS spacecraft and the ground station. Two way communication is accomplished in K_u-band with a repeater that uses both remodulation and simple frequency translation techniques. To achieve the required reliability



COMPONENT	PRIME POWER, WATTS RAIN (17.5dB)	PRIME POWER, WATTS CLEAR (7.5dB)
• FDM/FM	6.0	6.0
• HDR	44.6	14.8
• RECEIVER	5.3	5.3
• ANTENNA	---	---
TOTAL POWER	55.9	26.1
TOTAL WEIGHT, kg	17.5	

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FIGURE 3-21. TDRS/GS TRANSPONDER BLOCK DIAGRAM

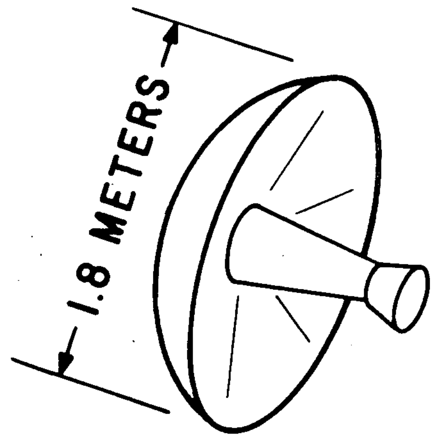
TABLE 3-22. TDRS/GS TRANSPONDER CHARACTERISTICS

FORWARD

- FDM
- BANDWIDTH = 240 MHz
- MIXER FRONT END (T_S = 33.6 dB)

RETURN

- CHANNEL (FDM/FM/FDM) BANDWIDTH
  - HDR: 150 MHz
  - FDM/FM: 209 MHz
- MODES OF SERVICE
  - HDR CHANNEL = 17.5 OR 7.5 dB RAIN MARGIN
  - FDM/FM CHANNEL = 17.5 dB RAIN MARGIN



PARAMETER	CHARACTERISTICS
● REFLECTOR TYPE	● FIXED DOUBLE MESH
● FEED TYPE	● 4-HORN PSEUDOMONOPULSE CASSEGRAIN
● HPBW	● ≈ 0.8°
● GAIN - FORWARD - RETURN	● 46.4 dBi ● 47.1 dBi
● WEIGHT	● 7.5 kg WITH BOOMS

FIGURE 3-22. TDRS/GS ANTENNA CHARACTERISTICS

TABLE 3-23. TDRS/GS ANTENNA EFFICIENCY AND GAIN BUDGET

Parameter	Receive	Transmit
A. Illumination Efficiency		
• Spillover/Amplitude Taper	85.0	85.0
• Primary Phase	99.0	99.0
• Blockage	98.1	98.1
• Primary Cross-Polarization	99.0	99.0
• Radome	96.5	96.5
Sub-Total	78.4	78.4
B. Reflector Efficiency		
• Surface Tolerance (0.018)	93.0	92.0
• Mesh Reflectivity	99.0	99.0
Sub-Total	92.0	91.1
C. Components Efficiency		
• Ku-Band Horn, Phasing, Polarizer/combiner	97.8	97.8
• Duplexer	98.5	98.5
• Power Divider	---	95.5
• Comparator	93.5	---
Sub-Total	90.0	91.9
Total Overall Efficiency (AXBXC)	65.4	65.7
Antenna Gain in dBi	46.4	47.1

TABLE 3-24. TDRS/GS ANTENNA WEIGHT BUDGET

Component	Weight (kg)
• Feed Support System	0.59
• Reflective Surface	1.95
• Ku-Band Polarizers, Diplexers, Power Splitter	0.90
• Scanner	1.13
• Comparator	0.34
• Ku-Band Rotory Joint (2)	0.45
• Subreflector	0.11
• Gimbal System	1.10
• Drive Electronics	0.93
Total	7.5

complete redundancy of all active components is utilized. The vital aspect of this link led to the addition of a microwave switch permitting the MDR/HDR #1 antenna to service the TDRS/GS transceiver providing additional redundancy and 6.5 dB more gain.

3.2.3.2.1 TDRS/GS RECEIVER

The function of the TDRS/GS receiver is to accept a 240 MHz spectrum at K_u-band and convert it to appropriate signals for distribution to the LDR, MDR/HDR, frequency source and telemetry sectionw within the spacecraft. As transmitted from the ground station, the 240 MHz spectrum is comprised of 5 signals in an FDM format depicted in Figure 3-10A, Forward Link. Due to the necessarily close spacing of the individual signals, double conversion is used to separate the IF frequencies and to circumvent a difficult RF filtering problem.

The TDRS/GS receiver block diagran is shown in Figure 3-23. The antenna signal is first fed to the diplexer which provides filtering and isolation sufficient to prevent interfering transmitter power from entering the receiver front end within the passband. In addition to this function it provides some degree of RF preselection to avoid passing wideband noise and RFI from the antenna to the receiver.

Following the diplexer, a low-loss switch is used to select either the primary or redundant RF module. The input of this module is a low noise balanced mixer. Downconversion of the input signal, centered at 13,520 MHz (Figure 3-3), is accomplished with a 14,020 MHz LO. Balanced mixers are used to maintain a low spurious response level. LO drive for the mixer is +10 dBm which is necessary to realize low conversion loss. Loss here must be kept low in the absence of RF amplification. After mixing, the IF spectrum is filtered in the first IF band-pass filter. The first IF spectrum is centered at 500 MHz and extends from 380 to 620 MHz.

The IF filter is a six-pole lumped constant design with particular emphasis on low insertion loss. This loss, together with the mixer loss and IF preamp noise figure determines receiver noise and subsequently the ground station power requirements to provide an adequate S/N ratio. Overall receiver noise figure, using this method, is 7.5 dB. The first IF amplifier uses ultra low noise devices such as 2N5650's. The 500 MHz IF amplifier increases the signal to -55 dBm. This amplifier, providing a total of 31 dB gain, is sufficient to eliminate any effect of second stage noise figure. A double balanced mixer is used to downconvert the 500 MHz spectrum to the baseband frequencies. The mixer operates with a local oscillator at 670 MHz which in turn produces the desired spectrum of 500 to 290 MHz (Figure 3-10).

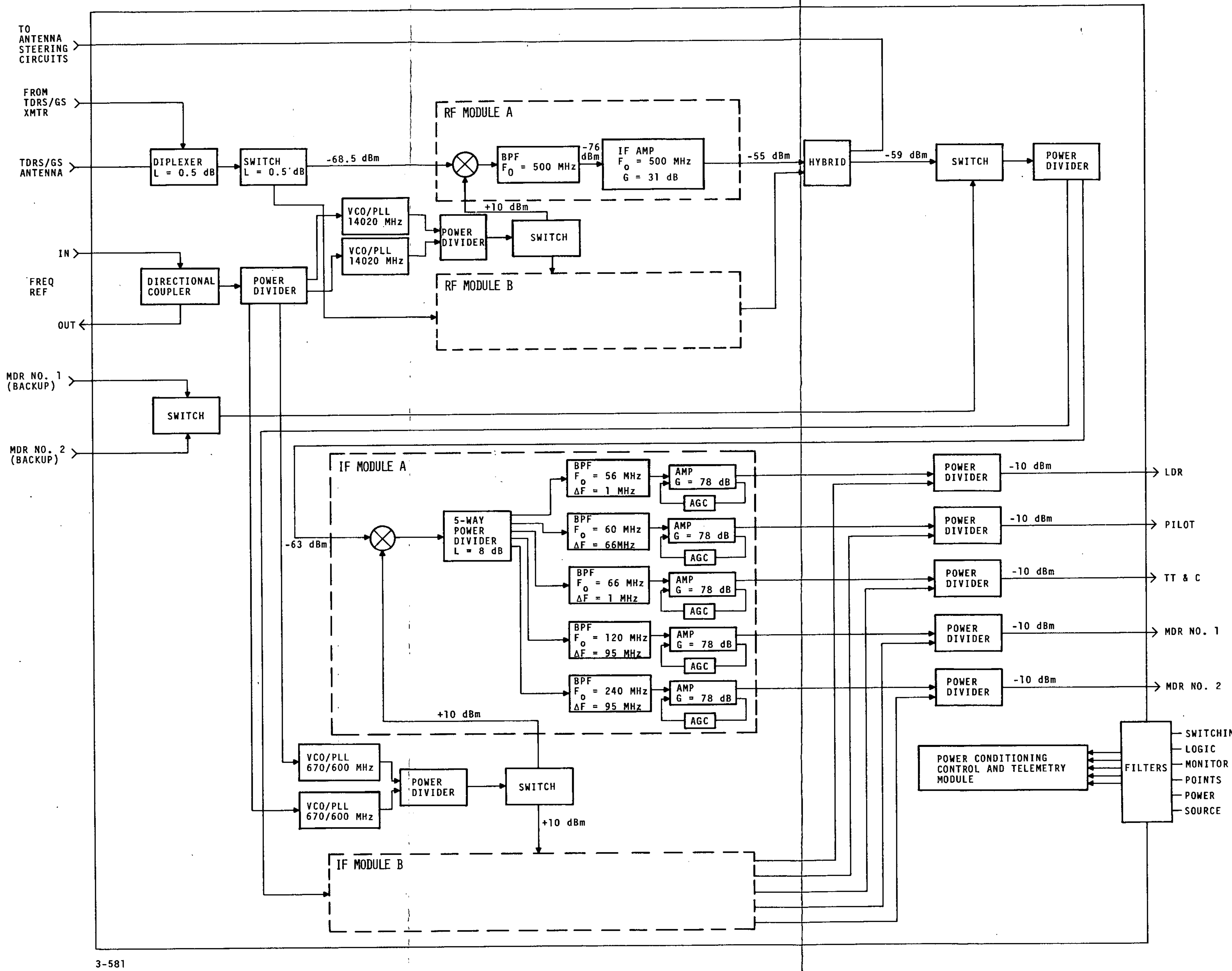


FIGURE 3-23. TDRS/GS RECEIVER BLOCK DIAGRAM
FIGURE 3-23

FOLDOUT FRAME 1

FOLDOUT FRAME 2

The 50 to 290 MHz second IF is split five ways in a power divider. Each output is fed to a separate IF band-pass filter which selects the appropriate signal.

Center Frequency	Function	Bandwidth
56 MHz	LDR	1 MHz
60 MHz	Pilot	1 MHz
66 MHz	TT&C	1 MHz
120 MHz	MDR/HDR #1	100 MHz
240 MHz	MDR/HDR #2	100 MHz

Each IF filter is a six-pole lumped constant design with sufficient skirt rejection to eliminate spurious responses and interference from adjacent spectrums. By keeping the IF power levels low ($<$ -60 dBm) the in-band harmonics of the low frequency IF signals do not cause interference in the other channels. After the individual IF's are separated by filtering, the signals are amplified to approximately -10 dBm by tuned AGC amplifiers. The amplifiers maintain a constant output power level to the transmitters.

In addition to the reliability obtained through redundancy, both 1st and 2nd LO's are cross-strapped to permit either LO to function with either mixer. Further flexibility is obtained by enabling the MDR/HDR receivers to function as backupunits for the TDRS/GS receiver. In this mode, the backup signal from the MDR/HDR receiver is used to drive the second converter, located in the IF module. This signal is at -59 dBm and centered at 500 MHz. However, the bandwidth in this mode is reduced to 100 MHz (limited by the MDR/HDR receiver and the elimination of the MDR/HDR #2 channel) and extends from 450 to 550 MHz. To be compatible with the demultiplexer, a 600 MHz LO is used, resulting in a second IF extending from 50 to 150 MHz.

3.2.3.2.2 TDRS/GS TRANSMITTER (FIGURE 3-24)

This transmitter accepts six input signals (MDR #1, Order Wire, TTAC, TDRS Tracking, LDR and MDR #2) for combining and frequency modulation and one signal (either HDR #1 or HDR #2) for frequency translation. The first group of inputs are combined to form an FDM signal extending from 1 to 50 MHz which is used to frequency modulate a K_u-band VCO (Voltage Controlled Oscillator). The output of this low level VCO, in FDM/FM format, is amplified in a solid state amplifier. Either HDR #1 or HDR #2 is selected for transmission to the ground station. Each signal can occupy a 150 MHz bandwidth. The HDR signal is upconverted to 14,675 MHz, filtered, amplified

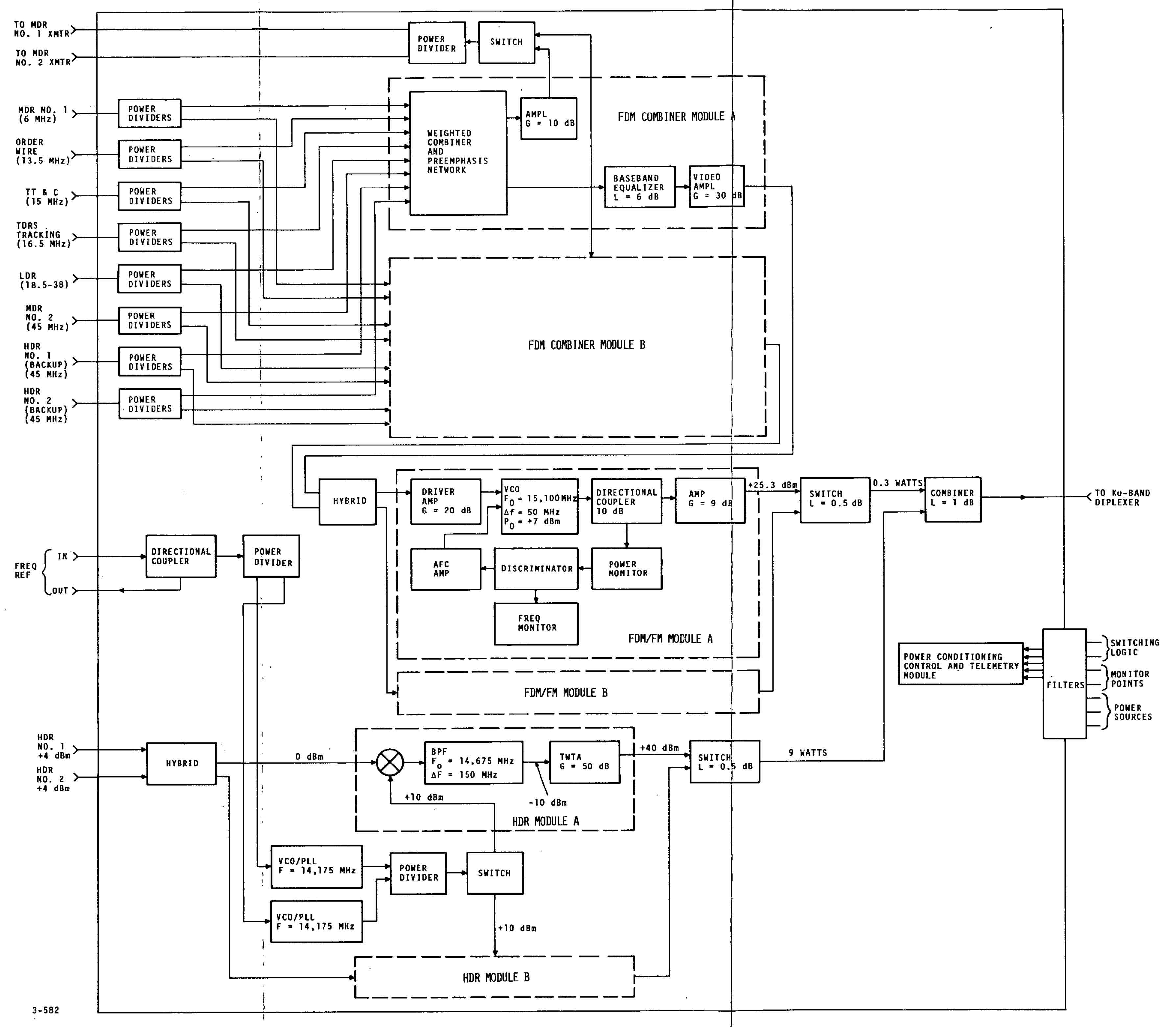


FIGURE 3-24. TDRS/GS TRANSMITTER BLOCK DIAGRAM
FIGURE 3-24

FOLDOUT FRAME 1

FOLDOUT FRAME 2

in a TWTA and combined with the FDM/FM signal. The combined signals are fed to a K_u-band diplexer and then to the TDRS/GS antenna.

Figure 3-25 depicts the baseband frequency arrangement. Also shown in Figure 3-25 is the required S/N for each channel at the output of suitable filters following the ground station detector (phase lock detector)[1]. The S/N ratio per channel is directly related to the frequency deviation attributable to that channel. Channel deviations were calculated using the relationship:[1]

$$(S/N)_o = (C/N)_i \frac{(B_{RF})}{(B_c)} \frac{(Fm)^2}{(f_n)^2}$$

where:

$(C/N)_i$ = carrier-to-noise ratio in B_{RF}

B_c = channel bandwidth

B_{RF} = RF (or IF) bandwidth

f_n = channel center frequency

$(S/N)_o$ = channel signal-to-noise output

F_m = maximum frequency deviation

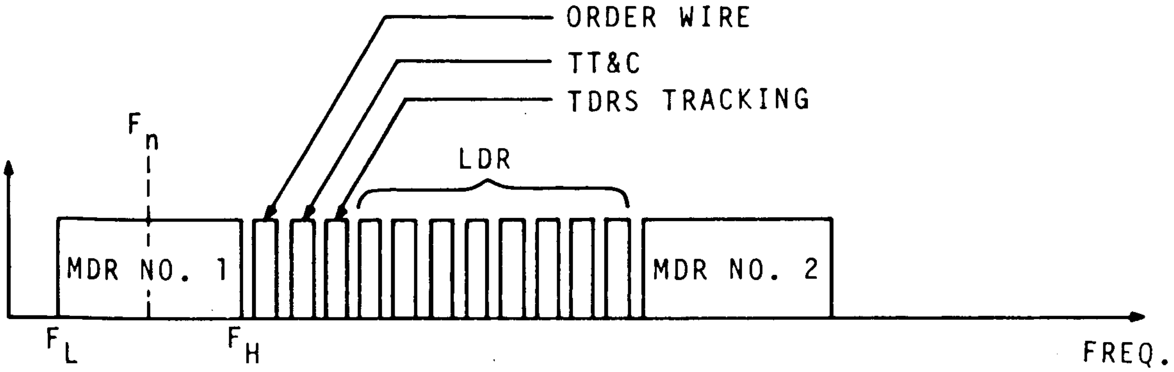
A 209 MHz bandwidth is assumed, operation at 0 dB $(C/N)_i$ is used, and the F_m is calculated. Figure 3-25 presents the results of these calculations. To verify the BW, Carson's rule for bandwidth computation is used:

$$B_{RF} = 2 (F_p + F_H)$$

---

1 Camp, "A Comparison of the Threshold Extension Capabilities of FMFB, and Phase-Lock-Loop Demodulators Employed in FDM-FM Communications System," PGCT, June 1970.

2 Panter, "Modulation, Noise and Spectral Analysis," McGraw-Hill, page 442.



FUNCTION	CHANNEL BANDWIDTH	REQUIRED S/N	F_L		F_n	F_m
MDR NO. 1	10 MHz	10 dB	1 MHz	11 MHz	6 MHz	4.15 MHz
ORDER WIRE	1	10	13	14	13.5	2.95
TT&C	1	10	14.5	15.5	15.0	3.28
TDRS TRACKING	1	10	16.0	17.0	16.5	3.61
LDR NO. 1	2	6	18.5	20.5	19.5	3.81
2	2	6	21.0	23.0	22.0	4.29
3	2	6	23.5	25.5	24.5	4.78
4	2	6	26.0	28.0	27.0	5.27
5	2	6	28.5	30.5	29.5	5.76
6	2	6	31.0	33.0	32.0	6.25
7	2	6	33.5	35.5	34.5	6.73
8	2	6	36.0	38.0	37.0	7.22
MDR NO. 2	10	10	40	50	45	31.13

3-564

FIGURE 3-25. TDRS/GS TRANSMITTER BASEBAND FREQUENCY ASSIGNMENT where:

B_{RF} = RF (or IF) bandwidth

F_p = peak frequency deviation for the composite FDM signal

F_H = maximum frequency component in the FDM signal

The peak deviation for the composite signal (F_p) is the rms combination of the MDR #1, order wire, TT&C, TDRS tracking, LDR, and MDR #2 signals. The LDR signals are first combined by adding all the F_m and using the result in the rms addition. This is done to account for the fact that the LDR signals are not independent and are, in fact, correlated. Thus, they add first on a peak basis, then on a power basis with the other signals. The result of this combination is 54.45 MHz. The highest baseband frequency is 50 MHz. Therefore, Carson's rule results in an IF bandwidth of 208.9 MHz, sufficiently close to the assumed 209 MHz. The final parameters are as follows:

F_p = 54.5 MHz

F_H = 50 MHz

Peak index = 1.09

$\frac{(C)}{(N)_i}$ = 0 dB

The Weighted Combiner shown in Figure 3-23 functions both to combine all signals to a single line and to adjust their relative amplitudes to satisfy the S/N requirement. This adjustment of the relative amplitudes of each signal is equivalent to a step-type preemphasis to obtain the required S/N for each channel. Following the combiner is an network whose function is to provide baseband equalization to compensate for system nonlinearities. The combining and equalization loss of approximately 30 dB is made up in a video amplifier whose gain is 30 dB and bandwidth extends from 1 to 50 MHz. This signal is further amplified and shaped in the driver that modulates the VCO. Signal shaping is required to obtain linear modulation of the VCO.

The VCO uses a Gunn effect diode whose frequency is controlled by a varactor coupled to the oscillator. The Gunn diode approach is used to achieve low noise. The penalty of low efficiency is not important here because only about 5 milliwatts of RF power is required. With one percent efficiency the dc power is only 0.5 watt. The varactor diode is closely coupled to the Gunn diode to obtain wide deviation capability (54.5 MHz peak deviation). The modulation sensitivity of 10 MHz per volt specifies 5.5 volts of modulation signal to obtain 55 MHz peak deviation. The driver output easily provides this level into the varactor capacitance of 2 to 3 picofarads. An AFC loop is provided to stabilize the center frequency of VCO to 15.0 GHz. This is accomplished by a 10 dB directional coupler feeding a discriminator at 15.1 GHz. The discriminator output is used to adjust the varactor bias to correct oscillator drift. The entire VCO discriminator subassembly is constructed on strip line. Following the VCO is an amplifier assembly consisting of two stages of amplification. Each stage is virtually identical and are, in fact, constructed on the same substrate. Each comprises an avalanche diode amplifier circulator-coupled to the input-output lines. The strip line configuration leads to a compact arrangement with 50 ohm input and output lines coupling to a low-loss circulator.

The HDR input signal is received by this transmitter at a center frequency of 500 MHz. Upconversion is accomplished in a balanced mixer together with an LO of 14,175 MHz. The resulting 14.675 MHz signal is filtered and then amplified up to 10 watts in a dual mode TWTA whose maximum gain is 50 dB. Also a variable is 40 dB gain with one watt output. The TWTA approach is used because the 10 watt level required is achieved with significantly less prime power than could be achieved by going solid-state. Redundancy is used to increase the reliability of the system. Each TWTA package comprises a power supply and two TWT's interconnected to permit the power supply to operate either tube. In this way the overall reliability is increased because the power supply reliability is much higher than that of the tube with minimum impact on system weight.

Combining of the FDM/FM signal with theTWTA output is accomplished in a low-loss multicoupler. The bandwidth of the multicoupler extends from 14,600 to 15,200 MHz to cover the required return spectrum as shown in Figure 3-3.

As shown in Figure 3-23, two additional signals are applied to the Weighted Combiner and Preemphasis Network besides those described in this subsection. In particular, the HDR #1 and HDR #2 signals, centered at 45 MHz and having a 10 MHz bandwidth, are used in the backup mode in which the MDR/HDR transceiver functions as the TDRS/GS transceiver. The weighting required for thismode is different because an FDM signal is transmitted instead of FDM/FM. The weighted combiner therefore has two weighting functions. The backup mode weighting function is less lossy and an equalizer is not required; more importantly, linear amplification is required in the FDM mode. Therefore, a video amplifier with a gain of 10 dB gain is used instead of 30 dB gain in this application.

3.2.4 FREQUENCY SOURCE

The changes required in the frequency source from that of the Phase I configuration are minimal and the corresponding description is virtually the same. Therefore, it will not be repeated here. The only changes are those required to generate 12,852 MHz (instead of 13,020 MHz) and 14,280 MHz (instead of 15,000 MHz). Figure 3-26 is a block diagram of the frequency source. There are no changes in its size, weight or power consumption.

3.2.5 TDRS TRACKING AND ORDER WIRE TRANSPONDER

The function of the TDRS tracking system is to provide tracking and position location data of the TDRS spacecraft by trilateration. It is accomplished by transmitting a PN code from the ground station to the TDRS spacecraft. This signal is retransmitted to two transponders located on earth. The return signals (distinctive PN codes from each transponder) are received by the TDRS spacecraft and relayed back to the ground station where range calculations are made.

The ground station combines the PN_1 code with command data and transmits it to the TDRS spacecraft on the K_u-band GS/TDRS link. PN_1 is extracted from the command data and transmitted at 2285 MHz by the earth-coverage antenna depicted in Figure 3-27. It consists of a bifilar winding on a fiber glass tube and is fed from a 180 degrees hybrid to achieve circular polarization. Two remote ground transponders, with high gain 30 foot antennas pointing at the TDRS spacecraft, receive this signal and retransmit codes PN_2 and PN_3 on a 2120 MHz carrier. PN_2 and PN_3 are unique codes permitting identification of each station. The TDRS tracking receiver converts the received signal to 16.5 MHz and feeds it to the TDRS/GS transmitter, where it is multiplexed with other signals and sent to the ground station on a K_u-band carrier.

The function of the order wire system is to provide requests for priority access to the TDRS MDR and/or HDR transponder service. These requests are received by the order wire receiver at a frequency of 2218 MHz, downconverted to 13.5 MHz and fed to the TDRS/GS transmitter where it is multiplexed with other signals for transmission to the ground station on a K_u-band carrier. All requests are sorted at the ground station and the MDR/HDR commands enable service to the selected users. The order wire re ceiver is sufficiently sensitive to service users with +16 dBw EIRP, corresponding to omniantennas.

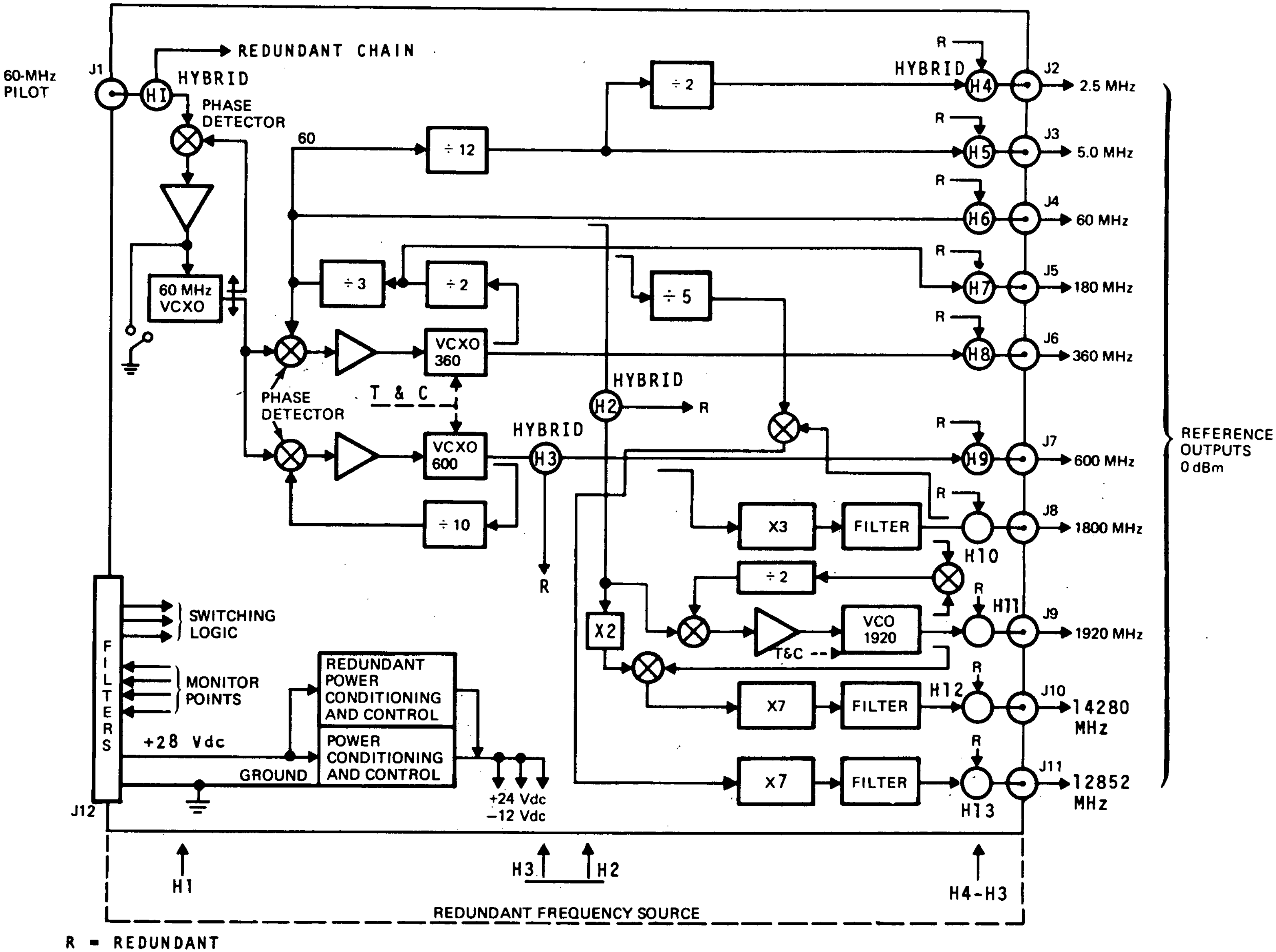


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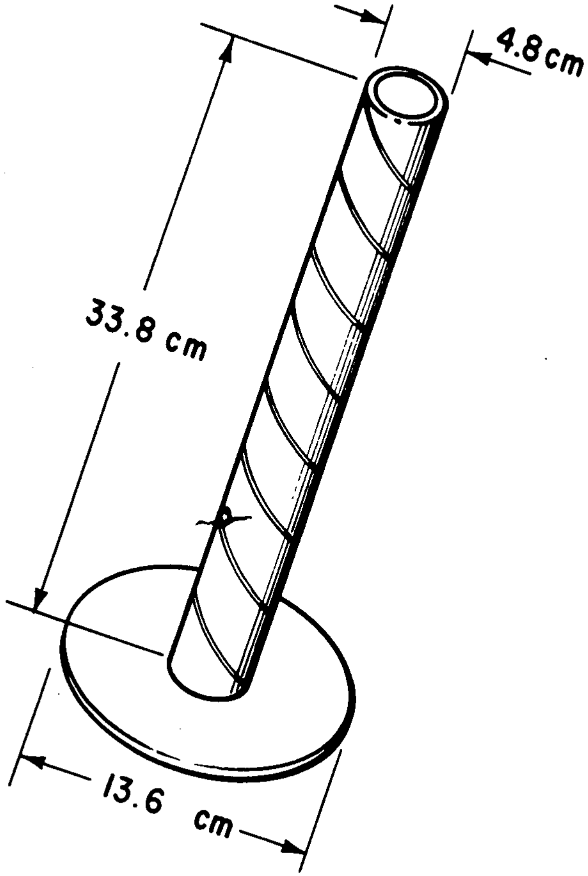
FIGURE 3-26. FREQUENCY SOURCE BLOCK DIAGRAM



PARAMETER	CHARACTERISTICS
HPBW	30°
GAIN, dBi	
• PEAK	15.0
• 17.2° FOV	14.5
• 30° FOV	12.0
WEIGHT, GRAMS	100

FIGURE 3-27. TDRS TRACKING/ORDER WIRE ANTENNA

Figure 3-28 is a block diagram of the TDRS tracking and order wire transponder. The TDRS tracking transmitter input at 15 MHz comes from the TT&C Modem Unit at a level of 0 dBm. Double upconversion is used to transmit at a frequency of 2285 MHz. The first conversion is to 500 MHz and is accomplished with a 485 MHz VCO/PLL. After filtering in a 2 MHz band-pass filter, the second upconversion takes place with a VCO/PLL at 1785 MHz. The required 12.2 watts of S-band power is generated in an all solid state amplifier providing 45.4 dB of gain. A coaxial switch accepts either the primary or redundant transmitter output for selection as the diplexer input. The diplexer output drives the S-band antenna.

Figure 3-28 also contains block diagrams of the order wire and TDRS tracking receivers. Diplexer outputs at 2218 MHz and 2120 MHz, respectively, feed these receivers. Redundant receivers are provided for each and are selectable by a low loss switch. Following this switch is a low noise (3.5 dB noise figure) transistor amplifier that provides 26.5 dB gain, necessary to minimize noise contributions of the following stages. The first mixer of each receiver is fed from the same VCO/PLL local oscillator. In the order wire receiver, the LO is below the input frequency. The LO frequency has been so chosen that the two IF frequencies are separated by 3.0 MHz. This is the exact separation required in the TDRS/GS transmitter multiplexer to achieve the frequency stacking depicted in Figure 3-25. A second downconversion, using a 64 MHz LO places the two received signals at the required frequencies of 13.5 MHz (order wire) and 16.5 MHz (TDRS tracking). This scheme permits the use of identical local oscillators for both converters of each receiver, resulting in a savings of power, size and weight.

Amplification in the two receivers is divided between the two IF amplifiers, with the most gain at the first IF, 50.5 MHz (order wire) and 47.5 MHz (TDRS tracking). These IF's are sufficiently close to permit the use of identical techniques in their construction. The output signal of each receiver is at a level of 0 dBm for transmission to the TDRS/GS transmitter.

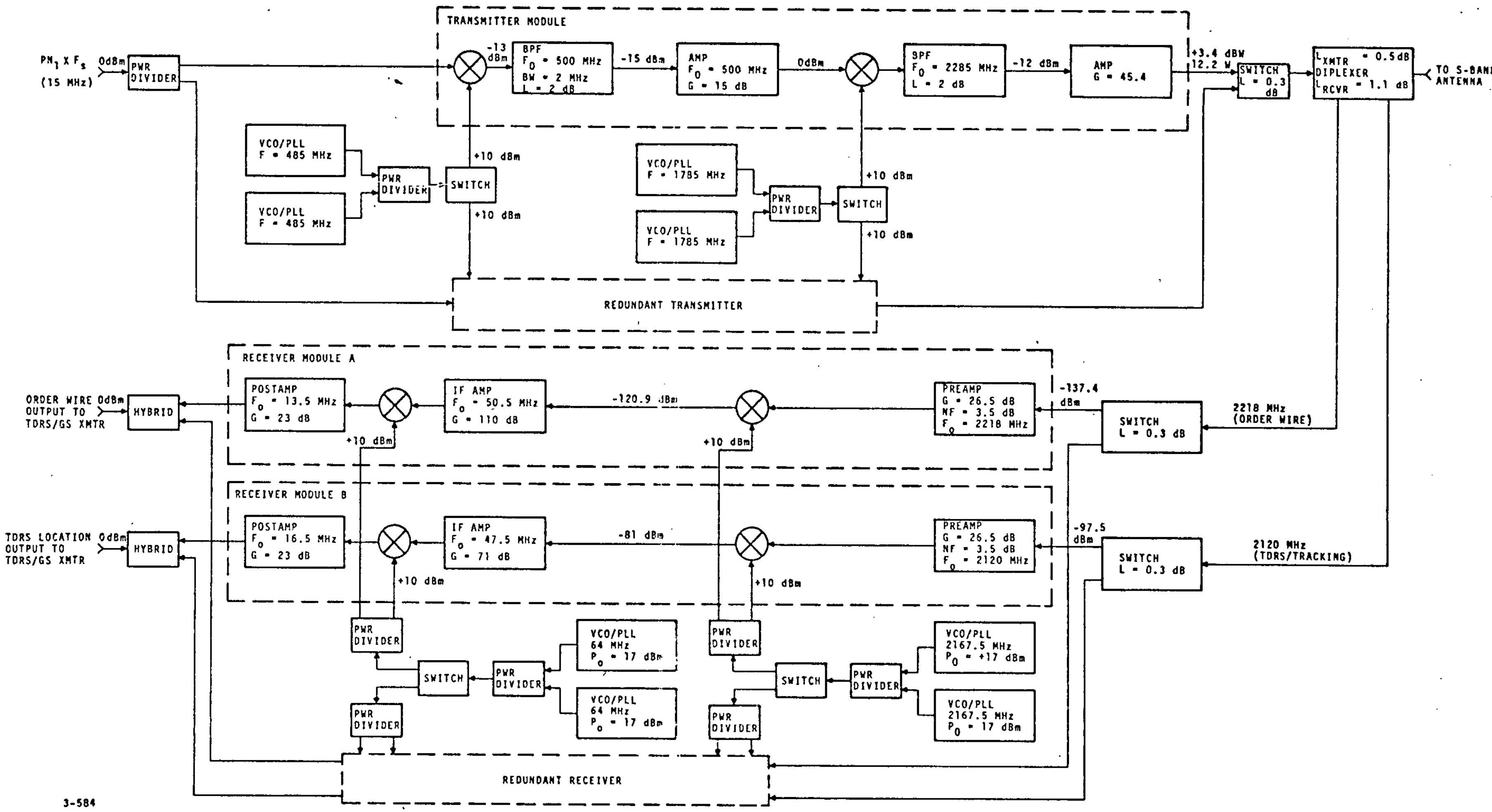


FIGURE 3-28. TDRS TRACKING/ORDER WIRE TRANSPONDER BLOCK DIAGRAM

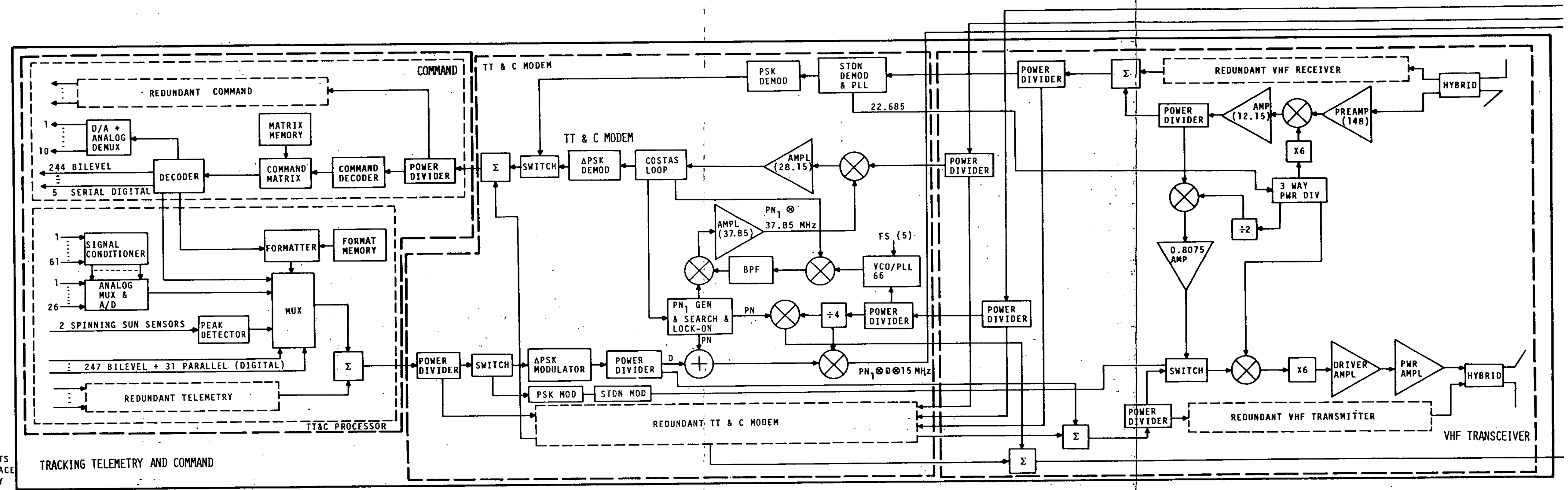
3.2.6 TRACKING TELEMETRY AND COMMAND TRANSPONDERS

An overall block diagram of the Tracking Telemetry and Command (TT&C) Transponder is shown in Figure 3-29, and includes the VHF Transceiver, 2 sets of earth coverage antennas, TT&C Modem, and TT&C Processor components. The VHF Transceiver is the primary link during the inflight transit phases, but becomes a secondary backup system on-station when the main K_u-band TDRS/GS link becomes the primary link.

Referring to Figure 3-29, it is noted that the TT&C system is the same as described for the Part I base line configuration with the following exceptions:

- VHF transmitter and receiver use separate dipole elements, thereby negating the need for a diplexer. Each dipole is fed by the main and redundant receiver (or transmitter) through a hybrid network.
- Command output of the VHF receiver feeds a STDN demodulator which demodulates the command data transmitted from the STDN station during the inflight transit phases.

Since the remainder of the TT&C Transponder has been described in detail in Part I - Volume III Report, it will not be repeated herein.



NOTES

1. ALL FREQUENCY IN MHz
2. ALL POWER IN dBm
3. DOTTED LINES AND COMPONENTS USED FOR ATLAS CENTOUR/SPACE SHUTTLE CONFIGURATION ONLY

3-565

FIGURE 3-29. TRACKING TELEMETRY AND COMMAND TRANSPONDER BLOCK DIAGRAM

FIGURE 3-29

FOLDOUT FRAME 1

FOLDOUT FRAME 2

3.3 USER TERMINAL IMPACT

In this section the impact of the up-rated TDRS on the user spacecraft telecommunications terminal design is discussed. In defining the impact on the user, the general operational requirements imposed must be evaluated. These requirements evolve from the basic configuration of the TDRS and the information transfer requirements of the users. Information transfer requirements have been defined in previous sections; functional requirements, however, are listed in Table 3-25.

The functional design of the user satellite telecommunications terminal must be performed with three options in mind, namely; 1) compatibility with the optimum design of the TDRS, 2) a compromise in the design of the TDRS and user equipment resulting in an overall minimum cost system, or 3) a design resulting in a minimum impact on the user spacecraft. It should be noted that throughout this report the user spacecraft telecommunications terminal is referred to as the user transponder.

With the incorporation of the TDRSS in the NASA reporting network, the users should be compatible with both TDRSS and STDN. In the LDR user each of the TDRSS will use unique carrier frequencies in the UHF band (400.5 to 401.5 MHz). Therefore each user must be capable of receiving two frequencies and the STDN carrier frequency. This requirement for STDN interface is an option to the user as a backup to the primary TDRSS connection. All MDR and HDR users will be assigned a unique carrier frequency at S-band or K_u-band respectively, which may be handed over from one TDRS to another. In this case again compatibility with STDN is an option to the user.

3.3.1 USER SPACECRAFT TRANSPONDER CHARACTERISTICS

The LDR, MDR, and HDR user transponders differ somewhat for several basic reasons; therefore, the characteristics and functional design for each user type will be discussed separately. Other aspects of the user transponder such as code acquisition and ranging operations are functions common to all users and thereby deferred to the end of the section. The fundamental operations required of the transponder, however, are presented in the transponder discussions.

3.3.1.1 LDR USER TRANSPONDER

The LDR user (LDRU) transponder concept presented herein is a result of evaluating the functions to be performed and consideration of the manner in which these functions could best be accomplished. The characteristics of the LDRU transponder are listed in Table 3-26.

TABLE 3-25. USER SPACECRAFT TRANSPONDER REQUIREMENTS

LDR USER	MDR USER	HDR USER
° Compatibility with both TDRS and STDN	° Compatibility with both TDRS and STDN	° Compatibility with both TDRS and STDN
° Tuneable to one of two UHF receiver frequencies (TDRS) and one STDN frequency	° Tuneable to one of two S-band receiver frequencies (TDRS) and one STDN frequency	° Tuneable to one of two K_u-band frequencies and one STDN frequency
° PN modulation necessary for multipath rejection on both forward and return links	° PN modulation necessary on forward link to reduce signal power flux density	° PN modulation necessary on forward link to reduce signal power flux density
° PN modulation required for bandwidth spreading to meet IRAC requirements	° PN modulation on return link to aid in acquisition	° PN modulation in return link to aid in antenna acquisition
° Up to twenty (20) users per TDRS (simultaneous access in return link)	° Up to two users per TDRS	° Maximum of one user per TDRS

TABLE 3-26. LDR USER TRANSPONDER CHARACTERISTICS

Forward Link	
° Data Rate	100 to 1,000 bps
° Carrier Frequency	2 channels - 401 MHz
° PN Code Rate	668 Kchips/sec
° Code Length (Gold Sequence)	2047 chips
° Range Ambiguity	Coded word of duration ≈9 x 10^4 chips
° One PN Code Sequence To All Users	
Return Link	
° Data Rate	1,000 to 10,000 bps
° Carrier Frequency	136 MHz (1 channel)
° PN Code Rate	≈1 Mchip/sec
° Code Length (Gold Sequence)	≈8191 chips
° Unique Code Sequence for each User (TDRSS Access is CDMA)	

The LDR user transponder must be capable of receiving commands, transmitting data, and providing PN code coherence so that a ranging computation can be performed on the ground. For a number of reasons discussed* previously a PN code modulation is used in both the forward and return links to the LDR user, therefore before any function can be performed a connectivity must be established by synchronizing to the received PN code in the user transponder and at the TDRS GS and by properly tuning the user receiver to the one of two TDRS to user carrier frequencies being used. Having established connectivity the transmission of data in both directions can be accomplished. Tracking is accomplished by computing the transit time of the signal from TDRS to user to TDRS. This feature establishes the minimum code length of the PN sequence for a given code rate. The minimum code rate is established by the amount of bandspreading required.

*"Tracking and Data Relay Satellite System Configuration and Tradeoff Study," Part I Final Report, Vols. III & IV, prepared by Space Division, North American Rockwell, under NASA Contract NAS5-21705, October 1972.

The command message transmitted to a user will contain the instruction directing the proper tuning for the next transmission after the present communication is completed. The advantage of this technique is that very little hardware is required to implement the operation.

The concept for PN code acquisition is more involved. The need for PN comes from the need to spread the TDRS to user transmission such that IRAC requirements are met, and to discriminate against the multipath and give modulation for code division multiplexing of the user transmissions and for tracking. In Section 3.3.3 a discussion of the code acquisition and tracking illustrates the problems involved and the concept eventually selected for this system. In essence, a short code is transmitted for the initial acquisition phase of the communication in order to get a reasonable acquisition time (40 sec. or less). The resultant code sequence (i.e., code length) configuration results in ambiguous ranging information.

In the Part I study* a two mode coding technique was used: a short code (2047 chips) for acquisition, and a long code (11 x 2047 in the forward link and 66 x 2047 in the return link) for the tracking mode with the switching taking place in both the user and TDRS GS upon command so that coherence in the user would not be lost and sync could be maintained in the user and GS during the switch.

In Part II of the study the change from an LDR steered beam concept to one of fixed-field of view necessitated the shift to a short code only in the forward link. This change was necessary to prevent users tracking the long code from loosing code sync (on the long code) while the short code was being up-linked to acquire new users.

Two way ranging is achieved via the PN code transmission with range resolution, as described in the Part I report, a function of the PN code rate. Ambiguity resolution which was formerly achieved by implementation of the long code sequences is now accomplished by uplinking (after the links have been synchronized to the PN codes) a unique data word interleaved with normal command/telemetry transmissions whose length is approximately equal to the two way range ambiguity of 40,000 km or approximately 133 msec. If a 1 kbps command rate is assumed this amounts to an ambiguity resolving data word of approximately 133 bits. At 668 kchips/sec code rate in the forward link this amounts to approximately 9×10^4 chips; likewise in the return link ($\approx$1 Mchip/sec) the coded data word is approximately 13×10^4 chips. The PN code acquisition requires a search of frequency uncertainty and time uncertainty and time uncertainty regions. The LDRU Acquisition procedure is shown in Table 3-27.

*Ibid., pp. 11-1 to 11-20

TABLE 3-27. LDR USER ACQUISITION PROCEDURE

° Ground Station Uplink PN Code via TDRS F-FOV Antenna at UHF

° User Acquire PN Code; Acquire Carrier/Track Doppler

° User Downlink PN Code

° Ground Station Acquire PN Code

° AGIPA Processor Maximize Signal/Interference

° Link Synchronized for Transmission/Ranging

Aside from the serial time and frequency search two techniques were considered (during Part I) which allowed a faster search of the uncertainty ranges: digital matched filter and digital spectral analysis. The latter technique has been selected for implementation.

This approach involves an all-digital discrete Fourier transform which minimizes synchronization time resulting from doppler uncertainties. In this case, contrary to the digital matched filter, the time uncertainty is searched in 1/2 chip increments over the pseudonoise code length and for each increment the total doppler uncertainty is searched in real time. A simplified functional diagram of the real time doppler uncertainty resolver is illustrated in Figure 3-30. Where the digital matched filter is essentially a fast time search technique the digital spectral analysis (doppler processor) is a fast frequency search approach. A more detailed discussion of the parameters of the doppler processor is presented in Appendix B.

Convolutional encoding is applied to the data for error control following delta-coding so that the phase ambiguity introduced by the Costas Loop (which mathematically is equivalent to a squaring loop) may be resolved in the demodulation process. A discussion of a typical forward error control concept employing convolutional coder and a maximum likelihood (Viterbi) decoder is presented in Appendix C.

3.3.1.2 MDR USER TRANSPONDER CHARACTERISTICS

The MDR concept is similar to the LDR concept with a few exceptions. The MDR transponder is required to perform the same operational functions, but because of the difference in operating frequency, the number of users accommodated, and the difference in the user and TDRS antenna characteristics, the method of implementing these operational functions are different. PN code modulation also is used in the MDR case in both the forward and return links. Therefore, in establishing connectivity a code acquisition must be accomplished. The MDR user (MDRU) is divided in two

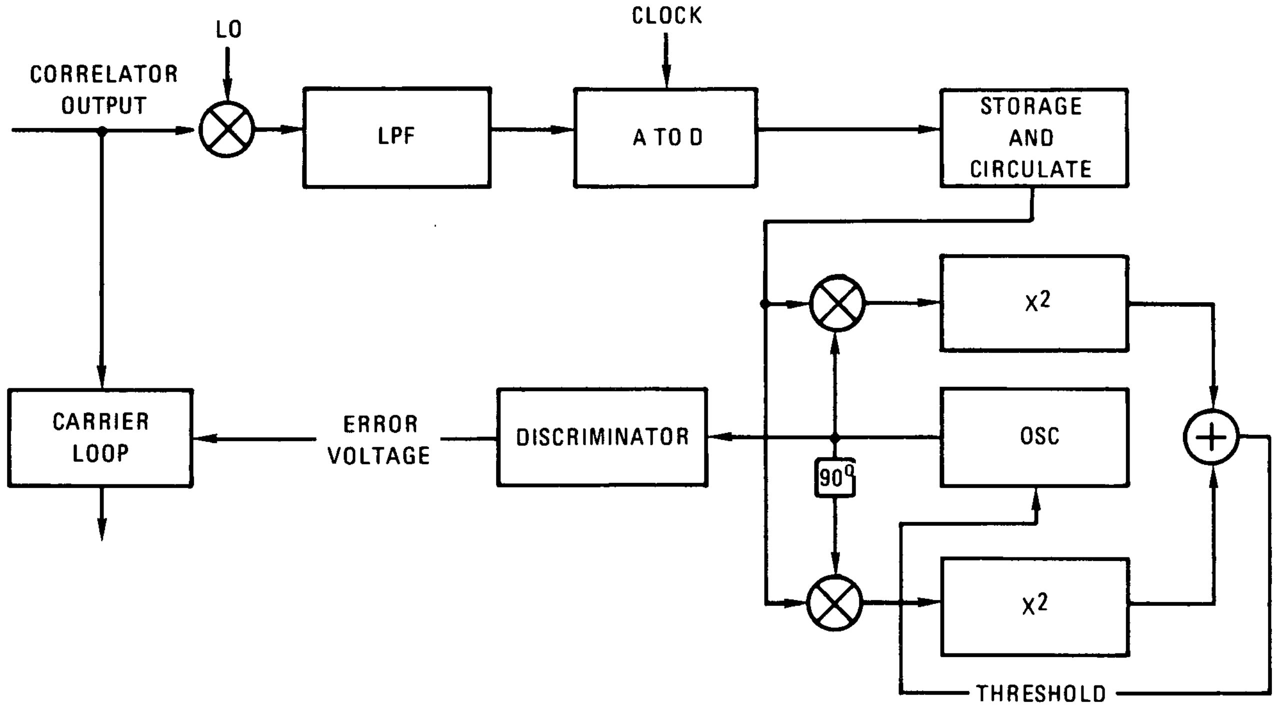


FIGURE 3-30. SIMPLIFIED FUNCTIONAL DIAGRAM OF THE DOPPLER RESOLVER

classes, the spec MDRU and the Manned MDRU. In this section the two will be discussed as one and differences between them noted.

One unique S-band carrier frequency is assigned to each MDR user. Since the TDRS for this case is a simple translating repeater, the unique carrier frequency for any of the MDR users can be selected on the ground and transmitted through either of the two TDRSS. The characteristics of the MDRU transponder is presented in Table 3-28.

TABLE 3-28. MDR USER TRANSPONDER CHARACTERISTICS

Parameter	Spec User	Manned User
Forward Link		
° Data Rate	100 to 1,000 bps	Data = 6 kbps
° Carrier Frequency	2025 to 2110 MHz (2 channels) (common PN code)	Voice = 48 kbps (ΔMod)
° PN Code Rate	5 Mchips/sec	20 Mchips/sec
° Code Length (Gold Sequence)	2047 chips	8191 chips
° Range Ambiguity	Coded word (Duration $\approx 7 \times 10^5$ chips)	Coded word (duration $\approx 3 \times 10^6$ chips)
Return Link		
° Data Rate	10 to 1,000 kbps	Data = 128 kbps Voice = 64 kbps (ΔMod)
° Carrier Frequency	2220 to 2280 MHz	
° PN Code Rate	5 Mchips/sec	5 Mchips/sec
° Forward Error-Control	Available	Required

The concept for code acquisition is somewhat similar, however, the PN code modulation for the MDR user is used primarily to derive range and range rate information and to spread the signal energy in the TDRS to user link to conform to IRAC requirements. The basic circuitry is the same as the LDRU case, i.e., use of a doppler processor, early-late gate tracker, and basic short code length.

A short code (2047 chips and 8191 chips for the spec and manned user, respectively) is transmitted from the TDRS Ground Station to the user. Code acquisition requires a maximum of 30 seconds. After code synchronization in the MDRU the next all 1's vector enables the return link code generator and transmitter to allow the ground station to synchronize on the code. Synchronization of both ends of the link is required before command or ranging signals can be uplinked to the user. The ranging signal for the MDRU is a coded data word (at the normal command/telemetry rates) which is 133 msec in duration (similar to the LDRU case). At a code rate of 4 Mchip/sec as in the spec MDRU this would amount to the equivalent of approximately 7×10^5 chips. However the ambiguity resolving data word is still, assuming a command rate of 1 kbps, on the order of 133 bits (133 msec x 10^3 bits/sec). A functional description of the MDRU Acquisition Procedure is shown in Table 3-29.

TABLE 3-29. MDR USER ACQUISITION PROCEDURE

Spec User	Manned User
° TDRS steers MDR antenna into user sector	° Shuttle request MDR service via order wire and turns on S-band beacon
° User commanded via S-band omni to steer S-band Directional antenna toward TDRS	° TDRS S-band antenna acquires shuttle beacon through programmed scan operation
° User commanded via high-gain S-band antenna to activate transreceiver	° Up-link PN code
° Ground station up-link PN code	° Shuttle acquire PN code - acquire carrier - track doppler
° User acquire PN code; acquire carrier/track doppler	° Shuttle down-link PN code
° User down-link PN code	° Ground station acquire PN code - acquire carrier/doppler
° Ground station acquire PN code; carrier/doppler	° Link synchronized for command/ telemetry transmission
° Link synchronized for command/ telemetry transmission	

The command message (MDR case) and TDM'd voice and data (manned case) can be extracted from the in-phase side of the Costas loop and processed to recover command data and two voice channels.

In the return link the PN code is mod-2 added to the data to form the modulating signal because the data rate is comparable to the code rate, and since the PN code is being transmitted only for tracking purposes, the PN code is treated like an additional data signal. The PN code bi-phase modulates an IF carrier which is then mixed with an IF carrier modulated by the data. The combined translated signal is then the transmission frequency.

3.3.1.3 THE HDR USER TRANSPONDER CHARACTERISTICS

The fundamental differences between the HDR user (HDRU) transponder and the others are the high return link telemetry rates (greater than 1 Mbps) and the requirement for a relatively high gain antenna. PN code modulation is used in the forward link for spectrum spreading and to provide a ranging signal. In the return link the code is used for ranging and identification during the HDRU antenna acquisition procedure. The basic HDRU transponder characteristics are presented in Table 3-30.

TABLE 3-30. HDR USER TRANSPONDER CHARACTERISTICS

Forward Link	
° Data Rate	100 to 1,000 bps
° Carrier Frequency	14.81 - 14.91 GHz
° PN Code Rate	60 kchips/sec (best case) 60 Mchips/sec (worst case)
° Code Length	2047 chips (gold sequence)
° Range Ambiguity	Coded Word (Duration ≈8000 chips)
Return Link	
° Data Rate	Greater than 1 mbps (plus 100 mbps)
° Carrier Frequency	13.85 - 14.0 GHz
° PN Code Rate	5 Mchips/sec
° Forward Error Control	Available to enhance return link performance at 100 mbps

The PN code in the forward link has a rate of approximately 60 kchips/sec with a code length of 2047 chips. To meet the IRAC requirements in the high power mode (i.e., 53.6 dBw to support forward link video) a chip rate of 60 Mchips/sec is required. The code length in the high power mode is on the order of 16,000 chips. Range ambiguity is resolved in the same manner discussed previously.

Acquisition of the HDRU antenna requires more sophistication than that for other types of users. Detailed descriptions of the procedure has been presented in previous sections. A summary of the HDR User Antenna Acquisition Procedure is presented in Table 3-31.

TABLE 3-31. HDR USER ANTENNA ACQUISITION PROCEDURE

° TDRS Steers Dual-feed MDR/HDR Antenna to User Sector
° Acquisition Commands Uplinked via User's S-band Omni
° Activate User K_u-band Beacon through Low Gain Antenna
° TDRS Acquire User K_u-band Beacon through Programmed Scan Procedure
° User Commanded to Switch to High Gain Antenna and Initiate Limited Acquisition Search Procedure
° TDRS Activate K_u-band Transmitter
° User Acquire TDRS K_u-band Signal

3.3.2 USER SPACECRAFT TRANSPONDER MECHANIZATION

Functional designs for the LDR, MDR, and HDR user transponders are presented herein. With exception of the HDR user, which has not been considered previously, the basic designs of the user transponders has not been altered to any great extent. In the discussion that follows operational features of the transponder (which are essentially the same for all three classes of users) are discussed once in the discussion of the LDR user transponder design. Similar functions for other user types are merely referred to.

3.3.2.1 LDR USER TRANSPONDER DESIGN

The block diagram for the LDR user transponder is shown in Figure 3-31. The receiver is shown segmented into: RF, code tracking loop, carrier tracking loop, doppler processor, demodulation, and frequency synthesizer sections. There is also a controller from which signals will be directed to segments to initiate specific functions and receive signals to maintain operational status of the receiver and transmitter.

The RF segment consisting of the RF and IF circuitry is straightforward, except for the need to switch to the appropriate injection frequency at the first or second mixer to tune the receiver to the proper channel to receive the command. The procedure used here is to transmit in the command message an instruction to the frequency synthesizer selecting the proper injection frequency for the next transmission such that at the end of the cur-

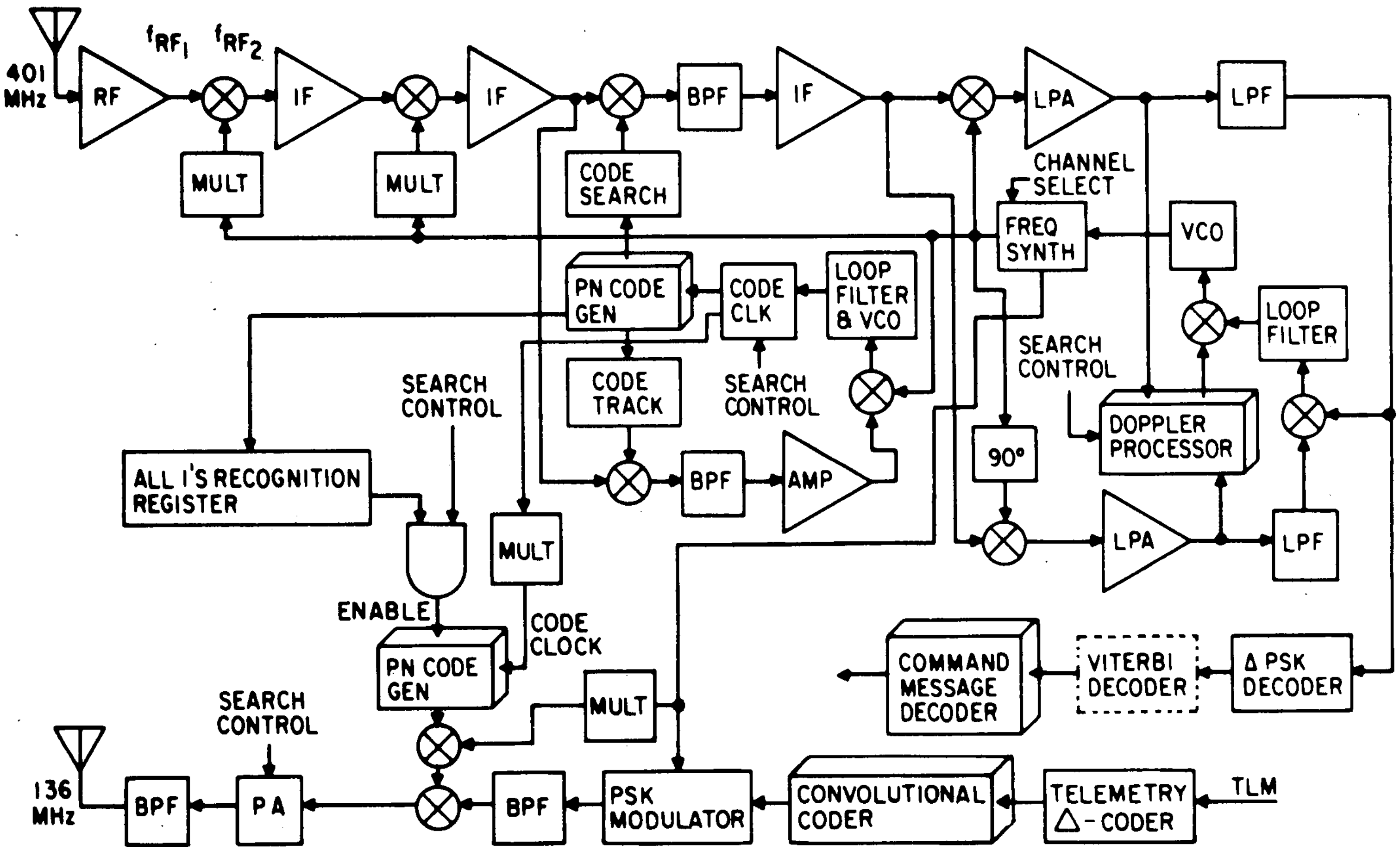


FIGURE 3-31. LDR USER TRANSPONDER

rent transmission the injection frequency will be switched to the proper channel. To insure against the synthesizer changing frequency in the case of inadvertent loss of lock, the frequency synthesizer will only switch to the new programmed injection frequency upon command from the controller. The switch command from the controller will be generated after some N code acquisition cycles have been performed by the receiver assuring that loss of lock was due to end of communication.

The IF signal is split into two channels one going to the carrier tracking loop and one going to the code tracking loop.

The carrier tracking loop is a Costas loop (mathematically equivalent to a squaring loop). The function of this loop is to reduce the signal to baseband and strip the data off at the output of the in-phase low pass filter. The signal output of the loop filter is the error signal denoting the difference between the reference signal and the IF signal. This error signal is used to vary the VCO to the proper injection frequency to account for frequency uncertainties. But the magnitude of the doppler uncertainty will require unreasonable pull-in times even with large acquisition bandwidths, therefore a frequency search mode is provided to quickly sweep the loop VCO to within a reasonable error frequency (equivalent to the reciprocal of the data rate). This is accomplished by means of the doppler processor.

The doppler processor provides a correction voltage to the carrier loop VCO which is then used as one of the two reference frequencies in the frequency synthesizer. In this way the doppler correction is injected in all frequencies where the doppler correction is necessary. The processor will also indicate code acquisition. The inputs to the doppler processor are taken from the in-phase and quadrature signals in the carrier tracking loop following the low pass amplifiers which have bandwidths equivalent to the reciprocal of the data rate. Therefore, when the codes are out of sync the signal energy in the data bandwidth will be small. When the codes are in sync, the signal energy in the data bandwidth will increase by the PN process gain. The doppler processor will indicate a "hit" or signal presence in a 100 Hz frequency cell when the PN codes are in synchronism because of the threshold exceedence. On every frequency sweep, the average energy level over the entire frequency uncertainty region is computed. This value is used as the floating reference point. The threshold is established by selecting a margin by which the maximum signal must exceed the floating reference. This margin is the threshold factor. Details of the doppler processor can be found in Appendix B.

Having detected the presence of a signal as the doppler processor scans the frequency range in one pass, further decision strategy is employed whereby two out of three consective detections, called "hits" are required to be declared a valid hit. This increases the true detection probability and decreases the false alarm rate under the threshold condition of 10 dB S/N in 100 Hz. At the conclusion of a valid hit, an analog voltage corresponding to the detected doppler frequency is sent to the carrier and code loop VCO's. This voltage effectively drives the local oscillator frequency to the input carrier for rapid acquisition.

After each sweep, if the presence of a signal is not detected, then the reference PN code is retarded by 1/2 chip in the code tracking loop and the entire frequency search process repeated in the doppler processor.

After code acquisition is detected then the controller will perform a number of functions indicated as search control in the block diagram:

1. The 1/2 chip retard function is stopped, and the early-late gate tracker is activated.
2. Activate the output gate of the all 1's recognition register such that the transmitter PN code generator start pulse is outputted to the transmitter.
3. Activate the transmitter power amplifier.
4. Initiate loss of lock count in the frequency synthesizer in case of inadvertant loss of lock.

The early-late gate tracker has now been activated in the code tracking loop. The IF signal is mixed with a reference IF modulated by the reference PN code through the code tracker.

The frequency synthesizer will generate a number of frequencies from two reference sources. The injection frequencies in the RF/IF segment of the receiver are generated from a stable source while the injection frequency for the carrier tracking loop is derived from the carrier tracking loop VCO making the injection frequency doppler corrected.

The receiver is now ready to receive a command message transmitted from the TDRS ground station. The data is taken off the in-phase side of the carrier tracking loop and applied to a ΔPSK decoder to remove the differential coding. The command message is then applied to the message decoder and the various instructions identified. The instructions of concern to the mechanization of the transponder are the injected frequency selection for the next communication and the command verification from the user spacecraft. Receipt of command verification at the ground station indicates that the link is fully operational and ready to execute command/telemetry and tracking functions. Although not indicated in the transponder block diagram, the LDRU transponder should interface with the STDN transponder.

The functioning of the LDRU transmitter is rather straightforward except for the control signals from the receiver. The PN code generator is an 11-stage code generator with logic to initiate operation of the code. This is done upon verification of sync on the forward link PN generator. When the start pulse from the receiver all 1's recognition register is received, the transmitter code generator is started and the transmitter activated.

The PN code is transmitted to the TDRS ground station. After the code acquisition is achieved in the ground station, the command message is transmitted to the user on short code.

Telemetry data from the user spacecraft is applied to a delta-coder whose output is convolutional coded and applied to a bi-phase modulator. The modulator is mixed with the upconverter frequency which has been bi-phase modulated by the return link PN code, and the composite signal applied to the power amplifier.

LDR Transponder - Size, Weight, and Power Estimates

The estimates for a size, power, and weight have been made with the assumption that the LDR will have a transmit power of 37 dBm (5 watts). Additional assumptions made are as follows:

1. No secondary power supply
2. Hardware is 1974 technology
3. Micropower logic
4. No hybrid technology
5. No radiation hardening

The transponder size and power are distributed as follows:

Item	Size (cc)	Power (watt)
Receiver:		
RF/IF assembly	492	1
Local ref. correlator	492	1
Costas demodulator	738	3
Doppler resolver	246	2
Coder/clock	246	1
Controller	246	1
Data processor	246	1
RF synthesizer	492	2
Transmitter:		
RF power amp (40.8 dBm)	2460	10
Modulator	492	2
Data processor	246	1
RF synthesizer	492	3
Total	6888	18

The weight distribution is: transmitter - 1.8 kilograms, receiver - 2.7 kilograms.

3.3.2.2 MDR USER TRANSPONDER DESIGN

As mentioned, the MDR user is divided into two generic types, the spec MDRU and the manned MDRU. The basic design of the spec and manned MDRU transponder is the same. Throughout this discussion no distinction will be made between the basic spec and manned user transponders except where there are significant differences.

Functional block diagrams of the spec MDRU and manned MDRU terminal are shown in Figures 3-32 and 3-33 respectively. Each receiver can be segmented into RF, code tracking loop, carrier tracking loop, demodulator/decoder, and doppler processor. There is also a controller from which signals will be directed to segments to initiate specific functions and receive signals to maintain operational status of the receiver and transmitter.

The RF segment consisting of the RF and IF circuitry is rather straightforward. The output of the IF is split into two channels, one going to the carrier tracking loop and one going to the code tracking loop. The carrier tracking loop is a Costas loop. The function here is to reduce the signal to baseband and to extract the data and voice for further processing in the decoder segment. The functional description of the Costas loop was given previously.

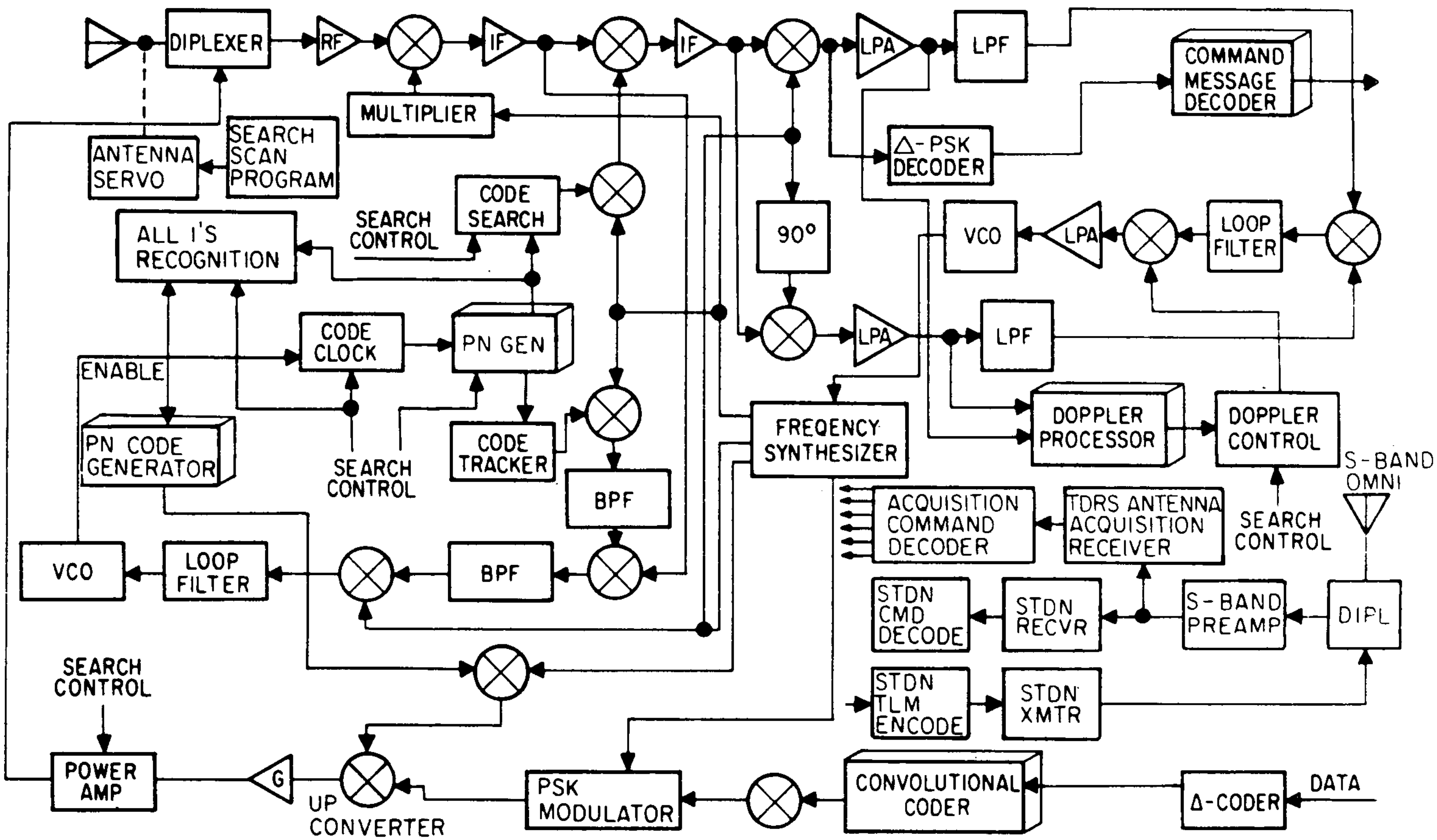


FIGURE 3-32. SPEC MDR USER TRANSPONDER

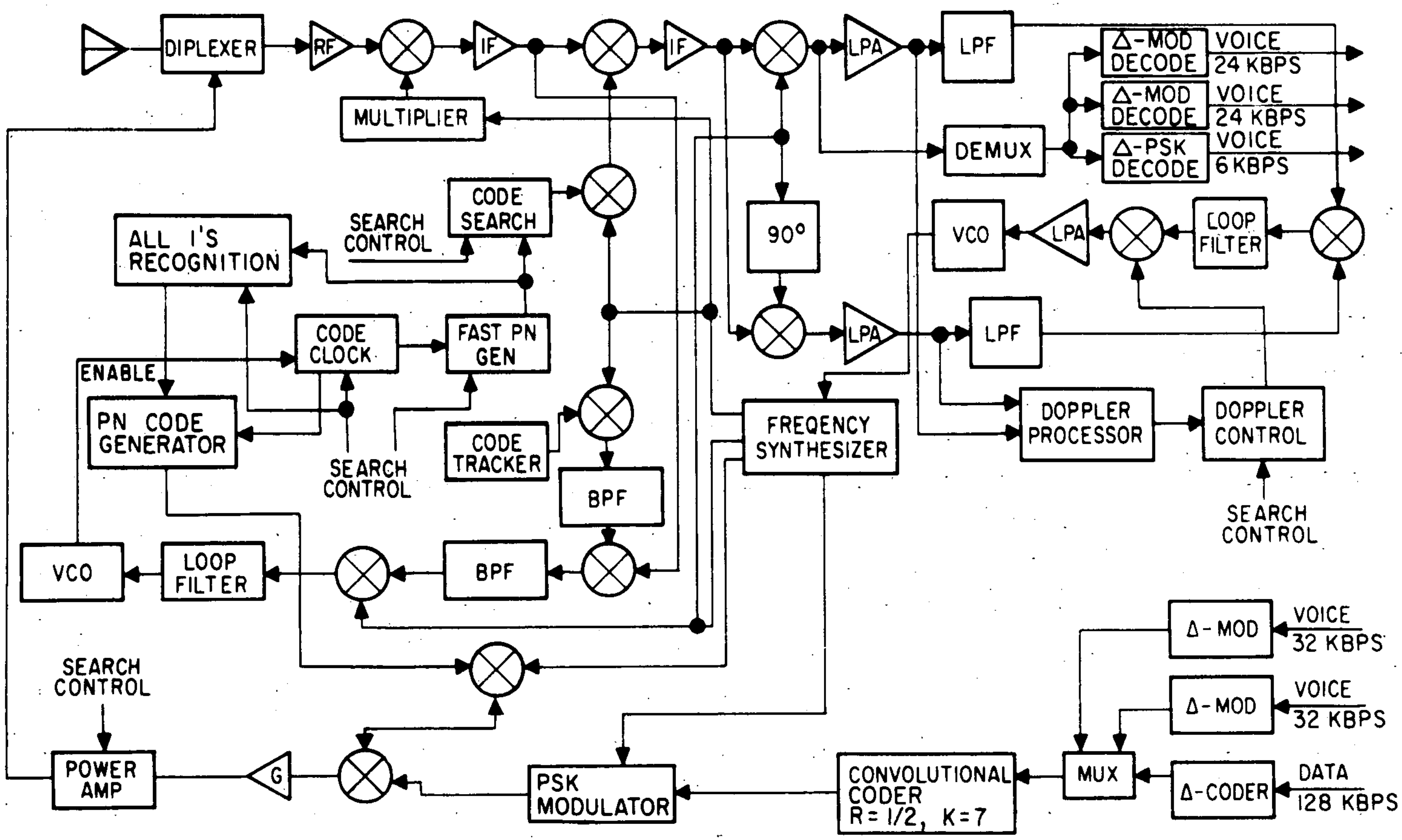


FIGURE 3-33. MANNED MDR USER TRANSPONDER

The code acquisition function is essentially the same as that used in the LDR case except for some small modifications which will be discussed here. The code acquisition operation is started by the TDRS GS repeatedly transmitting a single 2047 chip sequence (8191 for the manned user). In the user, the receiver will first search the frequency uncertainty band. If a signal is not detected, the reference PN code is retarded by 1/2 chip and the frequency search repeated. This procedure is continued until code acquisition is achieved. The frequency search is performed by the doppler processor.

The functional description of the doppler processor has been presented previously but the mechanization parameters must be changed to accommodate the MDR case. That is to say, with an increase in the doppler rate the search strategy must change. The computation of the Fourier coefficients are performed in one mode (1000 bps). The evaluation of this mode of operation is presented in the Part I, Final Report.* Each frequency search will take 1 msec. The maximum synch time is equal to a search through 2047 chips at 1/2 chip per frequency search with 1 msec required for each frequency search or a synch time less than 10 seconds. For the manned MDRU case the sync time is on the order of 40 sec.

After the code has been acquired, the next all 1's vector of the forward link code enables the return link PN generator and the power amplifier. The ground station acquires the code through a method similar to the forward link acquisition.

Two way range and range rate measurement is accomplished via the PN code. Ambiguity resolution is achieved in the same manner outlined for the LDRU, namely a code data word (133 msec in length). Assuming forward link command rates of 1 kbps and 3 kbps for the spec and manned user respectively the corresponding ambiguity resolving data word would be 133 bits and 399 bits respectively. Similarly, the number of bits in the return link ambiguity resolving code is proportional to the return link telemetry rate. It must be emphasized that the basic range and range rate accuracies are a function of the PN code rate, ** and not the data word used to resolve ambiguity.

The demodulated output of the spec MDRU transponder is applied to a PSK decoder and then to the command decoder; in the manned MDRU transponder the Costas loop output is applied to a demultiplexer which is clocked at 54 kHz to demultiplex the uplinked TDM voice and data. The three outputs of the DEMUX consists of one delta-coded data channel at 6 kbps and two delta-modulated voice channels of 24 kbps each. These three channels are applied to delta decoders to recover the baseband equivalents.

*Ibid. pp. 11-15 to 11-17; also, see Appendix C.
**Ibid. pp. 3-51 to 3-67.

The MDRU transmitters are identical for the manned and spec user, with the exception of data encoding. Typically, in either system the coded return link data stream is encoded (convolutional coder), applied to a bi-phase modulator whose output is mixed with the carrier frequency (which is bi-phase modulated by the PN code) and up-converted to the desired return link frequency. The up-converted signal is applied to the power amplifier for transmission. In the spec user transponder the input to the convolutional coder is merely delta-code telemetry data. The manned user, however, is required to delta-modulate two voice channels at 32 kbps each, delta-code a 128 kbps telemetry channel, and then time-division multiplex the three together into a serial data stream at a 192 kbps rate.

MDR Transponder - Size, Weight, and Power Estimates

The estimates for size, power, and weight were made assuming the unmanned MDRU has a transmit power of 40.8 dBm (12 watts). Additional assumptions made were:

1. No secondary power supply
2. 1974 technology hardware
3. Micropower logic
4. No hybrid technology
5. No radiation hardening

The transponder size and power are distributed as follows:

Item	Size (cc)	Power (watt)
Receiver:		
RF/IF assembly	738	1
Local ref. correlator	492	1
Costas demodulator	738	3
Doppler resolver	246	2
Coder/clock	246	1
Controller	246	1
Data processor	246	1
RF synthesizer	492	2
Transmitter:		
RF power amp (40.8 dBm)	2460	25
Modulator	492	2
Data processor	246	1
RF synthesizer	738	5
Total	7298	45

The weight distribution is: transmitter - 1.8 kilograms, receiver - 3.6 kilograms.

Deleted from the size, weight and power estimates for the spec MDRU transponder are those details relevant to the STDN transceiver and the antenna acquisition receiver and command decoder.

Estimates of size, weight, and power for the manned MDRU transponder have not been developed specifically. With the spec and manned user transponders being so similar it seems reasonable that the increased size, weight, and power requirements of the manned MDRU transponder (allowing for some increase in the RF power amplifier and the modems) will not exceed that of the spec user transponder by more than 10 percent.

3.3.2.3 HDR USER TRANSPONDER DESIGN

The basic functional design of the HDR User (HDRU) transponder is similar to that of the MDRU transponder. Fundamental differences are higher RF frequencies and a dual-gain antenna system. The HDRU transponder is shown functionally in Figure 3-34.

The HDRU transponder like the MDRU and LDRU transponders is comprised of a receiver section including: RF/IF subsystems, code tracking loop, doppler processor, decoders and frequency synthesizer and a transmit section consisting of a delta-coder, convolutional coder, PSK-modulator, return link PN code generator, and power amplifier. In addition, the HDRU system includes a relatively sophisticated servo system for steering the HDRU high gain antenna. Furthermore, the HDRU requires a simple autotrack system for maintaining antenna lock during the users transit across the TDRS coverage angle. Details of these two functions are not presented in this report.

The carrier tracking loop in the HDRU receiver is a Costas loop. Doppler acquisition is aided by the "Doppler Processor" which analyzes the frequency uncertainty range. The doppler frequency at K_u-band is approximately seven times that at S-band or about ±500 kHz. Assuming the doppler has a sample rate of 500,000 samples/sec, and allowing 2 msec per search (500 bps) the equivalent of 1000 filters (whose width is 1 kHz) is provided to search the ±500 kHz frequency uncertainty. Theoretically at slower frequency search rates, say on the order of 10 msec, there are effectively 5000 filters each with an effective bandwidth of 200 Hz.

For the former case each frequency search will take 1 msec. The maximum sync time therefore (at 500 bps) is equal to a search through 2047 chips at 1/2 chip per frequency search or approximately 8 seconds.

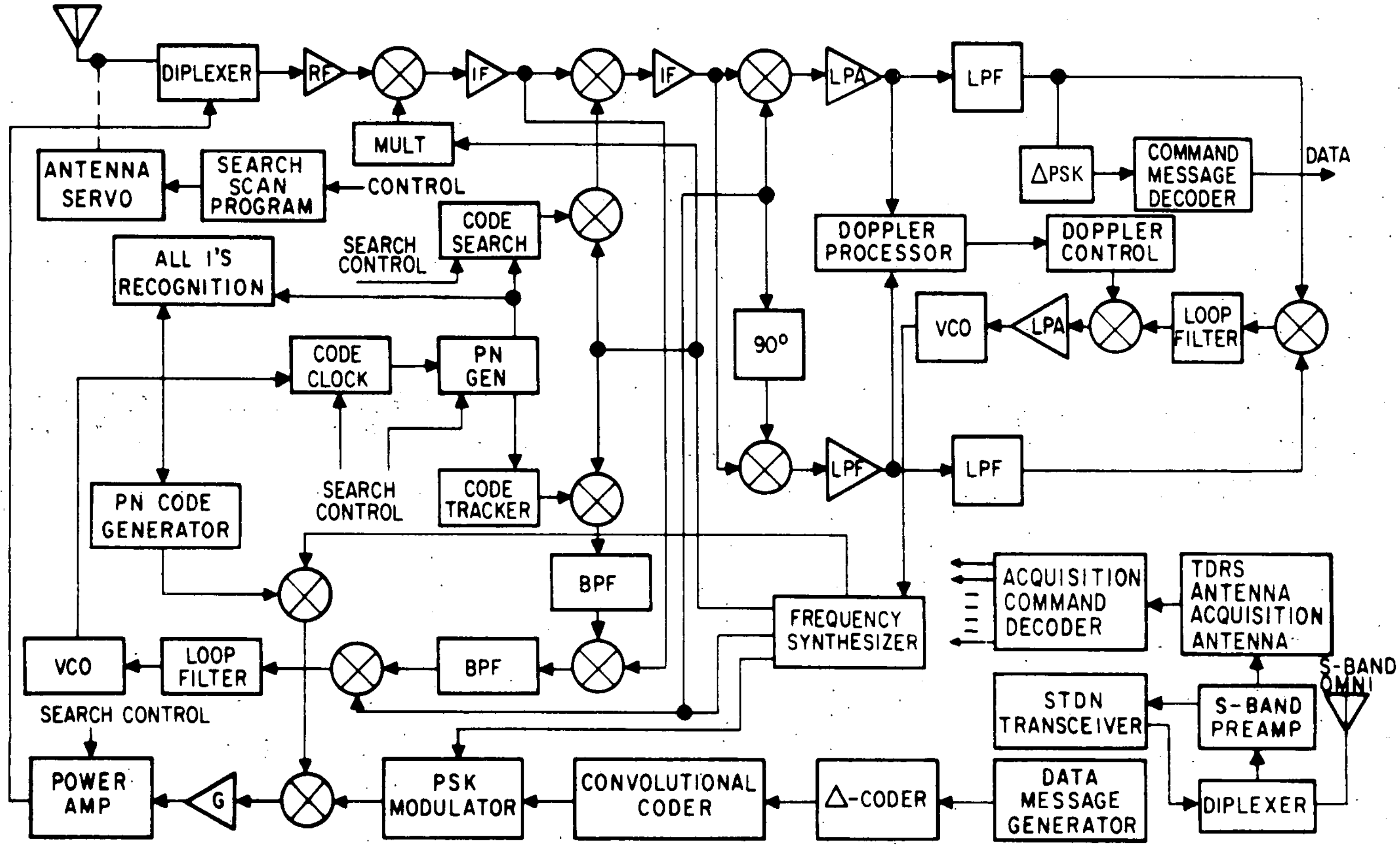


FIGURE 3-34. HDR USER TRANSPONDER

The code acquisition function is essentially the same as that used in the LDRU and MDRU case. Code acquisition operations are initiated by the TDRS ground station which repeatedly transmits a single 2047 chip PN sequence. The HDRU receiver searches the frequency uncertainty band; if a signal is not detected the PN code is retarded 1/2 and the frequency search executed again. The process is repeated until the uplinked code has been acquired and carrier lock has been established, after which the next all 1's code vector enables the return link PN generator and the power amplifier. Code acquisition by the TDRS ground station is performed in a similar operation.

As with the previously discussed user transponders, the HDRU transponder accomplishes range and range rate measurements via the 60 kchip/sec (or 60 Mchip/sec) PN code and the 5 Mchip/sec PN code in the forward and return links respectively. Assuming command rates of 1 kbps and telemetry of 10 Mbps the interleaved ambiguity resolving data word would be 133 bits and 1.33 Mbits in length in the forward and return links respectively. As mentioned, coded words of this type are employed not to achieve the specified range accuracy (determined by the PN code rate) but to resolve the 40,000 km two-way range ambiguity.

Size, power, and weight estimates have been made for the HDRU transponder having a transmit power of 32 dBm (≈1.6 watts). Assumptions relevant to hardware are the same as those previously described for the MDRU and LDRU. The HDRU transponder size and power are as follows:

Item	Size (cc)	Power (watt)
Receiver:		
RF/IF assembly	800	1
Local reference correlator	492	1
Costas loop demodulator	738	3
Doppler resolver	246	2
Coder/clock	246	1
Controller	246	1
Data processor	246	1
Frequency synthesizer	492	2
Transmitter:		
RF power amp (32 dBm)	2541	27.4
Modulator	492	2
Data processor	246	1
RF synthesizer	738	5
Total	7441	47.4

The weight distribution is: transmitter - 1.8 kilograms, receiver - 3.6 kilograms.

It should be noted that size, power, and weight estimates for the HDRU transponder do not include that for the antenna, antenna servo system, or the antenna acquisition transponder that shares an omnidirectional antenna with the STDN transponder.

3.3.3 PSEUDO-RANDOM CODE ACQUISITION AND TRACKING

Pseudo-Random code (hereafter referred to as PN code) modulation is used in the communication links to all the user satellites. In the LDR user satellites, PN modulation is used to:

1. Distribute the signal energy emanating from the TDRS to the user over a bandwidth such that the signal flux density at the earth will conform to the requirements of the IRAC (these requirements are presented in Table 3-32).

2. Discriminate against the multipath signals which will exist in the LDR case because of the omni-antennas.

3. Provide code division multiplexing from 20 users on the return link. Each of the up to 20 users will have a unique PN sequence to permit signal differentiation.

4. After synchronization has been accomplished, the two-way range information from TDRS to user is contained in the relationship between the transmitter and receiver PN code generators.

The need for PN modulation for the MDR and HDR user satellites is for conformance with the IRAC requirements, for ranging, and identification.

The use of PN modulation necessitates a synchronization procedure which must meet several requirements and is constrained by number of parameters. Synchronization is achieved in two steps: initial acquisition (i.e., synchronizing the received and reference coded to within a bit) and tracking (i.e., pulling into correlation after initial acquisition and maintaining this correlation of the two codes). Acquisition can be implemented in a number of ways and the choice is dependent primarily on the time allowed to acquire.

The time required to acquire a received PN code is a function of the amount of frequency and time uncertainty which must be resolved. The frequency uncertainty is a function of the doppler and the time uncertainty is due to the unknown transit time from transmitter to receiver. The procedure

TABLE 3-32. FORWARD LINK CLIP RATE REQUIRED TO MEET IRAC RECOMMENDATION

User Service	TDRS EIRP (dBW)	Bandwidth Requirement	Approximate PN Chip Rate
LDR	30	250 kHz	167 kchips/sec
	36	1 MHz	668 kchips/sec
	42	4 MHz	2.7 Mchips/sec
MDR	41	8 MHz	5 Mchips/sec
	47	32 MHz	20 Mchips/sec
HDR	23.6	91.4 kHz	60 kchips/sec
	53.6	91.4 MHz	60 Mchips/sec

used here is to search the total frequency uncertainty region (approximately ±12 kHz at 400 MHz) for each increment of the time uncertainty. If the signal is not present then the search is advanced to the next time increment and the frequency search repeated. This procedure is repeated until the signal is found within a particular time interval (in terms of the PN code chips) and frequency interval. The decision as to whether the signal, if found, is based on the magnitude of the signal of a given frequency cell relative to the average level (multiplied by a fixed threshold factor) over the total frequency uncertainty band.

The procedure described illustrates that the time to acquire the received code will be based on the number of uncertainty cells (both frequency and time) searched and the dwell time per cell. The number of time cells to be searched is dependent on the length of the PN sequence which is in turn dependent on the PN code rate.

3.3.3.1 FREQUENCY SEARCH

The frequency search is accomplished by the doppler processor. A functional block diagram of the processor is shown in Figure 3-30. A basic approach would be to sweep over the frequency uncertainty range. A better approach from a standpoint of saving time is a bank of parallel matched filters centered at intervals of the data rate. This, in effect is what the doppler processor does.

The doppler processor performs a digital spectral analysis of the input signal over the frequency uncertainty range. At 400 MHz the maximum doppler which will be experienced will be ±12 kHz. Assuming two modes of operation (i.e., 100 bps and 1,000 bps data rates) the doppler processor is mechanized in the following manner. At 100 bps, samples are taken for a period of 10 ms with a frequency resolution of 100 Hz and 240 frequency slots in the frequency uncertainty range. The Fourier coefficients for each of the 240 frequency slots are computed sequentially. This procedure is functionally represented by the squaring and integration block in the block diagram. As each set of coefficients are computed they are compared with the previous set and the largest set retained until all 240 sets of coefficients have been computed and compared. This largest set is then compared against an average level. A decision that the signal has been acquired is based on the peak set of Fourier coefficients exceeds the average level by some threshold factor. The value of this threshold setting will be dependent on the desired probability of acquiring probability of false lock and the bandwidth of the baseband circuitry preceding the doppler processor while operating in the acquisition mode.

Therefore, the dwell time for each frequency cell is a function of the number of computations required to compute a set of Fourier coefficients. The total time for each frequency search (over total frequency uncertainty range) is the reciprocal of the data rate or 10 ms and 1 ms corresponding to 100 bps and 1 kbps respectively.

3.3.3.2 TIME SEARCH

After each frequency search if the PN code has not been acquired the reference PN sequence is retarded by 1/2 bit and the entire process repeated until acquisition is made. The choice of 1/2 chip shifts in the reference code is made because even under the worst case phasing relationship (no worse than a 1/4 chip displacement) the cross-correlation of the received and reference PN codes will be down only 1 dB from the ideal autocorrelation peak. The steps could have been made smaller but the acquisition time would be increased proportionally.

3.3.3.3 ACQUISITION TIME

The maximum one way unknown range increment which could exist is 20,000 km. The IRAC requirement for 400 MHz is that the signal flux density on the earth shall not exceed -150 dBw/meter^2 in a 4 kHz frequency band. These requirements indicate that the 30 dBw EIRP from the TDRS to the LDR user must be uniformly distributed over 250 kHz. Using a one side receiver filter bandwidth-to-PN code chip rate ratio of 1.5, the minimum chip rate is on the order of 167 kilochips/sec. However, to take maximum advantage of available process gain a forward link chip rate of 668 kilochips/sec has been selected. At this chip rate the code length necessary for two-way unambiguous ranging over the 40,000 km (133 msec) uncertainty is on the order of 9×10^4 chips (133 msec x 668 kilochips/sec).

Assuming a 40 sec acquisition as reasonable, the curves of Figure 3-35 show that for a 100 bps data rate a 2047 chip code is the maximum length which can be used. This can be increased to a 4095 code by going to 1 chip shifts but at a 3 dB cost in required C/N_0. Given that 2047 is the maximum code length for the acquisition mode the following concept has been used to resolve the two contradictory problems, i.e., short acquisition time and large range uncertainty.

The transmission from the ground station will initially transmit a short code to the user satellite which will take a maximum of 40 secs (LDRU) to acquire. Acquisition of the received code in the user will initiate the transmitter and the PN code generator in the transmitter. Since the code sequences in the receiver and transmitter may be different, the all one's point in both the received and transmitted sequences will be synchronized to avoid a time uncertainty. The transmitted signal must be acquired in the ground station. Using the same basic acquisition procedures an additional acquisition time will be required, again a maximum of 40 seconds. Now the user and TDRS ground station are completely synchronized in both the forward and return links. Data can now be transmitted to and from the user.

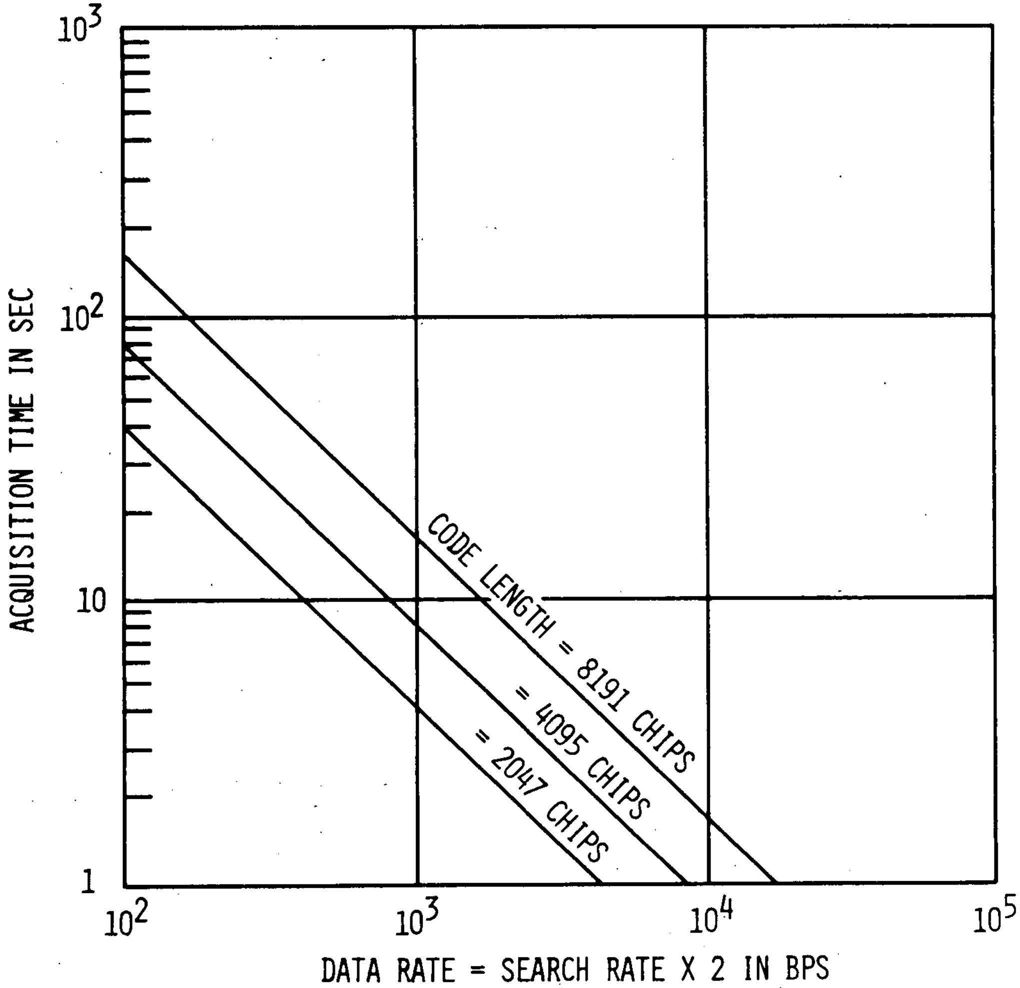


FIGURE 3-35. PN CODE ACQUISITION TIME

3.4 GROUND TERMINAL DESIGN

The TDRS system ground station is the unifying element in the overall system design. It is the source of all commands, tracking, and voice signals in the sense that all of these must pass through it and emanate from it to the two relay satellites. By the same token it will be the central gathering facility for the downlink telemetry, tracking, and voice before these are processed and sent to the users.

The TDRS ground station will be a fully automated facility. All switching, frequency selection, antenna patching, data transmission interfaces, modem selection and patching, power control, antenna pointing and tracking, etc., will be computer controlled with the need for station personnel kept to a minimum. The Central Processing Unit (CPU) used to do the controlling will also be under the command of TDRSCON. There will be a manual override mode provided so as to keep the system functioning, although, at a reduced level.

A functional design of the TDRS ground station is shown in Figure 3-36. The ground station (GS) can be divided into three functional subsystems; namely:

1. The RF group (generally at the antenna site)
2. The frequency conversion (up and down) group
3. The signal processor group (including modems, range and range rate trackers, AGIPA, etc.)

In the figure the ground station shown is that which supports only one TDRS. The total ground system is twice as detailed. The ground station design for supporting the TDRSS based on the Uprated Delta launch configuration is fundamentally no different than that design for the Part I baseline.* Deviations in the GS design to support the Uprated Delta system are in two areas. First, the GS has been modified to support the command, telemetry, and range and range rate requirements of the HDRU. Secondly, the modulation format of the TDRS/GS link has been altered to support HDR telemetry. In essence, the downlink format with exception of HDR data remains FDM/FM as in the baseline. The HDR telemetry is added to the carrier in a channelized manner. The resulting format, therefore, is a two channel system, one containing HDR data and the other consisting of LDR, TT&C, order wire, MDR, and pilot information which are modulated FDM/FM.

In the sections that follow a review of the TDRS ground station will be presented; for more detail the reader is referred to the Part I Final Report.

*Ibid. pp. 12-1 to 12-31.

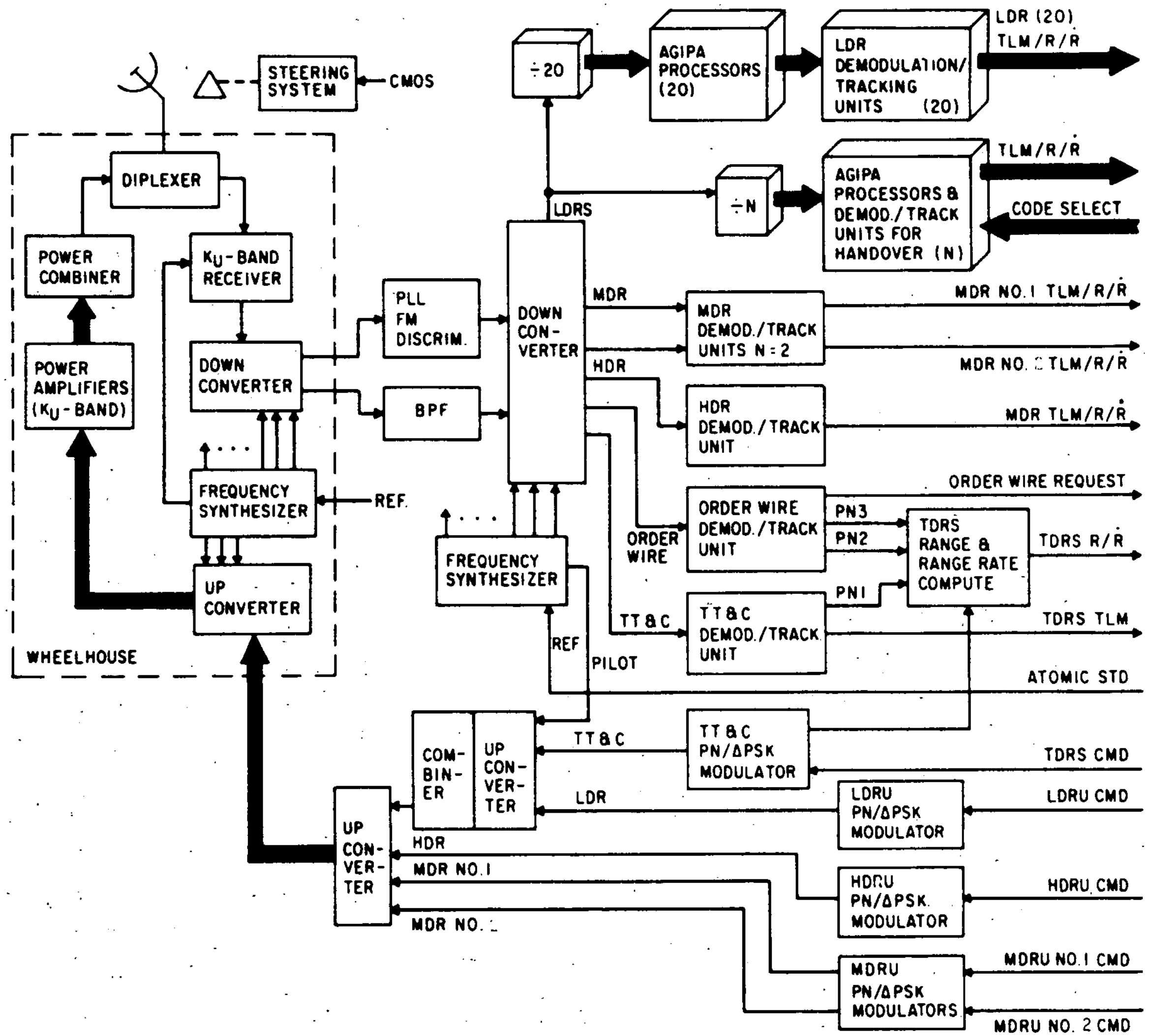


FIGURE 3-36. TDRS GROUND STATION FUNCTIONAL DESIGN: UPRATE DELTA LAUNCH CONFIGURATION

3.4.1 GROUND STATION SUPPORT REQUIREMENTS

The TDRS ground station is to provide support to the TDRS and the user spacecraft in the general areas of tracking, telemetry, and command. The ground station support requirements have not changed significantly from the baseline.

3.4.1.1 COMMAND SUPPORT

Two basic types of commands exist: commands for the relay satellites themselves, and those to be ultimately received and processed by the various user spacecraft (i.e., LDRU, MDRU, and HDRU). The commands for TDRS are generated at the ground station under the control of TDRSCON. The commands for the users can be generated at the ground station under NOCC control or can be received over the communications/data lines (NASCOM) and transmitted directly to TDRS. The source of commands in this case is the user control center.

The bit rates for the commands can range from 100 to 1000 bps for all users; therefore, normal voiceband data lines can be used for remote source commanding. The only hardware constraints foreseen are the availability of the necessary lines for remote sources and the necessary command generators for local sources. There may be a requirement to uplink video data to the HDRU; if so, modifications would be required if the data were to be uplinked directly from the user operations control center.

3.4.1.2 TRACKING SUPPORT

The need for range and range rate data on user spacecraft implies a need to know the location of both relay satellites within accuracies which yield the specified range and range rate accuracy on any user. The approach proposed is a trilateration technique for TDRS and a one code process on each user. In terms of the ground station, then, all code generation will be performed locally at the ground station by means of pseudonoise (PN) codes applied to the appropriate uplink carriers.

Due to the complexity of the tracking (i.e., the multiple relay method from ground to TDRS to user to TDRS to ground), the code processing will be accomplished by the ground station as opposed to user processing; however, provision is made to forward raw data to the user control center.

The basic deviation from the baseline ground station with regards to tracking is the change to a short PN sequence for ranging and the interleaving of a coded data word (at the CMD/TLM rates) to resolve ambiguity. Insertion of this coded word would be required infrequently (theoretically only once each pass); however, provision must be made to insert the word into the command data stream.

Each user satellite will return a code which must be processed. Since the location of the user is intimately related to the location of the relays, the range and range rate processing will have to be performed at the ground station before range and range rate counts can be transmitted to the user control center.

3.4.1.3 TELEMETRY SUPPORT

As in the case of command signals, there are two categories for the network telemetry. The first is the telemetry from the TDRS spacecraft and the second is that originating from the users. The TDRS will be transmitting housekeeping data, command verifications, etc., which must be received and interpreted by the network so as to ascertain the operating status or condition of the system. As for the users, each will upon command transmit similar telemetry.

In the case of TDRS telemetry, data will be received and handled by the ground station so that TDRSCON and TDRSNET will be able to receive display signals via communications/data lines which reflect the system condition. For the users there will be a minimum of handling/processing (except for range and range rate computation) performed at the ground station. Instead, the telemetry will be formatted for transmission to the individual users where it can be reduced per their unique requirements.

The GS telemetry support for the Uprated Delta launch configuration has been increased to handle the increased telemetry rates (greater than 1 mbps plus 100 mbps) provided by the HDRU spacecraft.

3.4.2 TDRS GROUND STATION CONFIGURATION

The elements of the TDRS ground station have not been altered significantly from the baseline system. Changes that have been made are merely to provide for the increased support required by the Part II constraints. A more detailed design of a typical TDRS ground station is presented in Appendix D.

3.4.2.1 THE RF GROUP

The TDRS GS antenna consists of two 60 ft (18.3m) diameter parabolic reflectors mounted on an elevation over azimuth wheel and track pedestal.

The antennas are high efficiency (>67%) Cassegrains using shaped reflectors and subreflectors. A horizontally stabilized RF equipment room is attached over the azimuth bearing. The RF room houses the four-horn feed assembly (including the monopulse bridge and diplexer), parametric amplifier and associated wiring and cabling.

The receiver temperature is the same as that specified in the Part I study (about 300°K as measured at the antenna input port). A receiver which corresponds to this consists of a diplexer with an input insertion loss of approximately 2 dB followed by two uncooled paramps each having a gain of 15 dB and a noise temperature of 100°K. Each paramp is backed up by a redundant paramp. Following the paramps the signal is mixed down to 2072.5 MHz, and amplified in an S-band amplifier to 0 dBm. This signal is then sent by cable to the central GS.

The transmitter consists of three power amplifiers providing 25 watts each. Two amplifiers are used for MDR #1 and MDR #2 command (of which one can be used to support HDR command). The third amplifier is a linear amplifier to transmit the two LDR, TDRS pilot reference, and the TDRS command data. The outputs are multiplexed in a multicoupler.

3.4.2.2 FREQUENCY CONVERSION GROUP

The frequency conversion group consists of the ground station FM demodulator, the frequency demultiplexer for the HDR and other channels, and the frequency division multiplexer for the uplinked signals.

The signal from the RF subsystem is filtered and amplified upon entering the main station. This signal is in two parts: a 150 MHz band for HDR telemetry and a 200 MHz band which contains six channels (i.e., MDR #1 and #2, LDR/JR, AGIPA, TT&C, order wire, and TDRS tracking) which are FDM/FM. A phase lock loop discriminator demodulates the FDM/FM into their baseband equivalents (IF) approximately 48 MHz wide.

The HDR telemetry data is bandpass filtered and applied to the frequency converters.

The LDR channels are each about 1.5 MHz and spaced 2.5 MHz apart. The LDR band is converted down in frequency, filtered, and routed to the AGIPA processors. The MDR channels are 10 MHz wide, but the information is not necessarily centered in the band. Therefore, an L.O. signal is made variable in 0.1 MHz steps so that the information in the MDR channel can be centered.

The forward link data from the ground station to TDRS contains five channels, of which three are data to be relayed to users (MDR/HDR #1, MDR/HDR #2, and LDR). The fourth channel is the pilot frequency used as a reference for the frequency synthesizers in the TDRS satellite, and the fifth is the TT&C signal. All five are FDM on the uplink carrier.

Two of the uplink channels are MDR/HDR channels for two MDR users or one MDR and one HDR. The channels are at 150 MHz initially and 1 MHz wide. The signals are mixed to a predetermined frequency with 0.1

MHz resolution anywhere within a 100 MHz bandwidth. The new frequency is determined by the particular user to be serviced. The TDRS system transmits the entire 100 MHz band for each MDR/HDR user.

3.4.2.3 SIGNAL PROCESSOR GROUP

The TDRS GS signal processor group consists of the AGIPA processor for the LDR users, Demodulation and Tracking Units (DTU) for all telemetry channels, and the PN/ΔPSK modulators for forward link commands.

3.4.2.3.1 THE AGIPA PROCESSOR

To provide polarization diversity the four-element Jr. AGIPA phased array requires eight receive channels, four each for the vertically polarized and horizontally polarized signal components. It can be shown that if the desired and interference signals are identified in the individual channels as well as after their summation sufficient information is then available to compute the weighting factors (amplitude and phase) to optimize the signal-to-interference ratio. A functional block diagram of the AGIPA processor is shown in Appendix D.

Since AGIPA processes both orthogonal polarization components of the desired and undesired signals independently, it effectively utilizes both polarization and spatial information to discriminate against the undesired signals. Detailed description of the processor has been provided in the Part I, Final Report.*

3.4.2.3.2 THE DEMODULATION AND TRACKING UNIT

The demodulation tracking unit shown in Figure 3-37 extracts an NRZ-L bit stream from the PN-ΔPSK input. It also incorporates all the necessary functions for determining the code delay and doppler shift due to the input source's range and range rate.

Several of these units and units with essentially the same objectives are used to process the various subcarriers throughout the ground station system. The following is a description of the unit depicted in Figure 3-37.

The PN-ΔPSK carrier (VHF band) is applied to the first mixer of the unit where a PN'ed local reference is mixed with it. The band pass filter isolates the difference frequency. At this point (until PN lock has been obtained) there is a noiselike waveform into the Costas loop phase detectors. The loop cannot possibly lock up to the carrier so the loop searches in vain for a predetermined period of time allotted for carrier lock.

*Ibid. pp. 4-32 to 4-51.

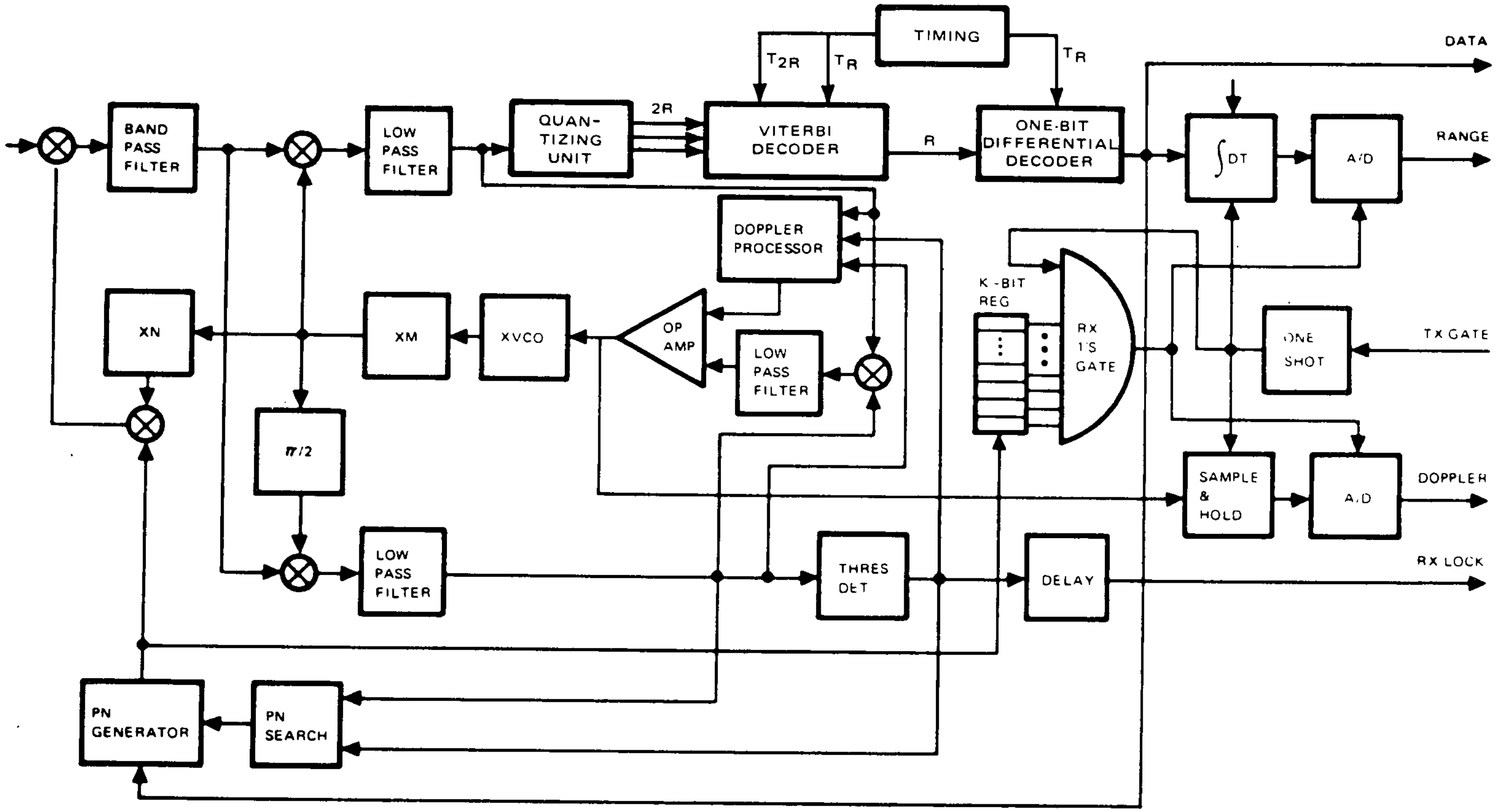


FIGURE 3-37. FUNCTIONAL BLOCK DIAGRAM OF THE DEMODULATION/TRACKING UNIT

The "PN Search" block provides a half chip delay coarse search after each carrier search time slot. At some point the local PN code comes within a half chip of perfect synchronization with the received code. The output of the band pass filter now contains sufficient carrier for the loop to lock.

In order to obtain carrier lock the "Doppler Processor" resolves the doppler shift and applies an error signal (proportional to the shift) to the "Operational Amplifier" which drives the crystal oscillator. Since the loop is now within its pull range it tracks the carrier. This having occurred, the "Threshold Detector" inhibits the "Doppler Processor, " the "PN Search, " and, after a delay sufficient to obtain fine lock on the PN code, sends an indication of the receiver lock ("RX Lock") to the transmitter.

Range delay is obtained as follows: the "RX Lock" enables the transmitter (TX) 1's gate. The first time an all one's transmitter code condition occurs the TX gate fires the "One Shot" which enables the receiver (RX) 1's gate. It also starts the integrator and causes the "Sample and Hold" to sample the error voltage to the "XVCO". When an all one's receiver code condition occurs (the receiver generator is locked to the incoming code) the integration is stopped, converted to a digital word and sent to the CPU for processing. Simultaneously the doppler error voltage is "A/D'ed" and also sent to the CPU.

The output of the Costas loop is applied to a bit quantization unit in the Viterbi decoder. Once the data has had the error control removed it is applied to a one bit differential decoder where the Delta coded bit stream is converted to NRZ.

As an added note, the "One Shot" prevents a stop count output from the RX gate unless a start count has occurred. Also the RX gate is enabled only for a period of time sufficient for maximum expected delay. This prevents accidental triggering outside the delay window.

3.4.2.3.3 THE PN/ΔPSK Modulator

The modulation unit shown in Figure 3-38 is the basic modulator for all uplink data and/or digital voice.

The NRZ-L data stream is converted to a delta coded signal, PN coded, and bi-phase modulated onto the appropriate subcarrier. The composite waveform is then filtered and the output sent to an uplink multiplexer or other uplink processor. The "TX 1's Gate" provides a start pulse for the receiver tracking circuits when all one's condition occurs in the PN code generator output.

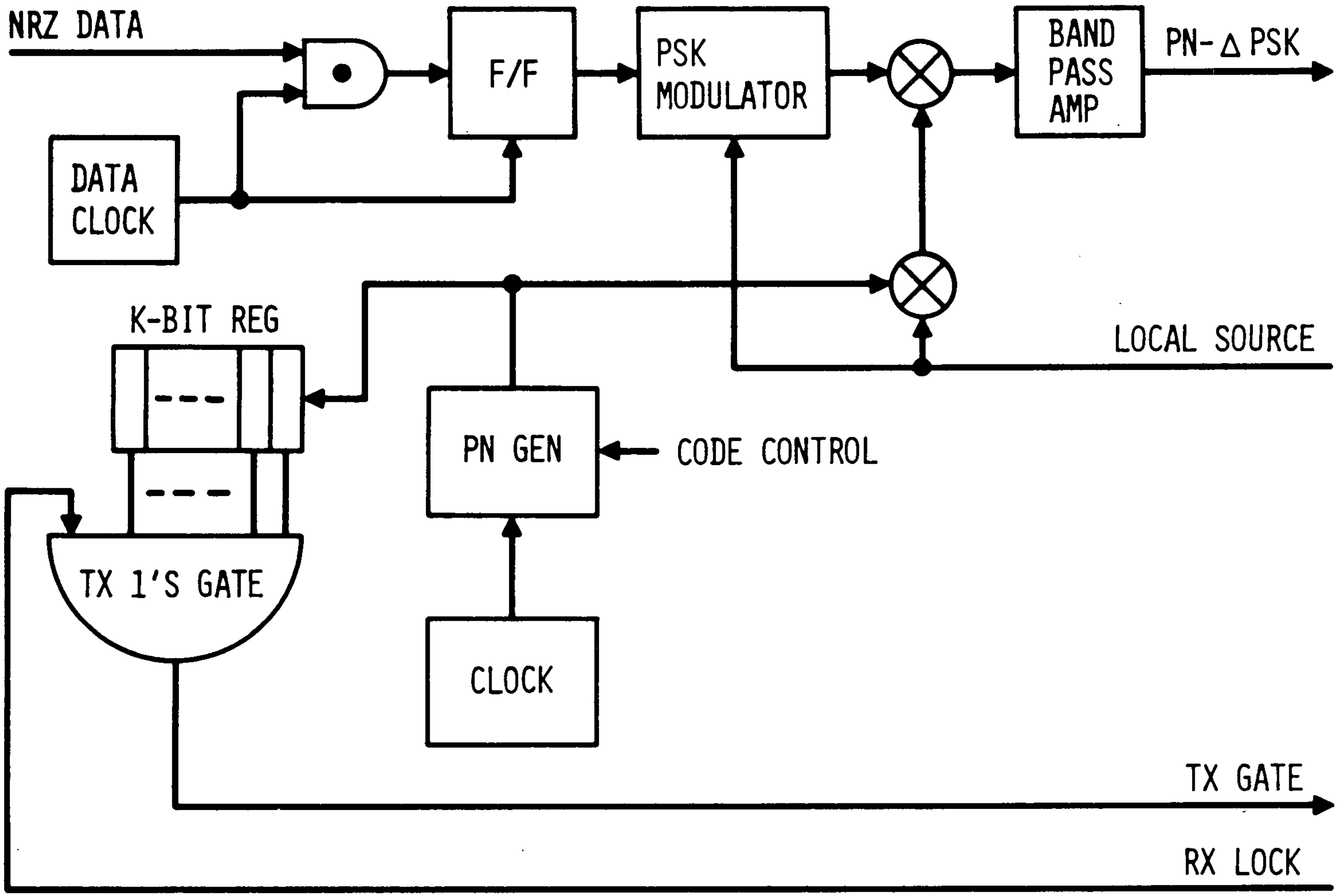


FIGURE 3-38. THE PN/ΔPSK MODULATOR

3.4.2.3.4 AGIPA PROCESSORS/DTU FOR HANDOVER

As shown in Figure 3-36 the LDR output from the down converted is channelized into two paths. One, a 20-unit group of AGIPA processors and demodulation/tracking units, the other a similar group of N units. The latter group, which conceivably can consist of only one or two units, is designed to aid in the handover of a user from one TDRS to another. Prior to handover one of the "extra" units would be assigned the code of the user of interest and would acquire that code and track along with the DTU that is dedicated to the user. The "extra" units therefore would be able to acquire the user through the TDRS that is accepting the handover, thereby reducing (or perhaps eliminating) loss of data during handover.

4.0 ATLAS CENTAUR AND SPACE SHUTTLE CONFIGURATION

The TDRS Telecommunication System designs for the Atlas Centaur and Space Shuttle launched configuration are:

- Multiple Launched Configuration where all 3 TDRS spacecraft are placed in geosynchronous orbit from a single launch. Each of the TDRS spacecraft and its Telecommunication System is the same as the Part II: Uprated Delta 2914 Configuration with a capability to support 20 LDR + 1 MDR + 1 HDR users. In this design the Telecommunication System has been constrained in performance and support capability by the size, weight and prime power limitation imposed by the Delta 2914 launch vehicle.
- Single High Performance Configuration uses the increased payload capacity of the Atlas Centaur and Space Shuttle to increase and improve the service and support to the spaceborne users.

Since the first configuration is the same as the Uprated Delta 2914 design, no discussion is included herein: only the alternate high performance system is described. The discussion is limited to the comparative difference in performance and system design, and where commonality exist with the Uprated Delta 2914 design, reference will be made to it.

The overall TDRS spacecraft antenna configuration is shown in Figure 4-1 which shows the major variation of the alternate configuration from the Uprated Delta 2914 Configuration. It is seen that the alternate configuration includes:

- LDR: A 5 element VHF array is used for Sr. AGIPA on return, and an independently mounted 5 element UHF array for the forward link

- MDR/HDR: 4 to 3.8 meters dual frequency (S-/K_u-band) antenna
- TDRS/GS: 1 to 3.8 meters antenna for the space-to-ground link

The major components of the Alternate Configuration are shown in Figure 4-2, and are as follows:

- LDR transponder
- MDR/HDR #1 transponder
- MDR/HDR #2 transponder
- MDR/HDR #3 transponder
- MDR/HDR #4 transponder
- TDRS/GS transponder
- TDRS Tracking and Order Wire transponder
- Tracking, Telemetry and Command transponder
- Frequency Source

4.1 SYSTEM DESCRIPTION

An overall detailed block diagram of the alternate TDRS Telecommunication System for the Atlas Centaur and Space Shuttle implementation is shown in Figure 4-3. The block diagram is very similar to the Uprated Delta 2914 configuration of Figure 3-11 with the exception of:

- One additional LDR-UHF transmit channel, and two LDR-VHF receive channels to support the 5th VHF and UHF antenna element
- Two additional MDR/HDR transponders
- TDRS/GS transponder changes in order to accommodate the additional LDR and MDR return data channels, and two MDR forward command data channels. The link power budget for this space-to-ground link also changes due to the larger 3.8 meter TDRS/GS antenna employed

The major difference between this alternate configuration and the Uprated Delta 2914 Configuration are described herein.

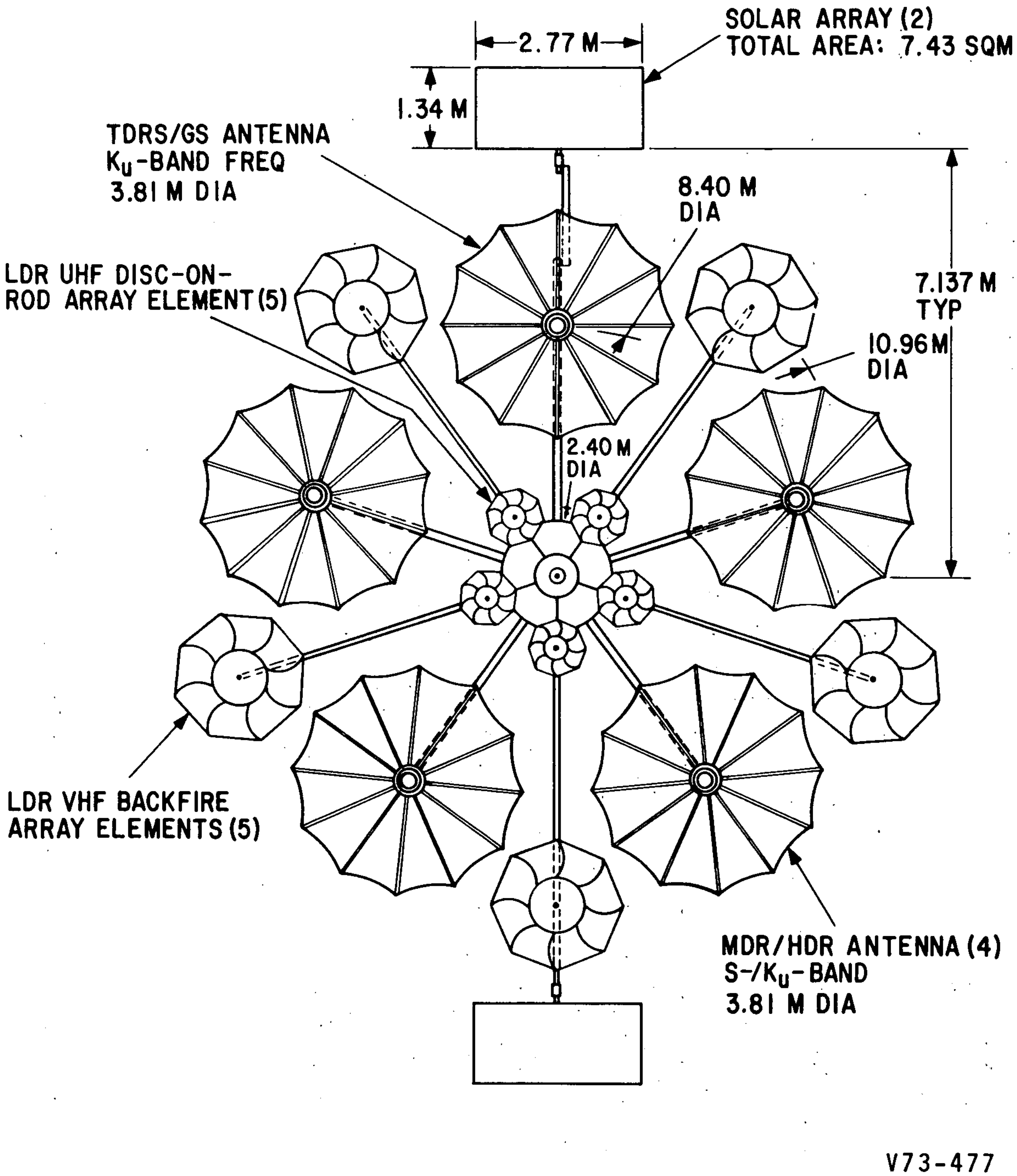
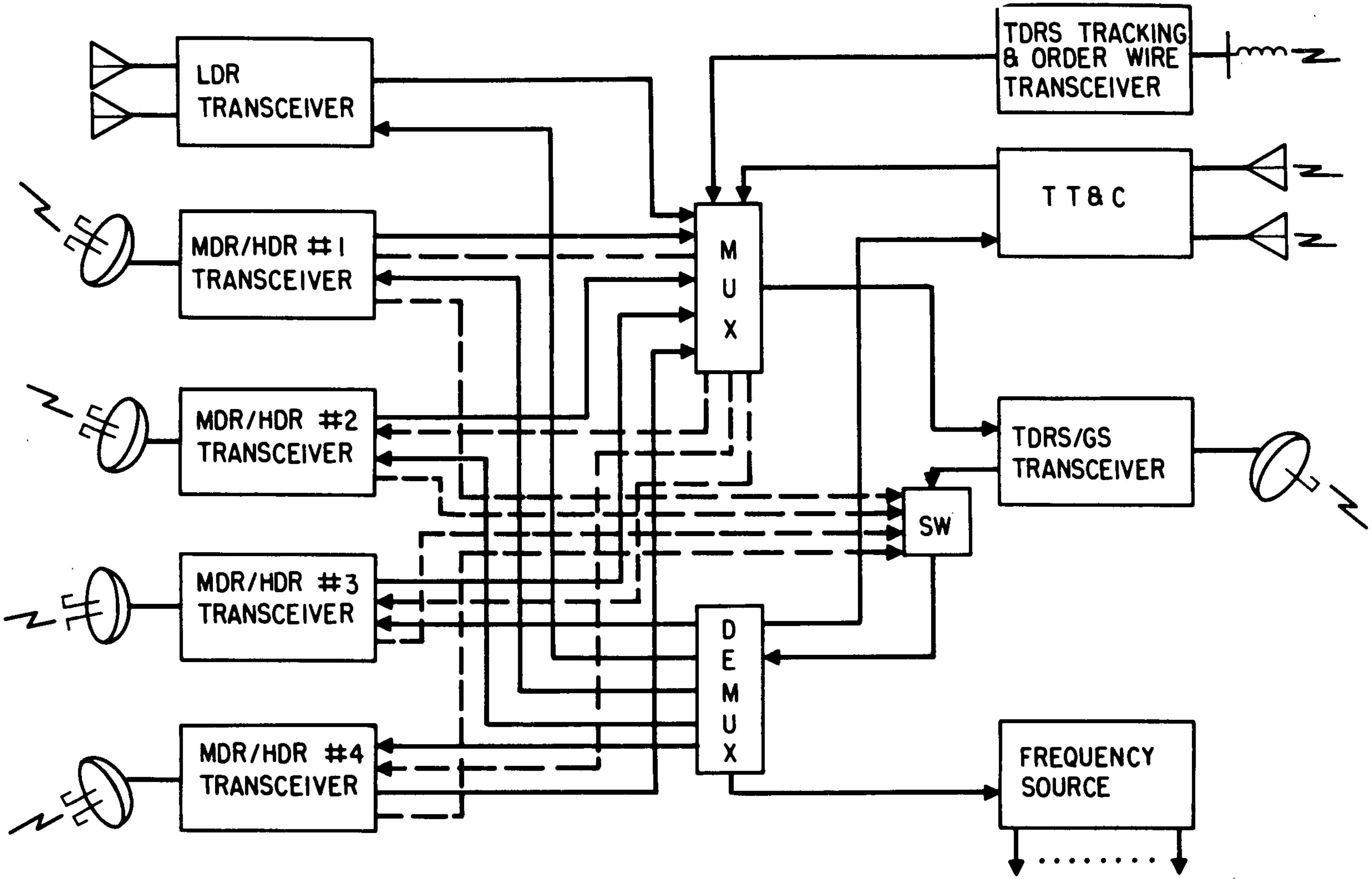


FIGURE 4-1. ALTERNATE TDRS TELECOMMUNICATION ANTENNA FARM (ATLAS CENTAUR/SPACE SHUTTLE CONFIGURATION



V73-478

FIGURE 4-2. TDRS TELECOMMUNICATION SYSTEM BLOCK DIAGRAM: ATLAS CENTAUR/SPACE SHUTTLE

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4.2 LDR TRANSPONDER

In the LDR return link, Sr. AGIPA is employed rather than Jr. AGIPA. Sr. AGIPA utilizes 5 VHF elements on a 5λ (11 meter) ring diameter, providing an approximate half power beam width (HPBW) of 8 degrees. Part I tradeoff analysis have shown the effectiveness of reduced beam width to more effectively discriminate the undesired interference signals from the desired signal, especially in the presence of high RFI environment. RFI model analysis for the Atlantic Scenario (Figure 4-4) as described in Part I for Jr. AGIPA has been extended to include Sr. AGIPA, and the resultant performance pattern for an unadapted case is shown in Figure 4-5; and for the adapted case with the user in location #2, #4, #5, and #7 are shown in Figures 4-6 through 4-9, respectively. A comparison of the performance patterns of Sr. AGIPA as compared with Jr. AGIPA for the same user locations are shown in Figure 4-10. It is evident that the reduced beam width provided with Sr. AGIPA considerably enhances its capability to discriminate against the RFI emitters. A summary of the improvement in signal-to-interference ratio (Δ SIR) as compared to a F-FOV system and Jr. AGIPA is shown in Figure 4-11 as a function of the user location. It is seen that Sr. AGIPA is more effective in providing improvements in the Δ SIR, varying typically from 9 to 15 dB as compared to 5 to 11 dB with Jr. AGIPA for the same locations. The data rate (H) that Sr. AGIPA can support as a function of the RFI power density environment is shown in Figure 4-12 as compared to the F-FOV. Typically, at RFI levels of -190 and -160 dBm/Hz, Sr. AGIPA can support 33 and 27 kbps whereas Jr. AGIPA can support 20 and 16 kbps (Figure 3-6), but the F-FOV approach is limited to 4.8 kbps and 900 bps, respectively.

Since Sr. AGIPA uses 5 elements (each element is the same short backfire element as used for Jr. AGIPA) its performance at low RFI is better than Jr. AGIPA by the approximate 1 dB array gain difference.

Referring to Figure 4-3, Sr. AGIPA utilizes a 5 element array with two receive channels per element (one each for the vertically and horizontally polarized antenna components), totalling 10 receive channels. Each receive channel and short backfire antenna element is identical to that described for the Uprated Delta 2914 Configuration and will not be described herein.

In the LDR forward link, 5 identical transmit channels are used, each of which feed a 15 dBi disc-on-rod antenna element, and is identical to that used in the Uprated Delta 2914 configuration. In the prime F-FOV mode, only one of the 5 transmit channels is used, and provides +30, +27, or +24 dBW at 26^{o} FOV as in the Uprated Delta Configuration. Since 5 identical channels are used, the 5 channels provide quintuple redundancy in the prime F-FOV mode. The 5 channels can, however, be used as a phased array in the backup mode which provides an EIRP of +44, +41, or +38 dBW; or up to

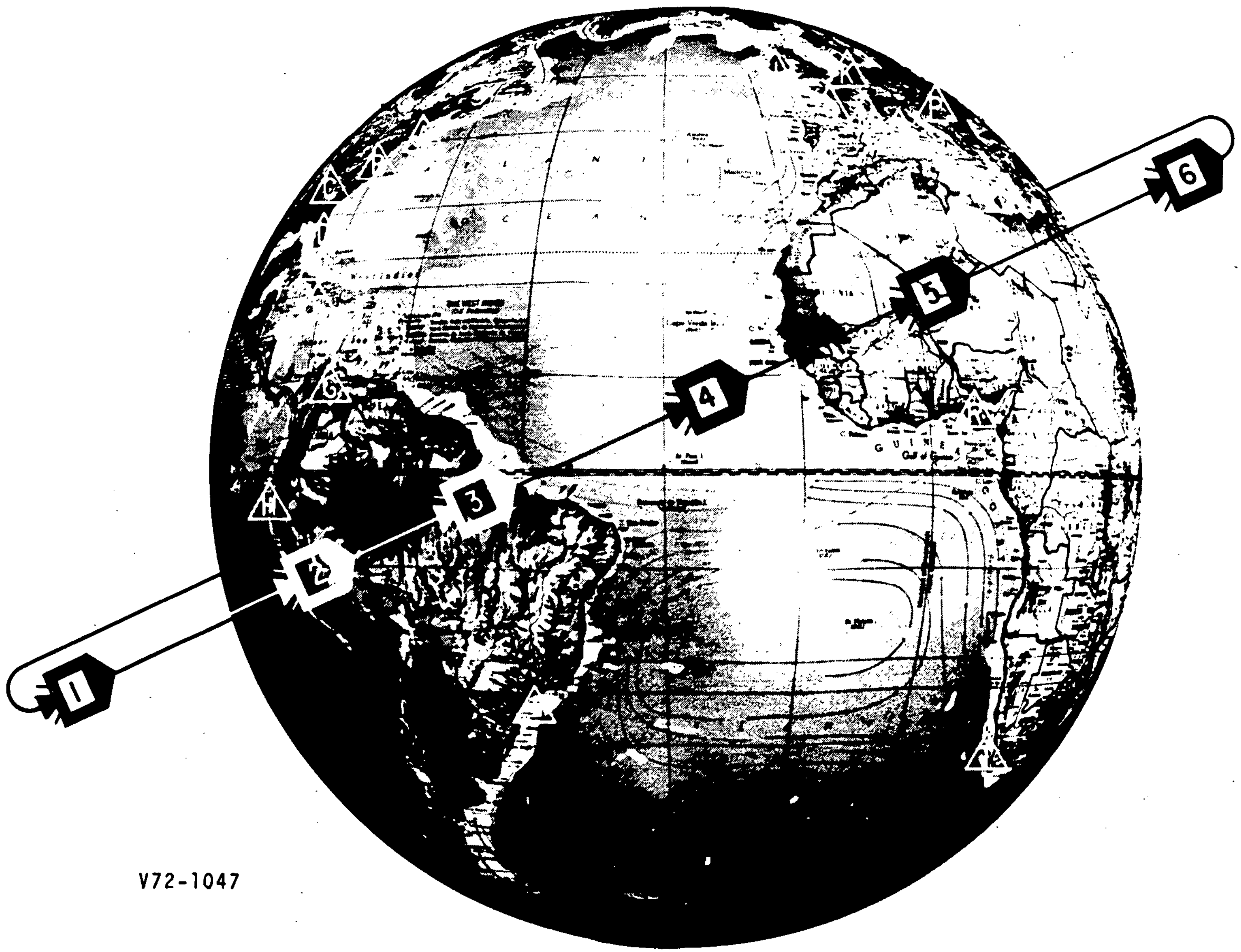


FIGURE 4-4. RFI MODEL FOR ATLANTIC SCENARIO

******************AGIPA ---- 5 ELEMENTS*****************

ATLANTIC VIEW

INITIAL SIG/INT -19.77 DB

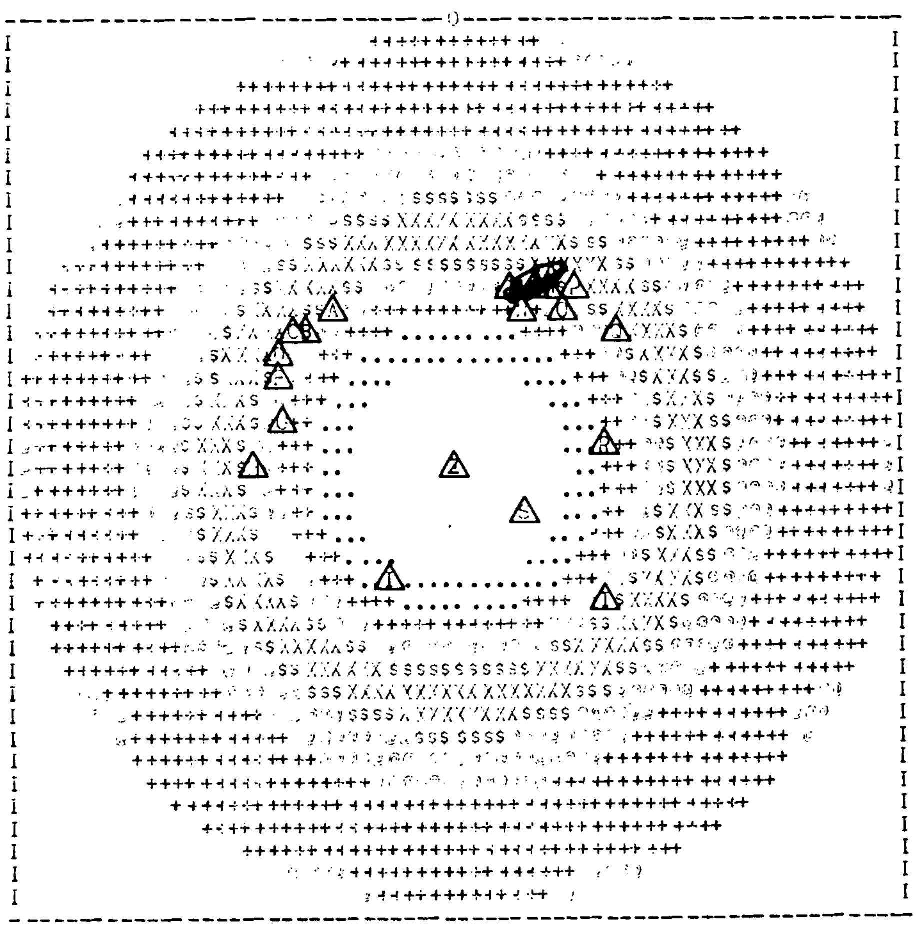
ELEMENT NUMBER		LOCATION RHO	THETA	WEIGHT MAG.	DEGREES
1	HORIZ.	2.5	0	1.	45.
	VERT.			1.	45.
2	HORIZ.	2.5	72	1.	45.
	VERT.			1.	45.
3	HORIZ.	2.5	144	1.	45.
	VERT.			1.	45.
4	HORIZ.	2.5	216	1.	45.
	VERT.			1.	45.
5	HORIZ.	2.5	288	1.	45.
	VERT.			1.	45.

SIGNAL	RHO	THETA	[illegible]-WATTS	POLAR
DESIRED	7.5	22	5	0

TOTAL RFI = 1000 WATTS

( ) = 0 TO 3DB . = 3 TO 6DB + = 6 TO 10DB
[illegible] = 10 TO 15DB $ = 15 TO 20DB X > 20DB

SCALE RADIUS 15 DEGREES



3-566

FIGURE 4-5. SR. AGIPA PATTERN: UNADAPTED

IAGIPA7 12:58:51 04/12/72

********************AGIPA---- 5 ELEMENTS************************

ATLANTIC VIEW

INITIAL SIG/INT-24.09 DB
PLUNGE IS SIG/INT-7.39 DB

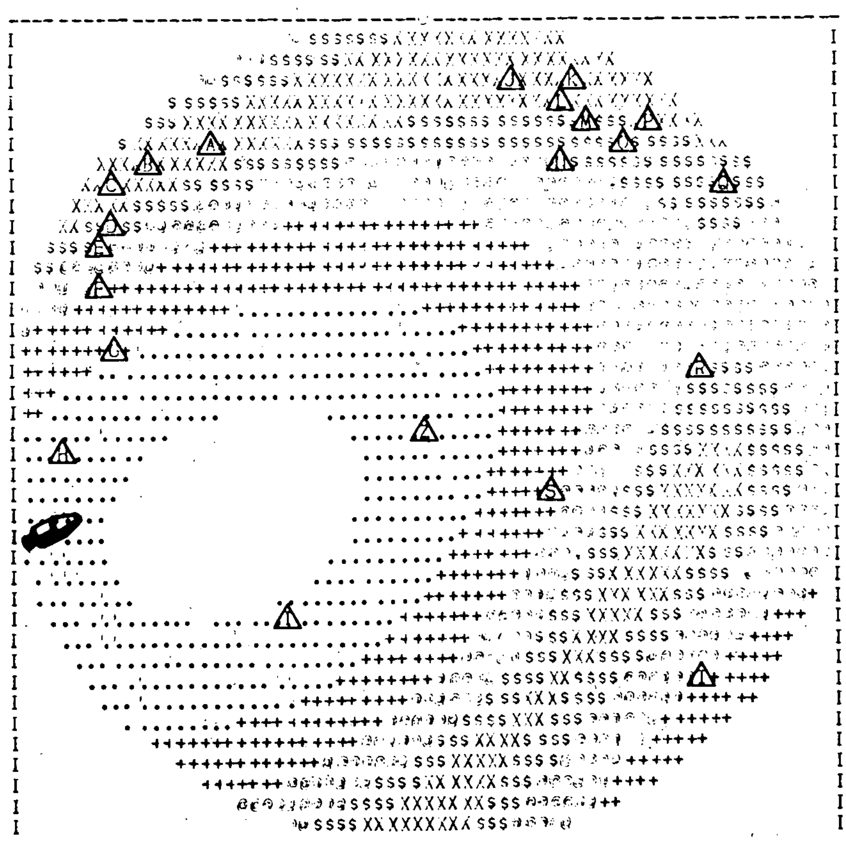
ELEMENT NUMBER		LOCATION RHO	THETA	WEIGHT MAG.	DEGREES
1	HORIZ.	2.5	0	1.125748	[illegible]
	VERT.			0.911057	45.
2	HORIZ.	2.5	72	1.02018	-2.363[illegible]
	VERT.			0.852937	45.
3	HORIZ.	2.5	144	1.0471	60.0154
	VERT.			0.[illegible]6659	45.
4	HORIZ.	2.5	216	1.34206	-87.389
	VERT.			0.320092	45.
5	HORIZ.	2.5	288	1.06415	-15.9[illegible]09
	VERT.			0.895154	45.

SIGNAL	RHO	THETA	JZA-WATTS	POLAR
DESIRED #2	8	285	5	0

TOTAL RFI = 1000 WATTS

( ) = 0 TO 3DB . = 3 TO 6DB + = 6 TO 10DB
@ = 10 TO 15DB $ = 15 TO 20DB X > 20DB

SCALE RADIUS 8 DEGREES



3-567

FIGURE 4-6. SR. AGIPA PATTERN: ADAPTED - LOCATION #2

RUN

TAGIPA7 13:07EST 04/12/72

******** ****** ****AGIPA ---- 5 ELEMENTS ****** ****** ****** ****

ATLANTIC VIEW

INITIAL SIG/INT-5.28 DB
PLUNGE 5 SIG/INT-3.69 DB

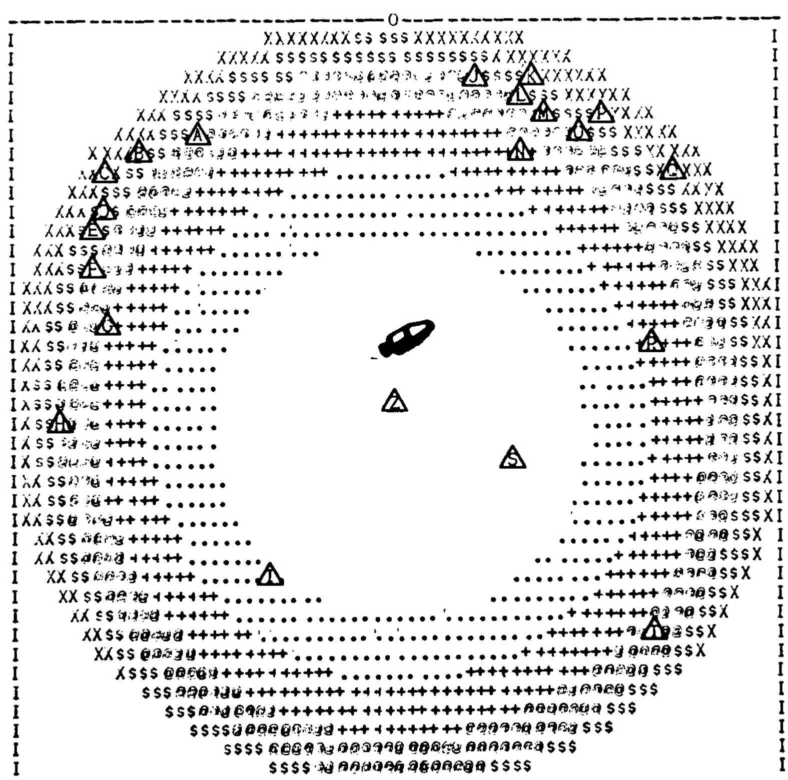
ELEMENT NUMBER		LOCATION RHO	THETA	WEIGHT MAG.	DEGREES
1	HORIZ.	2.5	0	0.911234	54.7381
	VERT.			1.02429	45.
2	HORIZ.	2.5	72	1.28122	47.6826
	VERT.			0.982337	45.
3	HORIZ.	2.5	144	0.864255	27.1618
	VERT.			0.992642	45.
4	HORIZ.	2.5	216	0.976087	40.0489
	VERT.			0.953573	45.
5	HORIZ.	2.5	288	1.09136	43.7584
	VERT.			1.01239	45.

SIGNAL	RHO	THETA	J/S-WATTS	POLAR
DESIRED #4	1.5	11	5	0

TOTAL RFI = 1000 WATTS

( ) = 0 TO 3DB . = 3 TO 6DB + = 6 TO 10DB
@ = 10 TO 15DB $ = 15 TO 20DB X > 20DB

SCALE RADIUS 8 DEGREES



3-568

FIGURE 4-7. SR. AGIPA PATTERN: ADAPTED - LOCATION #4

```
2700,_ DATA 5.8,55,5,0
RUN

TAGIPA7      13:12EST     04/12/72

**********************AGIPA ---- 5 ELEMENTS******************************

                    ATLANTIC VIEW

INITIAL SIG/INT-11.98 DB
PLUNGE 15 SIG/INT-9.7 DB
```

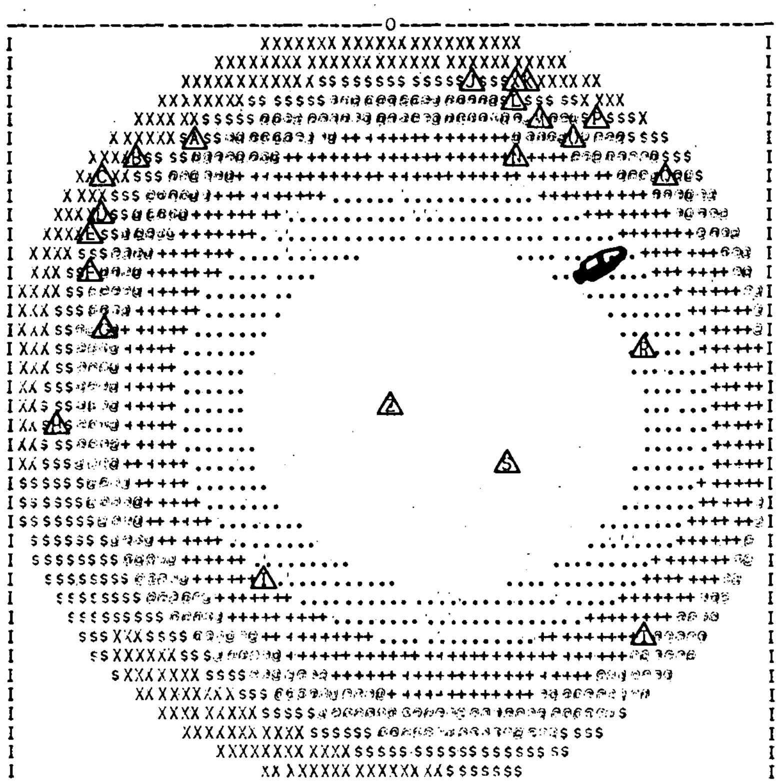
ELEMENT NUMBER		LOCATION RHO	THETA	WEIGHT MAG.	DEGREES
1	HORIZ.	2.5	0	1.2729	39.5346
	VERT.			1.07538	45.
2	HORIZ.	2.5	72	0.837461	25.7851
	VERT.			0.980777	45.
3	HORIZ.	2.5	144	1.16171	14.3439
	VERT.			1.01306	45.
4	HORIZ.	2.5	216	1.35102	63.0952
	VERT.			0.953328	45.
5	HORIZ.	2.5	288	0.629728	-68.7791
	VERT.			1.05473	45.

SIGNAL	RHO	THETA	J/N-WATTS	POLAR
DESIRED #5	5.8	55	5	0

TOTAL RFI = 1000 WATTS

```
( ) = 0 TO 3DB       . = 3 TO 6DB      + = 6 TO 10DB
@ = 10 TO 15DB       $ = 15 TO 20DB    X > 20DB
```

SCALE RADIUS 8 DEGREES



3-569

FIGURE 4-8. SR. AGIPA PATTERN: ADAPTED - LOCATION #5

```
2700 DATA 7.5,22,5,0
RUN

TAGIPA7       13:20EST     04/12/72

********************AGIPA ---- 5 ELEMENTS**************************

                    ATLANTIC  VIEW

INITIAL SIG/INT-19.77 DB
PLUNGE 15 SIG/INT-13.93 DB
```

ELEMENT NUMBER		LOCATION RHO	THETA	WEIGHT MAG.	DEGREES
1	HORIZ.	2.5	0	0.467076	20.2355
	VERT.			0.93798	45.
2	HORIZ.	2.5	72	1.77439	19.3756
	VERT.			0.793222	45.
3	HORIZ.	2.5	144	0.097423	-43.0098
	VERT.			0.602314	45.
4	HORIZ.	2.5	216	0.36908	27.1492
	VERT.			0.7931	45.
5	HORIZ.	2.5	288	2.41422	17.3578
	VERT.			0.913098	45.

SIGNAL	RHO	THETA	J/N-WATTS	POLAR
DESIRED #7	7.5	22	5	0

```
TOTAL RFI =  1000 WATTS

( ) = 0 TO 3DB      . = 3 TO 6DB      + = 6 TO 10DB
@   = 10 TO 15DB    $ = 15 TO 20DB    X > 20DB

  SCALE RADIUS 8 DEGREES
```

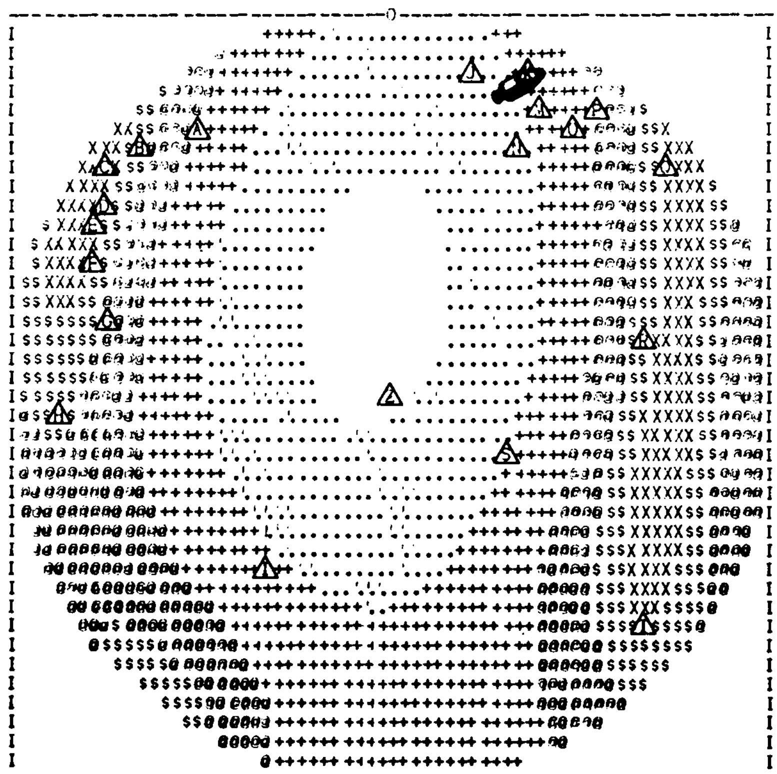


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FIGURE 4-9. SR. AGIPA PATTERN: ADAPTED - LOCATION #7

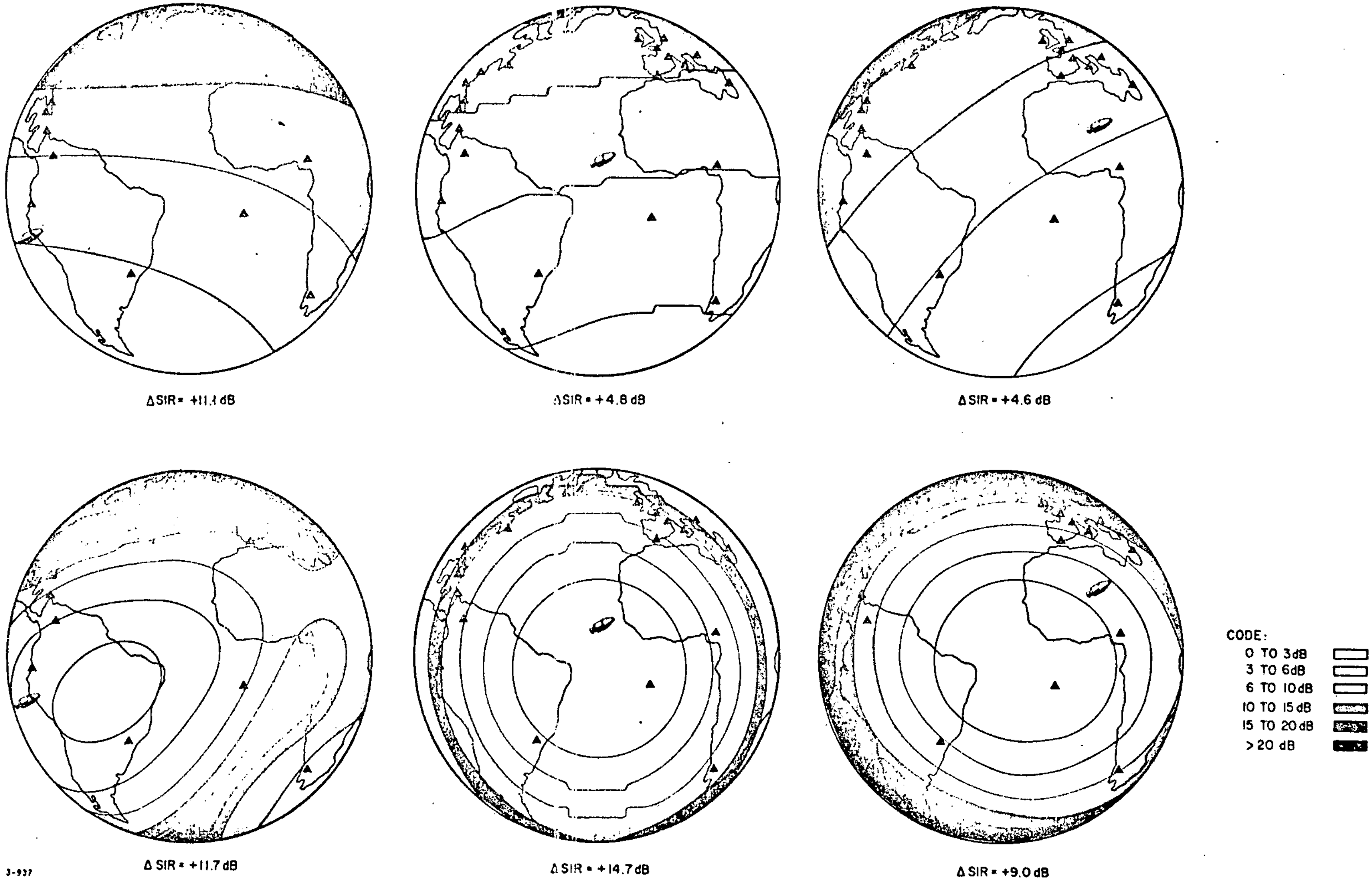

FIGURE 4-10. SR AGIPA VS JR AGIPA (ATLANTIC SCENARIO)

FIGURE 4-10

FOLDOUT FRAME 1

FOLDOUT FRAME 2

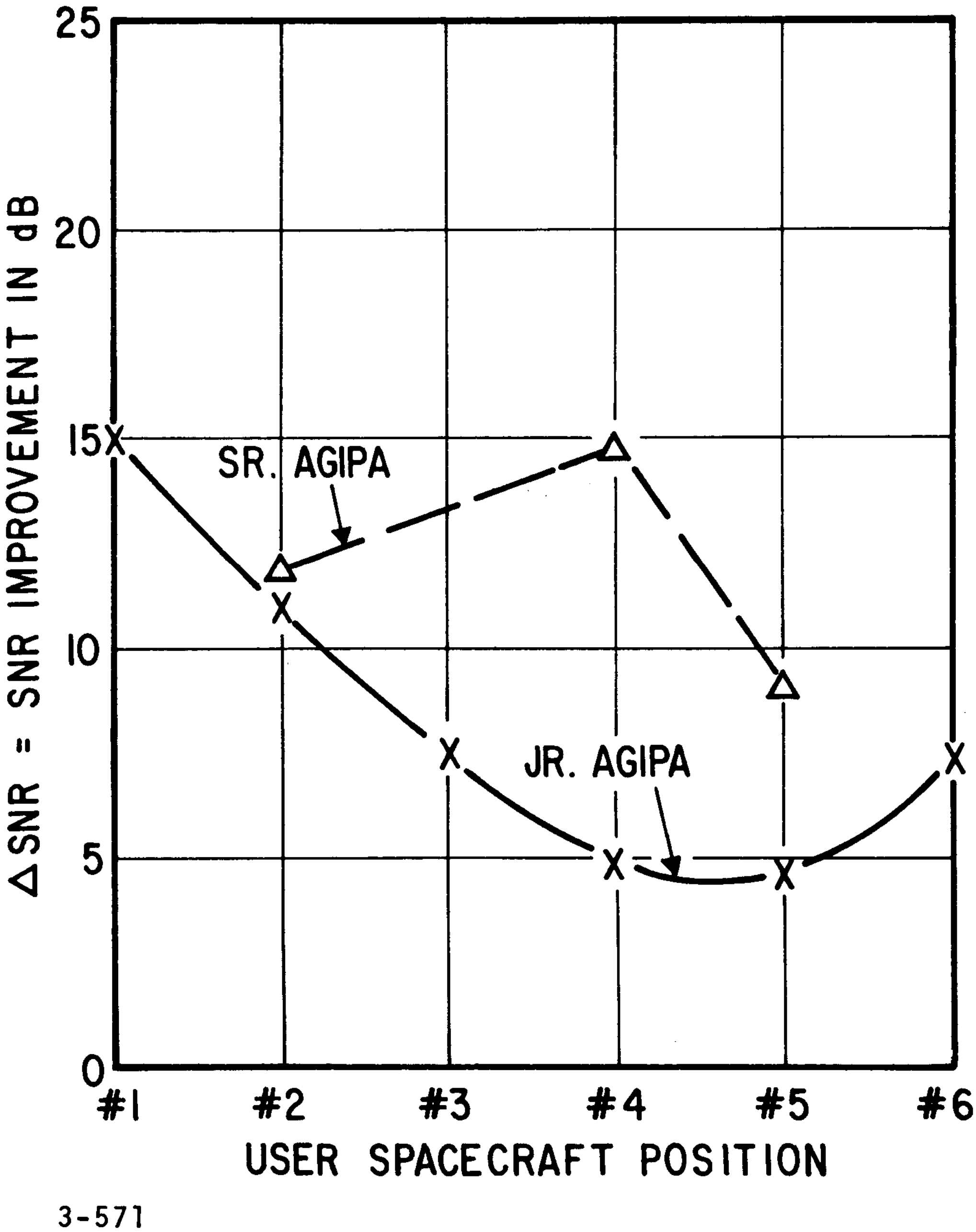

FIGURE 4-11. SNR IMPROVEMENT: ATLANTIC SCENARIO

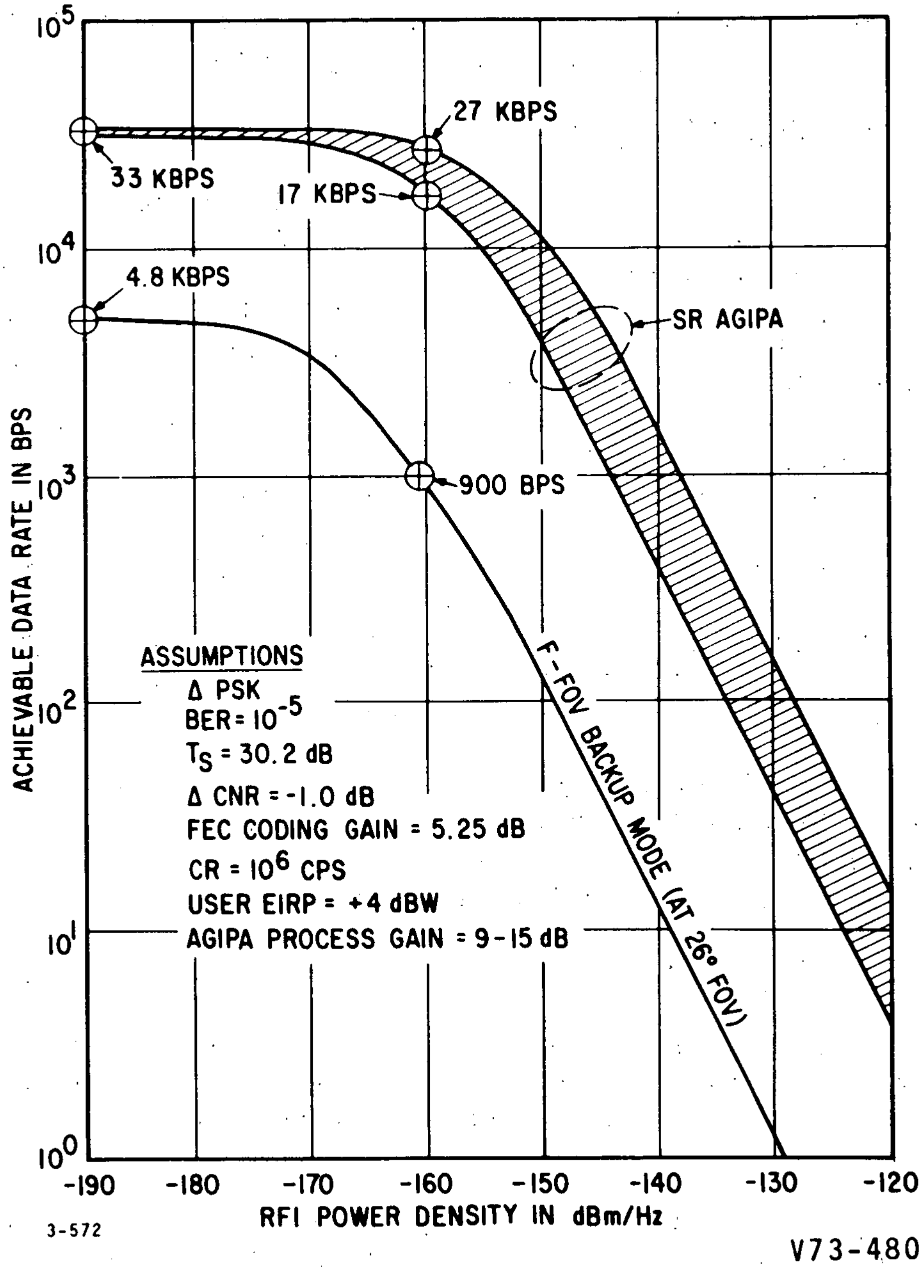

FIGURE 4-12. LDR RETURN LINK: DATA RATE VERSUS RFI POWER DENSITY (ATLAS CENTAUR/SPACE SHUTTLE)

+14 dBW of additional EIRP to reach a user in an emergency condition (e.g., in a tumbling state), or when he is in the midst of high RFI environment. The forward link data rate performance versus RFI as shown in Figure 4-13 is improved by +2 dB over that shown for the Uprated Delta Configuration.

Since each channel is the same as for the Uprated Delta Configuration, detailed description will not be repeated here.

4.3 MDR/HDR TRANSPONDER

Referring to Figure 4-3 the 4 MDR/HDR transponders are identical to that used for the Uprated Delta 2914 Configuration. Each transceiver is a dual frequency S-/K_u-band system, feeding a 3.8 meter deployable parabolic reflector antenna. As in the Uprated Delta Configuration, each transceiver is designed to receive simultaneously S- and K_u-band data from a common MDR/HDR user but can only transmit either S- or K_u-band. Although the 4 transponders can support up to 4 MDR users simultaneously, the TDRS/GS space-to-ground link is currently designed to support only 1 HDR link as required. As a future growth consideration, however, this space-to-ground link could be redesigned to support more than 1 HDR user by adding additional HDR transmit channels in the TDRS/GS transponder.

As in the Uprated Delta Configuration, two transponders (MDR/HDR #3 and #4) provide a functional redundancy to the TDRS/GS transponder, such that triple functional redundancy exists for the TDRS/GS transponder.

Since these two transponders are the same as that described for the Uprated Delta MDR/HDR Transponder, a description will not be repeated herein. MDR/HDR #1 and #2 are the same with the exception that the TDRS/GS backup functions have been removed.

4.4 TDRS/GS TRANSPONDER

The TDRS/GS transponder has been modified to include the additional return and forward link data that must be relayed to and from the Ground Station, and the larger 3.8 meter antenna that is used in this alternate configuration.

The TDRS/GS transmitter is still a dual channel system; however, the FDM/FM channel has been replaced with an FDM channel. In effect the TDRS/GS transmitter is an all FDM system, where the HDR data is still transmitted on a separate HDR channel employing a power amplifier operating in the saturated mode, and all other remaining FDM data (10 LDR + 4 MDR + Telemetry + TDRS Tracking + Order Wire) are combined and transmitted in a linear solid state amplifier channel. These transmit channels are referred to herein as the HDR and FDM channels.

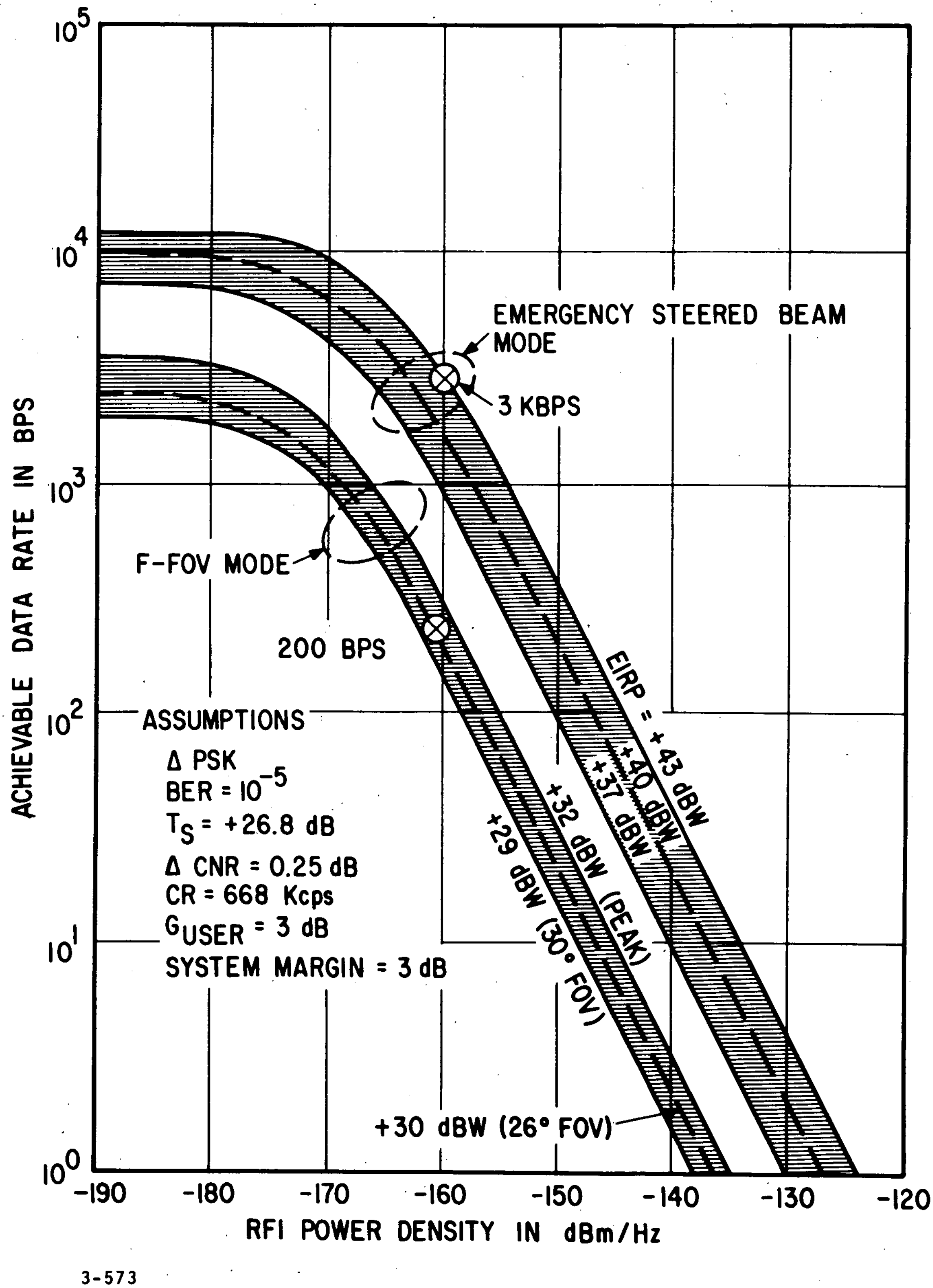

FIGURE 4-13. LDR FORWARD LINK: DATA RATE VERSUS RFI POWER DENSITY (ATLAS CENTAUR/SPACE SHUTTLE)

The FDM channels are combined as depicted in Table 4-1 to form the composite baseband signal which has a maximum frequency component at 85 MHz. Frequency modulation cannot be used because the RF bandwidth limitation of 200 MHz (see Figure 3-3) can only support an FM index of less than 0.2, a value that eliminates the effectiveness of FM.

Link requirements for the HDR channel and one MDR channel are shown in Table 4-2. The parameters used in preparing this table are the same as used in the preparation of Table 4-1, which summarizes the power requirement for all the FDM channel components. The total power of 0.15 watt for the FDM signal must be generated in a linear amplifier to reduce intermodulation distortion to acceptable limits. Typically, this means a 10 dB backoff from saturation. Therefore, the final amplifier for the FDM signal must have a saturated power output capability of 1.5 watts, within the capability of solid state amplifiers. The HDR signal requires +3.7 dBW, , or 2.4 watts, to be developed. However, this signal can be amplified in a saturated amplifier because only one signal is present. This amplifier is also solid state, a change from the TWT employed in the Uprated Delta configuration. The larger TDRS/GS antenna is largely responsible for this change.

The weighting function required for the FDM signal of the Atlas Centaur/Space Shuttle Configuration follows the power output allocation shown in Table 4-1. This weighting function is implemented in the Weighted Combiner shown in Figure 4-3. Following the combiner is an image reject mixer used to upconvert the 1 to 85 MHz spectrum to the 461 to 545 MHz region. This type of mixer is required to circumvent an impossible filter requirement. A 10 dB gain band-pass amplifier follows and drives a second mixer where upconversion to K_u-band occurs. The required 21.9 dBm output power is obtained by amplification in a linear solid state amplifier providing 39 dB of gain.

The HDR channel of the TDRS/GS transmitter differs from the Uprated Delta version in that the TWTA is replaced with a solid-state amplifier. The reduced gain (44 dB) and power output (2.4 watts) make this possible.

Figure 4-14 depicts the spectrum transmitted from the Ground Station and received by TDRS/GS receiver. Double conversion is employed to translate this K_u-band signal to the baseband region for separation and distribution to the appropriate sections of the spacecraft. A block diagram of this receiver is shown in Figure 4-3. It is identical to that of the Uprated Delta Configuration with the addition of two MDR/HDR channels. Therefore, a description of the receiver will not be repeated herein. Table 4-3 contains the Power Budget for this link. Only one MDR channel is shown because all four channels are identical.

TABLE 4-1. FDM BASEBAND LAYOUT

Function	Channel Bandwidth (MHz)	Required S/N (dB)	F_L (MHz)	F_H (MHz)	F_n (MHz)	Required(1) P_T (dBm)	(mW)
MDR No. 1	10	10	1	11	6	14.8	30.2
Order Wire	1	10	13	14	13.5	4.8	3.0
TT&C	1	10	14.5	15.5	15.0	4.8	3.0
TDRS Tracking	1	10	16.0	17.0	16.5	4.8	3.0
LDR No. 1	2	6	18.5	20.5	19.5	3.8	2.4
LDR No. 2	2	6	21.0	23.0	22.0	3.8	2.4
LDR No. 3	2	6	23.5	25.5	24.5	3.8	2.4
LDR No. 4	2	6	26.0	28.0	27.0	3.8	2.4
LDR No. 5	2	6	28.5	30.5	29.5	3.8	2.4
LDR No. 6	2	6	31.0	33.0	32.0	3.8	2.4
LDR No. 7	2	6	33.5	35.5	34.5	3.8	2.4
LDR No. 8	2	6	36.0	38.0	37.0	3.8	2.4
LDR No. 9	2	6	38.5	40.5	39.5	3.8	2.4
LDR No. 10	2	6	41	43	42.0	3.8	2.4
MDR No. 2	10	10	45	55	50	14.8	30.2
MDR No. 3	10	10	60	70	65	14.8	30.2
MDR No. 4	10	10	75	85	80	14.8	30.2
					Total Power =		153.8 mW (+21.9 dBm)

(1) RF power required at final power amplifier and includes internal losses (diplexer, combiner, switches, waveguide, etc.) of 2.6 dB.

TABLE 4-2. TDRS/GS RETURN LINK BUDGET

Parameter	FDM[1]	HDR
CNR required, dB	10	17.1
RF Bandwidth, MHz	10	150
System Noise Temperature[2], dB	25.2	25.2
Thermal Noise Power, dBW	-133.4	-121.6
GS Antenna Gain[3], dBi	67.9	67.9
α_T = space, dB	-208.1	-208.1
pointing + polarization, dB	- 1.6	- 1.6
Rain Margin, dB	17.5	17.5
EIRP, dBW	35.8	54.7
TDRS Antenna Gain, dBi	53.6	53.6
TRANSMIT RF POWER[4]		
• High Power Mode, dBW	- 15.2	+ 3.7
• Low Power Mode, dBW	- 15.2	- 6.3

NOTES:

1. Computed for MDR video. See Table 4-1 for other components of FDM signal.
2. Two uncooled paramps in cascade
3. 18.3 meter diameter antenna
4. Includes internal losses (waveguide, diplexer, switch, etc.) of 2.6 dB.

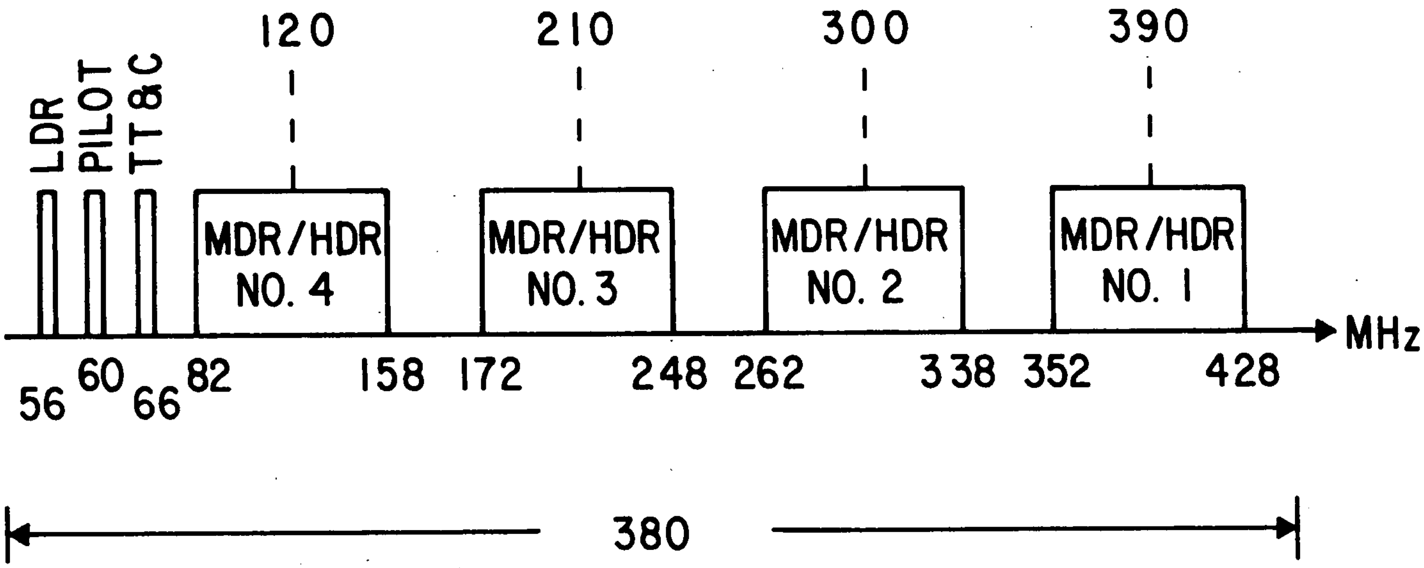

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FIGURE 4-14. TDRS/GS FORWARD LINK: FREQUENCY ASSIGNMENT

TABLE 4-3. GROUND STATION-TO-TDRS FORWARD LINK BUDGET

Parameter	LDR	MDR Unmanned	MDR Manned	HDR Unmanned	HDR Manned
Modulation	ΔPSK	ΔPSK	ΔPSK	Δ PSK	ΔPSK
Data Rate, kbps	1.0	1.0	54.0	1.0	10^4
CNR Required (1), dB	12.34	3.52	8.64	-7.66	16.16
RF Bandwidth, MHz	1.0	75.0	75.0	100.0	100.0
System Noise Temp. (2), dB	33.6	33.6	33.6	33.6	33.6
Thermal Noise Power, dBW	-135.0	-116.2	-116.2	-115.0	-115.0
TDRS Antenna Gain, dBi	52.8	52.8	52.8	52.8	52.8
Losses: Space, dB	-207.2	-207.2	-207.2	-207.2	-207.2
Pointing, dB	-1.0	-1.0	-1.0	-1.0	-1.0
Polarization, dB	-0.5	-0.5	-0.5	-0.5	-0.5
Atmospheric, dB	-0.4	-0.4	-0.4	-0.4	-0.4
Rain Margin (3), dB	17.5	17.5	17.5	17.5	17.5
EIRP Required, dBW	51.1	61.1	66.2	51.1	75.4
GS Antenna Gain (4), dBi	67.0	67.0	67.0	67.0	67.0
GS Transmitter Power (5) Required, dBW	-15.9	-5.9	-0.8	-15.9	+8.4
Watts	25.7	0.26	0.83	25.7	6.92

NOTES:

1. See Figure 3-11 for the required CNR for the GS/TDRS link.
2. Uses mixer front end (NF = 7.5 dB); T_{ant} = 290 K, losses = 1.5 dB, T_s = 33.6 dB.
3. Rain margin provides operation in 25 mm/hr rainfall rate.
4. 18.3 meter aperture with seventy-five percent efficiency.
5. Includes internal losses (waveguide, diplexer, switch, etc.) of 2.1 dB.

4.5 SUMMARY OF MODES OF SERVICE

Table 4-4 contains a summary of the Modes of Service provided by the Atlas Centaur/Space Shuttle Configuration together with the significant parameters for each mode.

4.6 WEIGHT AND POWER SUMMARY

An overall weight and power summary for the Alternate TDRS Telecommunication System for launch on the Atlas Centaur and Space Shuttle launch vehicle is shown in Table 4-5. The overall Telecommunication System weighs 230 kg.

TABLE 4-4. TDRS TELECOMMUNICATION: MODES OF SERVICE: ATLAS CENTAUR/SPACE SHUTTLE

Modes	Transmit EIRP at 26^{o} FOV	Receive G/T_s at 26^{o} FOV
LDR Link		
• Primary Mode	• F-FOV = +30/27/+24 dBW (command 1 at a time)	• Sr. AGIPA = -12.7 dB/K (support 20 simultaneously)
• Backup Mode	• Steered Beam = +44/+41/+38 dBW	• F-FOV = -19.7 dB/K
MDR/HDR #1, #2, #3 and #4		
• Primary Mode	• Support 4 MDR or 3 MDR + 1 HDR	• Support 4 MDR or 3 MDR + 1 HDR
	S-band: Unmanned = +41 dBW Manned = +47 DBW	+10.0 dB/K
	K_u-band: Unmanned = +23.6 dBW Manned = +53.6 dBW	+25.9 dB/K
• Backup Mode (#3 and #4 only)	• K_u-band = +53.6 dBW	+25.9 dB/K
	S-band = +47.0 dBW	+10.0 dB/K
TDRS/GS Link		
• Primary Mode	• FDM channel = +42.9 dBW HDR channel = +54.7 dBW	+19.2 dB/K
• Backup Mode		
K_u-band	+53.6 dBW	+25.9 dB/K
S-band	+47.0 dBW	+10.0 dB/K
VHF	+7.7 dBw	-30.2 dB/K

TABLE 4-5. TDRS TELECOMMUNICATION SYSTEM: WEIGHT AND POWER SUMMARY
(Part II - Atlas Centaur/Space Shuttle Configuration)

COMPONENT	WEIGHT (KG)	PRIME POWER (Watts) PEAK	PRIME POWER (Watts) AVE
1. LDR			
• Receiver	5.1	11.4	11.4
o Transmitter	3.2	503 403 205	106 56 32
• Antenna & Support Booms	45.7		
2. MDR HDR #1			
o Receiver	4.1	7.6	7.6
o Transmitter	6.1		
*S-band		66.0 15.0[3]	66.0 15.0[3]
*Ku-band		35.8 5.1[4]	35.8 5.1[4]
• Antenna & Support Booms	20.1	24.0	4.0
3. MDR HDR #2			
• Receiver	4.1	7.6	7.6
o Transmitter	6.1		
*S-band		66.0 15.0[3]	66.0 15.0[3]
*Ku-band		35.8 5.1[4]	35.8 5.1[4]
• Antenna & Support Booms	20.1	24.0	4.0
4. MDR HDR #3			
• Receiver	4.5	8.2	8.2
o Transmitter	6.4		
*S-band		66.0 15.0[3]	66.0 15.0[3]
*Ku-band		35.8 5.1[4]	35.8 5.1[4]
• Antenna & Support Booms	20.1	24.0	4.0
5. MDR HDR #4			
o Receiver	4.5	8.2	8.2
o Transmitter	6.4		
*S-band		66.0 15.0[3]	66.0 15.0[3]
*Ku-band		35.8 5.1[4]	35.8 5.1[4]
• Antenna & Support Booms	20.1	24.0	4.0
6. TDRS GS			
• Receiver	2.2	5.3	5.3
o Transmitter	6.5	77.0[5]	77.0[5]
o Antenna & Support Booms	19.0		
7. Frequency Source	3.5	8.0	8.0
8. TT&C			
o Processor	6.0	10.0	10.0
o Transceiver	1.8	13.5 4.5	0.5[6]
o Antenna (2)	1.4		
9. TDRS Tracking/Order Wire			
o Transponder	2.5	7.9	2.0[7]
• Antenna	0.1		
10. Cabling/Wires/Waveguide	12.0		
TOTAL	231.6		

NOTES:

1. Emergency steered Beam mode provides EIRP of +44, +41, or +38 dBw.
2. F-FOV mode provides EIRP of +30, +27, or +24 dBw at 26° FOV.
3. S-band mode emits EIRP of +47 and +41 dBw to support manned and unmanned user, respectively.
4. Ku-band mode emits EIRP of +53.6 and +23.6 dBw to support video to manned user and 1 kbps to unmanned user, respectively.
5. With 17.5 dB rain margin.
6. Transmitter normally turned off "on-station".
7. Transmitter turned on only periodically; receiver is always on.

4.7 USER TERMINAL AND GROUND STATION IMPACT

The impact of the Space Shuttle/Atlas Centaur Launch Configuration on the user spacecraft telecommunications terminal design and the TDRS ground station design is discussed in this section. The service provided by this maximum capability system has been extended to support 4 simultaneous MDR users simultaneously in addition to the services previously provided. In addition, the return link LDR support is via a Sr. AGIPA concept (i.e., 5 element arrays).

4.7.1 USER TRANSPONDER IMPACT

The impact of the Shuttle/Atlas Centaur configuration on the user spacecraft transponder (regardless of the type of service) is virtually nil. With the exception of increased forward link EIRP in the LDRU emergency mode, the forward link support on a per user basis is unchanged.

4.7.2 THE TDRS GROUND STATION

The impact of this maximum capability service on the TDRS ground station is in two areas. First the PLL/FM discriminator used in the up-rated Delta system has been eliminated as shown in Figure 4-15. The down converted signal (at S-band) is applied directly to an FDM demultiplexer, whose output is channeled to the demodulation/tracking units for each of the users. In the MDR/HDR telemetry channels the ground station can support simultaneoulsy in the telemetry link either four MDR users or three MDR plus one HDR users.

In the forward link the various user command channels are combined in FDM in a manner similar to that in the ground station designed to support the up-rated Delta system. The maximum capability system up links nine channels in FDM. There are four MDR channels and one each for HDR, LDR, order wire, TT&C, and the pilot.

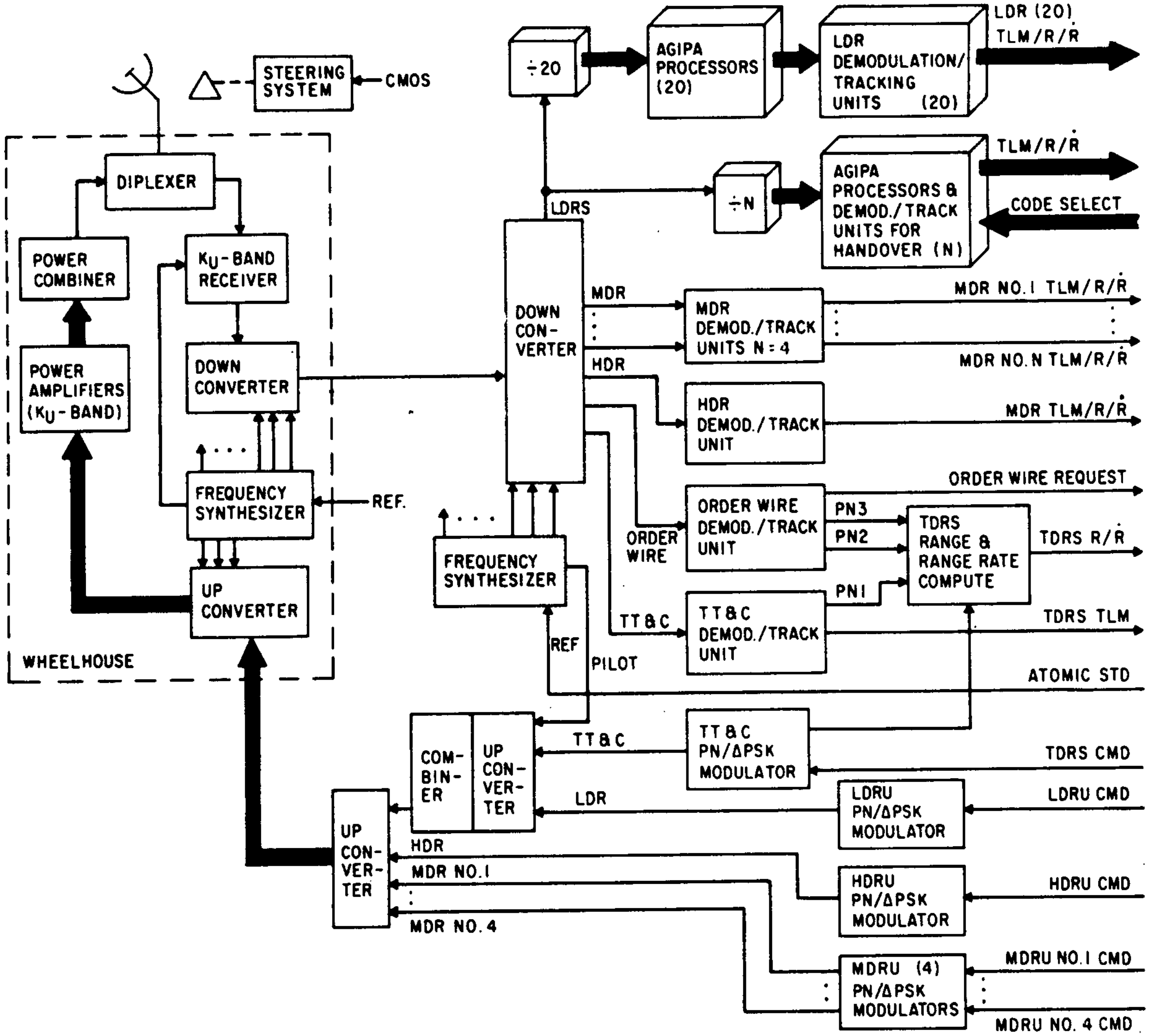

FIGURE 4-15. TDRS GROUND STATION FUNCTIONAL DESIGN: SHUTTLE/ATLAS CENTAUR LAUNCH CONFIGURATION

APPENDIX A

FREQUENCY SELECTION FOR THE HDR LINK

A tradeoff was conducted to evaluate the comparative impact of utilizing X-band (7.7 to 8.5 GHz) versus K_u-band (13.4 to 15.2 GHz) for the high data rate (HDR) space-to-space link. The tradeoff includes consideration for the:

- Link performance
- Bandwidth consideration
- Hardware impact

The link performance can best be evaluated by examining the communication link equation; viz:

$$CNR = \frac{P_t G_t G_R}{K T_s B} \times \left[\frac{\lambda}{4 \pi R}\right]^2 \qquad \text{(A-1)}$$

where:

P_t = transmit power

G_t = transmit antenna gain

G_R = receive antenna gain

K = Boltzman's constant

T_s = system noise temperature

B = receive bandwidth

λ = wavelength

R = range

Equation (A-1) can be rewritten as shown below by letting $G = \frac{4 \pi A}{\lambda^2}$ where A is the effective antenna aperture:

$$\mathrm{CNR} = \frac{P_t \left[\frac{4 \pi A_1}{\lambda^2}\right]\left[\frac{4 \pi A_2}{\lambda^2}\right]}{K\, T_s\, B} \times \left(\frac{\lambda}{4 \pi R}\right)^2 \tag{A-2}$$

$$= \frac{K}{\lambda^2} = K^1\, F^2$$

where K and K^1 include all parameters of the equation which are independent of the frequency. Consequently, if we assume aperture constrained antennas at the user and TDRS spacecraft, equation (A-2) shows that the CNR for this link is essentially proportional to the frequency squared; or in other words, this link should be operated at the highest practical frequency that can be implemented within the user and TDRS spacecraft. The relative link performance improvement for the pertinent X-band and K_u-band frequency has been summarized in Table A-1, and shows +4.6 dB in the forward link and +4.7 dB in the return link, and is also plotted in Figures A-1 and A-2 as a function of user antenna aperture for the forward and return links, respectively.

Table A-1 also shows the impact on the prime power source - due primarily to the RF-to-dc convertion efficiencies of the transmitter. When using a solid state amplifier (6 percent at K_u-band versus 10 percent at X-band) there is a 2.2 dB advantage in operating at X-band. In general, the net link gain might be expressed as the sum of the link performance and prime power improvements; and Table A-1 shows the net relative advantage of approximately +2.5 dB when the HDR space-to-space link is operated at K_u-band.

The prime power impact on the user has been plotted in Figure A-3 as a function of the user antenna aperture for data rates (H) of 10^8, 10^7, and 10^6 bps. The incremental step change in each curve reflects a change from using the more efficient (but heavier and less reliable) TWT amplifier at the higher RF power levels to the less efficient and lighter/ more reliable solid state amplifiers at the lower power levels. The curve essentially shows that K_u-band is approximately +2.5 dB more efficient than X-band.

TABLE A-1. RELATIVE FREQUENCY COMPARISON: X-BAND VERSUS K_u-BAND

A. LINK PERFORMANCE = Kf^2

PARAMETER	Δ IMPROVEMENT: K_u/X-BAND RATIO, dB	
	FORWARD	RETURN
TDRS ANTENNA GAIN	+5.0 dB	+4.8 dB
USER ANTENNA GAIN	+5.0 dB	+4.8 dB
SPACE LOSS	−5.0 dB	−4.8 dB
SYSTEM NOISE TEMPERATURE	−0.4 dB	−0.1 dB
Δ PERFORMANCE GAIN	+4.6 dB	+4.7 dB
B. PRIME POWER/WEIGHT ADVANTAGE		
TRANSMITTER EFFICIENCY		
−SOLID STATE	−2.2 dB	−2.2 dB
−TWTA	0	0
NET GAIN	**+2.4 dB**	**+2.5 dB**

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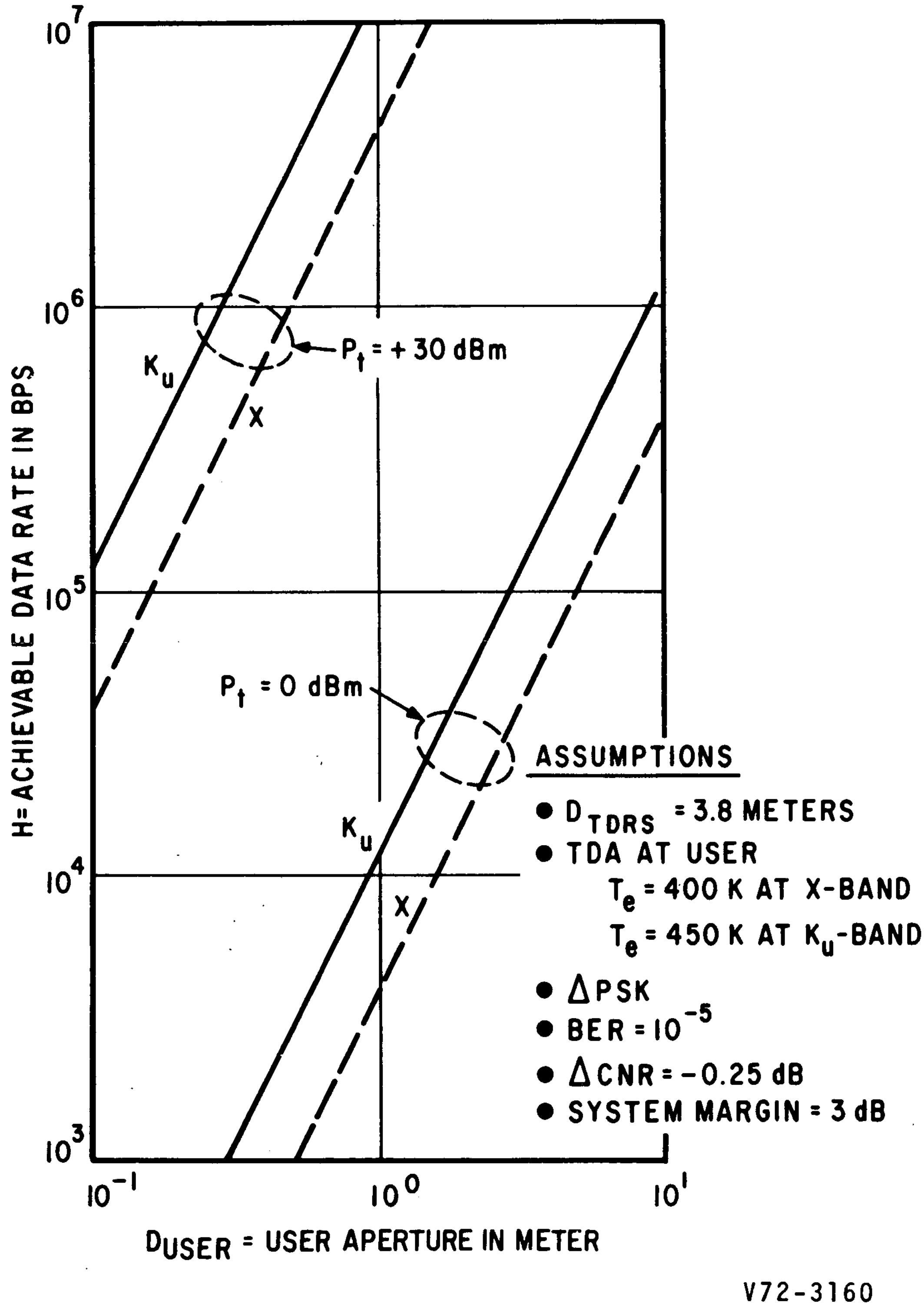

FIGURE A-1. HDR FORWARD LINK: DATA RATE VERSUS D_{USER}

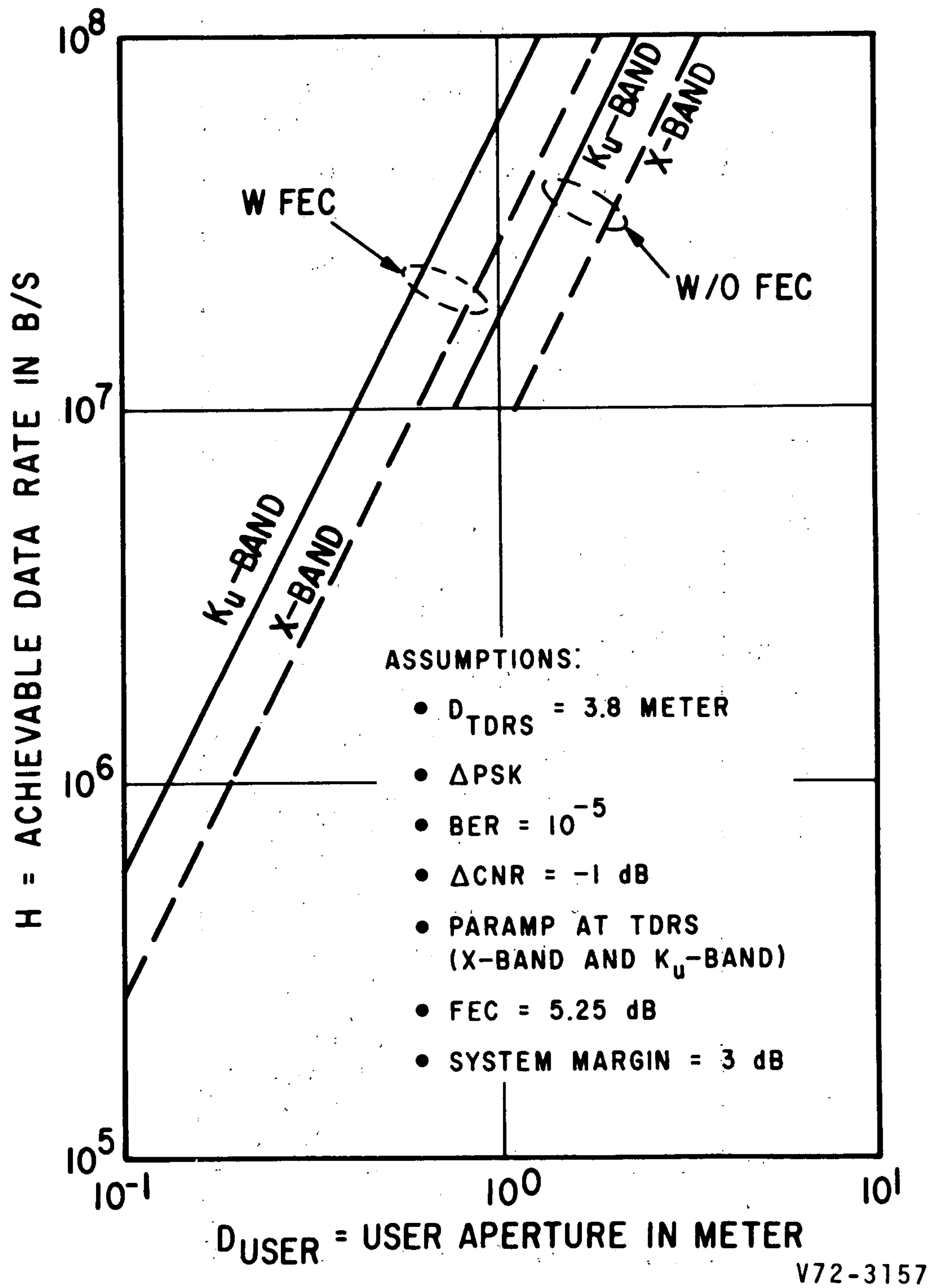

FIGURE A-2. HDR RETURN LINK: DATA RATE VERSUS D_{USER}

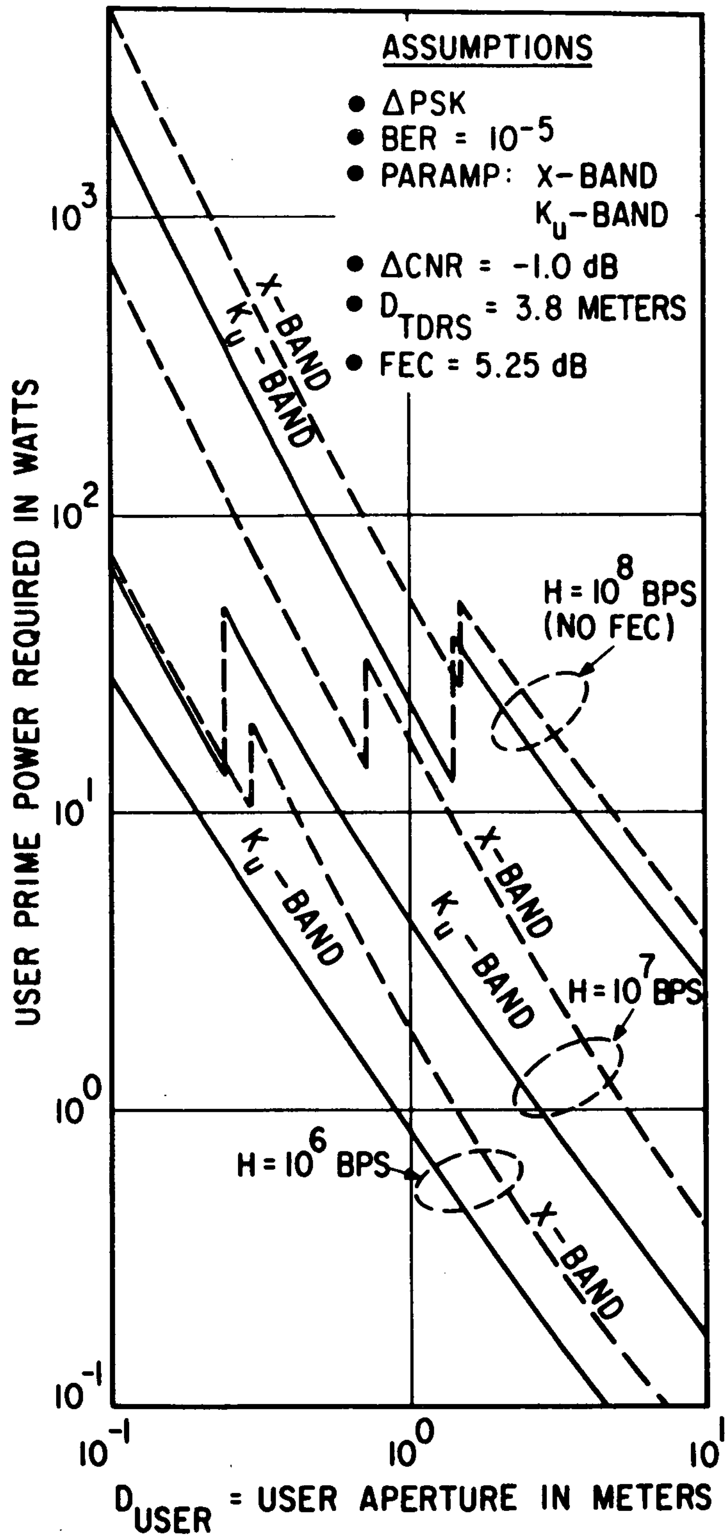

FIGURE A-3. FREQUENCY SELECTION: PRIME POWER IMPACT ON USER (HDR RETURN LINK)

The weight impact of X-band versus K_u-band was determined and is shown in Figure A-4. The transmitter was sized as shown in Figure A-5, which compares the equivalent weights (actual hardware weight + prime power weight at a conversion factor of 0.16 kg/watt) of solid state and TWT amplifiers at X-band and K_u-band. It is seen that it is more weight effective to use solid state amplifiers up to RF power levels of approximately 3 to 5 watts, above which the TWT amplifier becomes more weight effective.

Figure A-4 shows the differential weight penalty (Δ weight) when operating the HDR link at X-band versus K_u-band as a function of data rate (H) for the forward and return links for typical HDR users with antenna apertures of 0.61 and 1.5 meters. In the forward link the Δ weight penalty of operating the link at X-band is primarily at the TDRS spacecraft and approaches approximately 1.2 kg at low data rates for both antenna sizes; but at high data rates (typically 10^7 bps or video) the Δ weight penalty increases rapidly and requires the user to use a larger size antenna. For a 1.5 meter user antenna, the Δ weight penalty is 2 kg, whereas at 0.61 meter the penalty is well above 10 kg.

In the return link, the Δ weight penalty is primarily at the user spacecraft. For return data rates of 10^8 bps, the user must use a larger antenna in order to minimize the Δ weight penalty. Nominally for a 1.5 meter user antenna, the Δ weight is 1.3 kg.

The spectral bandwidth consideration is based on using the same allocated K_u-band for both the space-to-space as well as space-to-ground links. Since adequate bandwidth is available as shown in Figure A-6 to share the K_u-band for both links, the choice of using X-band lies only in future growth systems where the allocated K_u-band may become inadequate when many HDR users must be supported. For the case of one HDR and 4 MDR as described for the Atlas Centaur/Space Shuttle configuration, the allocated K_u-band is more than adequate.

Other frequency selection criteria are summarized in Table A-2, viz. in the hardware impact:

- Space qualified X-band hardware is flying today; however, it can be expected that K_u-band hardware can be similarly available in time for the planned 1977 launch.
- Existing STDN S-band antenna may be convertible to operate at X-band by simply modifying the feed, whereas a new antenna will be required to support K_u-band.

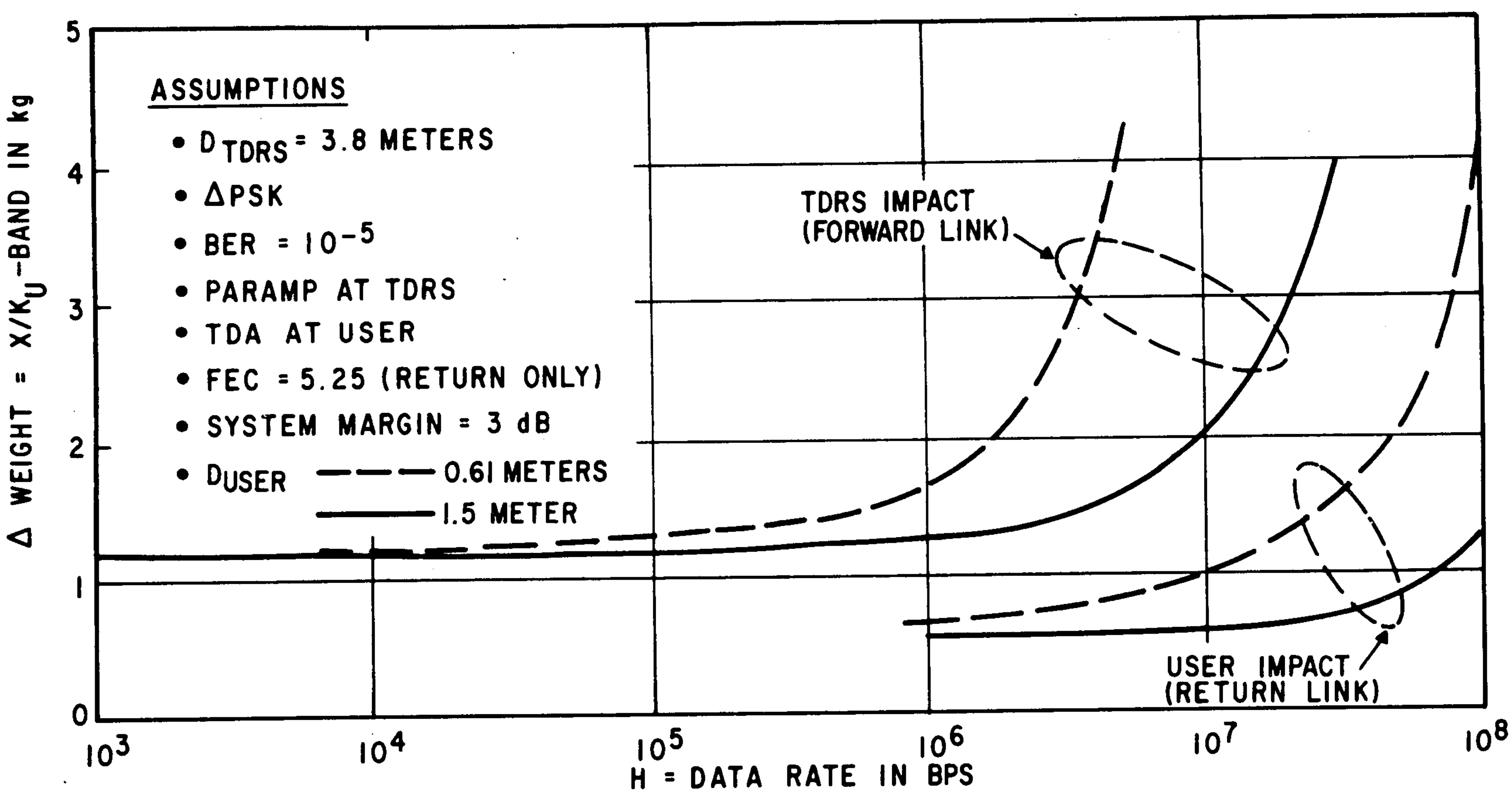

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FIGURE A-4. FREQUENCY SELECTION: RELATIVE WEIGHT IMPACT (X/K_u RATIO)

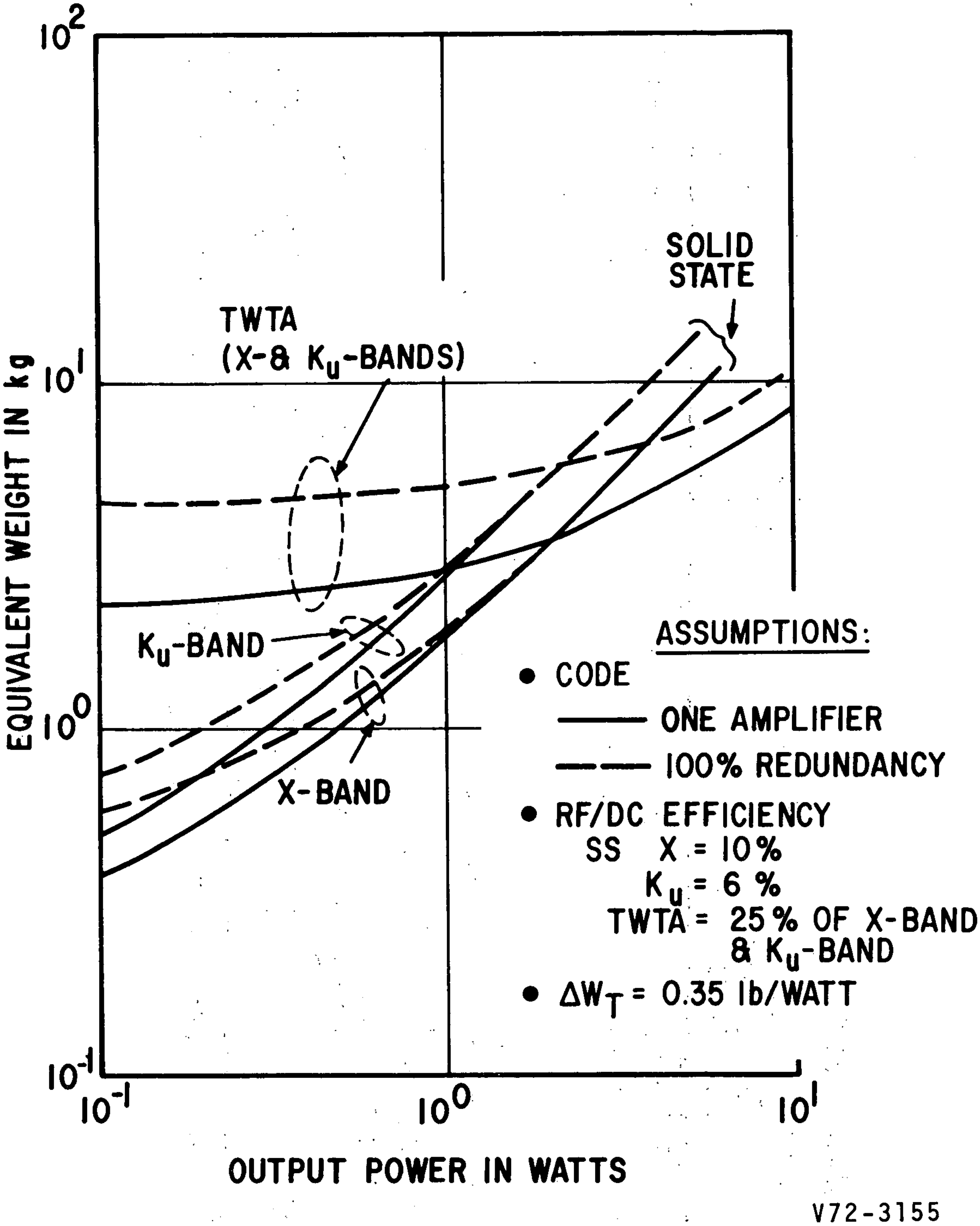

FIGURE A-5. HDR TRANSMITTER SOURCE: SOLID STATE VERSUS TWTA

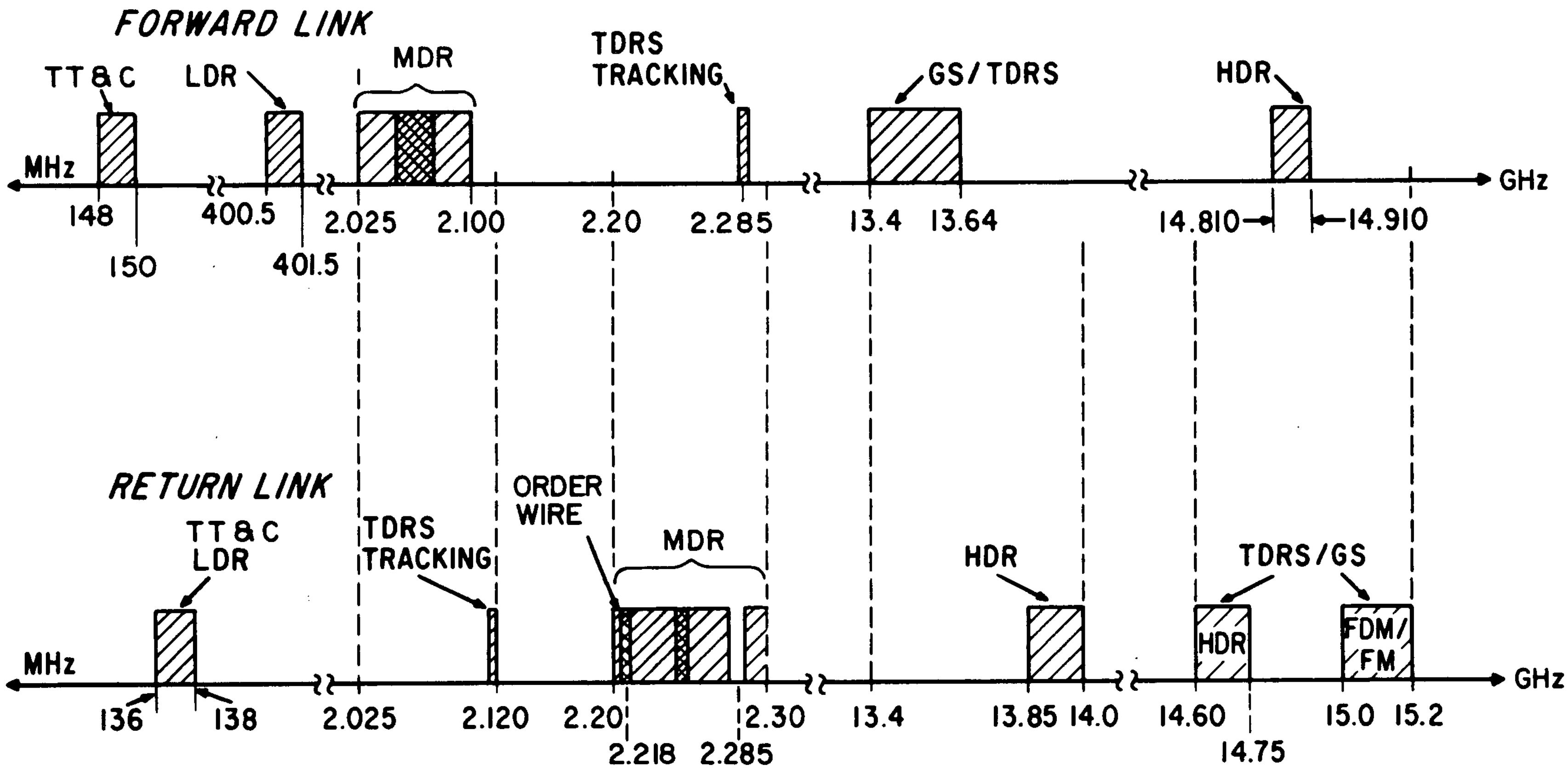

FIGURE A-6. TDRS TELECOMMUNICATION SYSTEM FREQUENCY PLAN

- Operation of the space-to-space link at K_u-band provides an inherent K_u-band functional backup transponder for the space-to-ground link. Separate transponder must be added at X-band to provide similar functional redundancy.

In summary, the improvement in link performance, reduced weight at both TDRS and user spacecraft, and inherent functional backup for the space-to-ground link provides a clear choice to operate the HDR space-to-space link at K_u-band. Therefore it is recommended that K_u-band be selected for this link.

TABLE A-2. SUMMARY OF HDR FREQUENCY SELECTION

PARAMETER	X - BAND	K_U - BAND
LINK PERFORMANCE	LESS	BETTER
HARDWARE IMPACT		
• SPACE QUALIFIED	• FLYING NOW	• CURRENTLY IN DEVELOPMENT; EXPECT TO BE AVAILABLE FOR PLANNED 1977 LAUNCH
• AT TDRS	• LARGER, HEAVIER, AND REQUIRES MORE PRIME POWER	• REDUCED SIZE AND WEIGHT PENALTY; ALSO ≈ 2.4 dB LESS PRIME POWER REQUIREMENT
• AT USER	•	•
• AT GROUND STATION		
— TDRS	• REQUIRES X-BAND TRANSPONDER FOR SPACE-TO-GROUND BACKUP	• HAS BUILT-IN SPACE TO-GROUND BACKUP CAPABILITY
— STDN	• MAY BE ABLE TO MODIFY EXISTING S-BAND ANTENNA FOR X - BAND	• REQUIRES NEW ANTENNA

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APPENDIX B
The Doppler Processor

It is well known that the optimum detector for a CW pulse of known frequency in Gaussian noise is a matched filter. When the frequency is known only within some range, the best strategy is to employ a number of matched filters, one for each resolvable frequency in the range. A practical digital implementation using this technique was used for mechanization of the doppler resolver, shown in Figure B-1.

If the frequency is known, a classical matched filter is optimum for enhancing the signal-to-noise ratio when the signal is accompanied by additive Gaussian noise, prior to envelope detection and decision. By "known" is meant that product of the pulse duration, T, and the frequency uncertainty, W, is less than 1. If the product is on the order of 1, or a little more, then without much loss in the quality of detection, the matched filter may be segmented so that it effectively integrates over n intervals T/n seconds long, such that $WT/n < 1$. Each segment is detected and the n detected values are then added. Thus, some postdetection integration is used in place of predetection integration. As n gets large the process becomes increasingly inefficient. For example, when n = 100, the loss relative to ideal predetection integration (the matched filter) is on the order of 4 to 6 dB.

A more efficient approach is to provide a set of matched filters, one for each frequency across the uncertainty region at intervals of about 1/T Hz. Thus, the number of filters required is on the order of WT.

The first step in the process is to bandpass filter the signal-plus-noise using a filter of bandwidth 2W centered at $W_c/2$ Hz. The principal operation next performed is the computation on the Fourier coefficients of the filtered signal-plus-noise on the interval (O, T). A pulse of unknown frequency can be represented as

$$P(t) = A \cos \left[(\omega_d + \omega_a) + \theta\right] \quad 0 < + < T$$

where ω_c is the nominal frequency, ω_a is unknown, uniformly probable over the range $\pm 2\pi W$ and T is the data bit period. Assume WT to be an integer, M; if necessary by over-estimating W slightly.

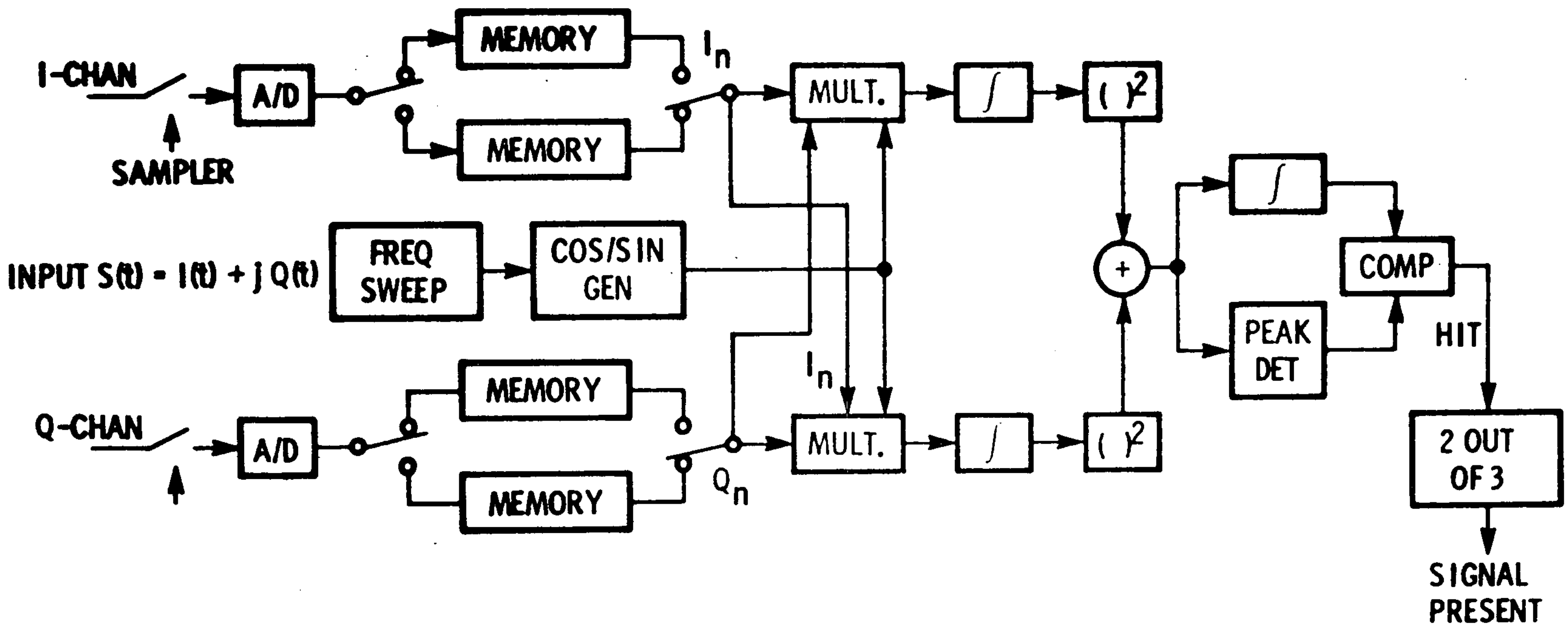

FIGURE B-1. FUNCTIONAL BLOCK DIAGRAM OF THE DOPPLER PROCESSOR

In particular, it is desired to compute the power in each component corresponding to frequencies in the filter passband. These quantities are the values of C_n^2 where

$$C_n = \frac{1}{T} \int_o^T f(t) \exp(-j2\pi n/T)dt$$

for values of n in the region $\frac{\omega_c T}{2} \pm WT$, where f(t) is P(t) + noise.

Alternatively, C_n^2 can be obtained through

$$C_n^2 = a_n^2 + b_n^2$$

where

$$a_n = \frac{1}{T} \int_o^T f(t) \cos 2\pi nt/T \, dt$$

$$b_n = \frac{1}{T} \int_o^T f(t) \sin 2\pi nt/T \, dt$$

In the mechanization it is necessary to store f(t) at the filter output. This is conveniently done by resolving f(t) into its quadrature components, sampling and quantizing so that digital memory can be used. The quadrature components of f(t) with respect to a carrier at ω_c are $f_c(t)$ and $f_s(t)$ such that

$$f(t) = f_c(t) \cos \omega_c t + f_s(t) \sin \omega_c t$$

where

$$f_c(t) = A \cos(\omega_a t + \theta) + n_c(t)$$

$$f_s(t) = -A \sin(\omega_a t + \theta) + n_s(t)$$

in which n_c and n_s are independent Gaussian noise processes of zero mean, the same power, both bandlimited to the frequency interval (-W, +W). On the basis of sampling theory, it would be adequate to sample f_c and f_s at the rate of 2W samples/second, however, as a practical matter sampling should be at 3W to 4W samples/second to allow for non-ideal filtering and to improve the accuracy of the numerical approximations to the integrals. Call the actual sampling rate R, such that RT is a convenient integer. Amplitude quantization of the samples can be performed as crudely as one bit, however, this entails a loss of nearly 2 dB in output signal-to-noise. The use of 3 bit (8 level) quantization reduces this loss to a few tenths of a dB. The sample, quantized values of $f_c(t)$ and $f_s(t)$ will be represented by $F_c(m/R)$ and $F_s(m/R)$ where the range of the integer m is 1 to RT corresponding to the range of t: 0<+<T.

Before writing a final expression for a_n and b_n, it is useful to note certain symmetries in the expressions for values of n spaced equally above and below the midband value, $\omega_c T/2$. To make these evident, let $n = k_c + k$ where $k_c = \omega_c T/2\pi$. The range of k which is of interest is ±WT. Making these changes in notation, approximating $f_c(t)$ and $f_s(t)$ by their sampled, quantized counterparts, and approximating the integrals by sums, we obtain:

$$a_{\pm k} = \frac{1}{2RT} \sum_{m=1}^{RT} F_c\left(\frac{m}{R}\right) \cos \frac{2\pi km}{RT} \pm \frac{1}{2RT} \sum_{m=1}^{RT} F_s\left(\frac{m}{R}\right) \sin \frac{2\pi km}{RT}$$

$$b_{\pm k} = \frac{1}{2RT} \sum_{m=1}^{RT} F_s\left(\frac{m}{R}\right) \cos \frac{2\pi km}{RT} \pm \frac{1}{2RT} \sum_{m=1}^{RT} F_c\left(\frac{m}{R}\right) \sin \frac{2\pi km}{RT}$$

Having computed the 2WT pairs of coefficients, a_k and b_k, the 2WT coefficients C_k^2 are formed. Since only one signal is sought, it is the maximum of all the C_k^2 which need be compared to a threshold to make the detection decision. Since the threshold setting should be proportional to the noise power, it may be convenient to set the threshold as a fixed fraction, β, of the noise power as estimated by the sum of all of the C_k^2. In the usual manner β is chosen to achieve a given false alarm rate, or a given detection probability for a given signal-to-noise ratio, or some similar basis.

Consider the computation of a_k and b_k performed in two modes (data rates of 100 bps and 1000 bps); to resolve the doppler at UHF (≈±16 kHz).

1. R = 32,000 sample/sec, T = 10 msec, and k ranges from -159 to +160. This is equivalent to having 320 filters each 100 Hz wide to cover the frequency uncertainty of ±16 kHz.

2. R = 32,000 samples/sec, T = 1 msec, and k ranges from -15 to +16. This is equivalent to having 32 filters each 1 kHz wide to cover the frequency uncertainty of ±16 kHz.

In the following only mode 1 is described since 2 is operationally identical to 1. Moreover, the example can be extended to cover the doppler frequencies occurring at S-band and K_u-band as well.

The first section of the mechanization concerns obtaining and storing the F_c and F_S data. The F_c and F_S signals are the I and Q channel outputs, respectively, from the baseband signal processing module. These signals are converted to 3-bit words and sampled at the rate of 32,000 samples per second. Thus, for RT = 320, a batch of data characterizing the signal over 10 milliseconds is obtained (i.e., 320 words each of 3 bits). The memory is of shift register type and consists of four 320-word x 3-bit sections, 2 for F_c and 2 for F_S. The two memory registers for each signal component are organized such that while one memory register is being loaded (gathering new data), the other is recirculating at accelerated rate for processing (computing a_k and b_k). The recirculating rate is 10,240 Hz so that 320 pairs of a_k and b_k are computed in 10 msec, which is the required time interval to gather a new batch of data by the other memory register. Thus, by alternating the two memory register's functions, input signals are continually processed until the unknown frequency is found.

The computation of a_k and b_k requires the multiplication of data samples, $F_c\left(\frac{m}{R}\right)$ and $F_S\left(\frac{m}{R}\right)$ by the sine and values, $\cos\left(\frac{2\pi km}{RT}\right)$ and $\sin\left(\frac{2\pi km}{RT}\right)$ and summing the products. $F_c\left(\frac{m}{R}\right)$ and $F_S\left(\frac{m}{R}\right)$ are read out serially from the circulating memory register. The arguments for the sine and cosine are generated by decoding the 4 most significant bits of a 7-bit accumulator which starting at zero accumulates the value of k as m indexes from 1 to 128. (k increments each time m cycles until k ranges from -159 to +160.) At the end of the computation for a particular k, (i.e., at m = 128) the $C_k{}^2 = a_k{}^2 + b_k{}^2 \simeq (\Sigma R_m)^2 + (\Sigma J_m)^2$ is obtained and is presented to the "auctioneer". This is a register and comparator arrangement which is preset to zero at the start of each data batch and thereafter compares the present content of its register with the newly computed $C_k{}^2$. Whichever is greater is then stored in the register. Thus, at the end of a k cycle (i.e., as k goes through the range from -159 to +160), the greatest value of $C_k{}^2$ seen is left in the auctioneer. A final comparison is then made with times the sum of $C_k{}^2$, which is accumulated in a separate register, to make the detection decision.

Having detected the presence of a signal as the doppler processor scans the frequency range in one pass, the doppler processor employs a further decision strategy, whereby two out of three consecutive detections, called hits in the block diagram, are required to be declared a valid hit. This increases the true detection probability and decreases the false alarm rate under the threshold condition of 10 dB S/N in 100 Hz. At the conclusion of a valid hit, an analog voltage corresponding the detected doppler frequency is sent to the carrier and code loop VCO's. This voltage effectively drives the local oscillator frequency to the input carrier for rapid acquisition.

APPENDIX C

FORWARD ERROR CONTROL UNIT

The error correction unit will consist of a rate 1/2 convolutional encoder with a Viterbi (Maximum Likelihood) decoder. The decoder unit can be built on a module which contains its own power supply. The encoder/ decoder designs will accommodate bit rates from 10 kb/s to 1000 kb/s. The units can use large scale integrated (LSI) circuitry.

The performance of Viterbi decoding when used with optimal (maximum distance) codes is shown in Table C-1 along with the non-optimal transparent code chosen for the present application. The values in the table were obtained by computer simulation.

For reasons to be discussed later, a nonoptimal code was decided upon precisely because it is transparent, i.e., complemented code words due to Costas loop demodulation phase ambiguity result in complemented sequences which can be delta decoded to retrieve the original information. Before delta decoding there is only a 0.2 dB degradation between the transparent code and the optimal non transparent one. After delta decoding the degradation is only 0.4 dB in the probability region of interest, i.e., 10^{-3} and 10^{-4}, when compared with the K=5 optimal code used without delta coding and decoding.

The error correction unit which will be discussed is a K=5, non-systematic, transparent, and noncatastrophic convolutional code with the corresponding Viterbi decoder. This decision was based on the following factors:

1. The required value of E_b/N_o can be obtained by the encoder/decoder design.

2. Low power considerations can be met with an LSI design.

3. The transparency of the code offers little degradation in E_b/N_o when used with differential encoding and decoding.

4. The K=5 specification requires the least amount of complexity in hardware.

5. The unit can be implemented with approximately 20 CMOS MSI or 4 custom LSI chips due to the low data rate.

The following sections describe the Viterbi algorithm and offer trade-off performance curves for the constraint lengths of five, six, and seven. Interface considerations are offered in the following sections together

TABLE C-1

VITERBI DECODING OUTPUT ERROR RATE PERFORMANCE

Soft Decision Q=8
(No Δ-decoding except the K=5 transparent code)

Output Error Rate	Uncoded CPSK E_b/N_o	Uncoded DCPSK E_b/N_o	K=7		K=6		K=5 Optimal Code		
			E_b/N_o	Coding Gain over CPSK	E_b/N_o	Coding Gain over CPSK	E_b/N_o	Coding Gain over CPSK	Coding Gain over DCPSK
1×10^{-5}	9.6	9.9	4.4	5.2	4.9	4.7	5.2	4.4	4.7
1×10^{-4}	8.4	8.8	3.7	4.7	4.1	4.3	4.4	4.0	4.4
1×10^{-3}	6.8	7.3	3.0	3.8	3.3	3.5	3.5	3.3	3.8
1×10^{-2}	4.3	5.2	2.1	2.2	2.3	2.0	2.4	1.9	2.8

K=5 Transparent Code (Δ-Decoded)

Output Error Rate	E_b/N_o	Coding Gain over CPSK	Coding Gain over DCPSK
1×10^{-5}	5.6	4.0	4.3
1×10^{-4}	4.8	3.6	4.0
1×10^{-3}	3.9	2.9	3.4
1×10^{-2}	2.8	1.5	2.4

with the hardware approach and some design results for the coder/decoder unit.

C-1 PERFORMANCE OF THE MAXIMUM LIKELIHOOD DECODER

In this section, the performance of the maximum likelihood decoder for K=5, 6 and 7; rate 0.5 convolutional codes will be presented. The codes studied are listed in Table C-2. The contraint length 5 and 6 optimal codes are nontransparent to phase reversals while the constraint length 5 nonoptimal and constraint length 7 optimal codes are transparent.

TABLE C-2

CONVOLUTIONAL CODES STUDIED

Constraint Length	Code Polynomial
5 (transparent)	{10011, 11001}
5 (optimal)	{10011, 11101}
6 (optimal)	{111101, 101011}
7 (optimal)	{1111001, 1011011}

A code is transparent if the coded output sequences of two complementary information input sequences are themselves complements of one another. Transparency of the code can be useful in a system which may have a phase polarity ambiguity but it is in no way a requirement, e.g., a phase reversal detector can be incorporated into the decoder itself in lieu of the transparency.

The codes shown in Table C-2 are the optimal codes for their constraint length except the K=5 transparent code. Optimality is defined here as possessing the largest minimum free distance among all codes of identical constraint length.

From a hardware standpoint, it is desirable to use a code of short constraint length (K). As the constraint length of the code is increased, the hardware growth is exponential.

Another hardware problem is to choose the length of the path memory truncation or decoding delay. The problem here is to determine the minimum number of bits to be retained in the path memory without a significant loss in performance. Since there are 2^{K-1} path memories, storage must be allocated for $N \cdot 2^{K-1}$ bits where N is the number of bits retained in each path.

The channel will be modeled as an Additive White Gaussian Noise Channel for all simulations in this section. This is an adequate model for satellite links. As a reference, the probability of a bit error as a function of the signal-to-noise ratio (E_b/N_o) for ideal CPSK and differentially encoded PSK (DCPSK) is shown in Figure C-1. A typical error rate of interest is 10^{-5} which for CPSK is achieved at E_b/N_o = 9.6dB; and for DCPSK, at 9.9 dB.

C.1.1 SOFT-DECISION DECODER

The performance as a function of code constraint length and decoding delay is considered in this section. For the results presented in this section, the decision statistic for the most delayed bit in the path memory is the most likely path metric. Also, the allocated metric storage is four bits with provisions for clamping and resetting.

The performance curves for the soft-decision maximum likelihood decoders for constraint length 5, 6, and 7 codes are shown in Figures C-2, C-3, and C-4. The soft-decision inputs are uniformly quantized to three bits.

Figure C-2 shows the performance of the constraint length 5 soft-decision maximum likelihood decoder for various decoding delays. The performance of the soft-decision decoder improves with increasing decoding delay. However, for a decoding delay greater than 5 constraint lengths, the return is insignificant. An average error rate of 10^{-5} for the constraint length five optimal code with a decoding delay of five constraint lengths is achieved at E_b/N_o = 5.25 dB. This represents a coding gain of 4.4 dB over ideal two-phase PSK.

In the case of the transparent code there is a 0.2 dB loss over the optimal code shown in Figure C-2 because the distance is 6 as opposed to the optimal of 7; also there is a 0.2 dB loss due to the delta decoding needed at the output of the error correction decoder. At 10^{-5} bit error rate, then, the coding gain is only 4.0 dB over coherent PSK.

Figure C-3 shows the results of the simulation of the soft-decision maximum likelihood decoder for the constraint length six code with variable decoding delay. The chosen decoding delay is five constraint lengths. For a

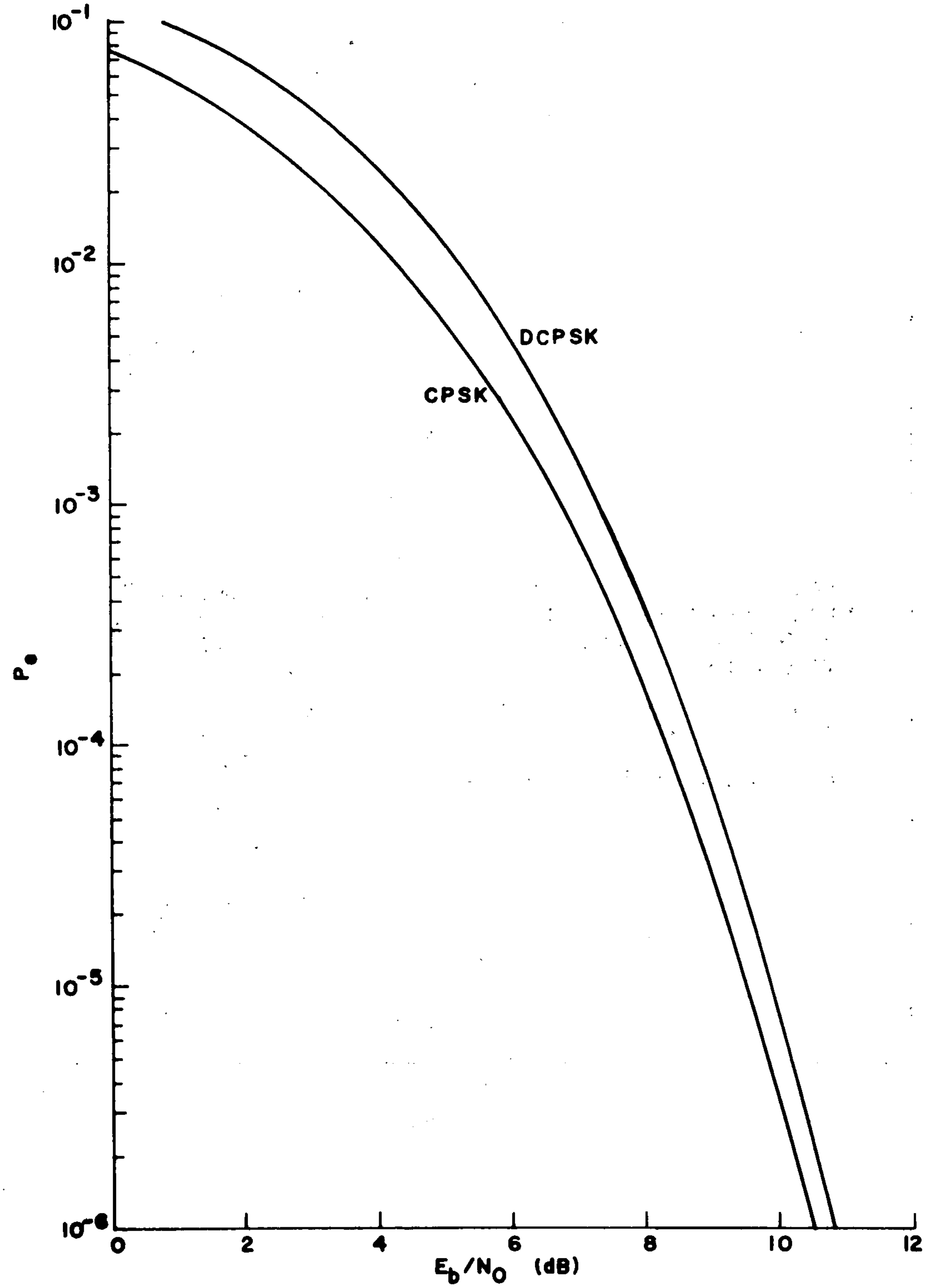

FIGURE C-1. PROBABILITY OF A RAW BIT ERROR VS. E_b/N_o FOR IDEAL CPSK AND DCPSK ON THE ADDITIVE GAUSSIAN NOISE CHANNEL

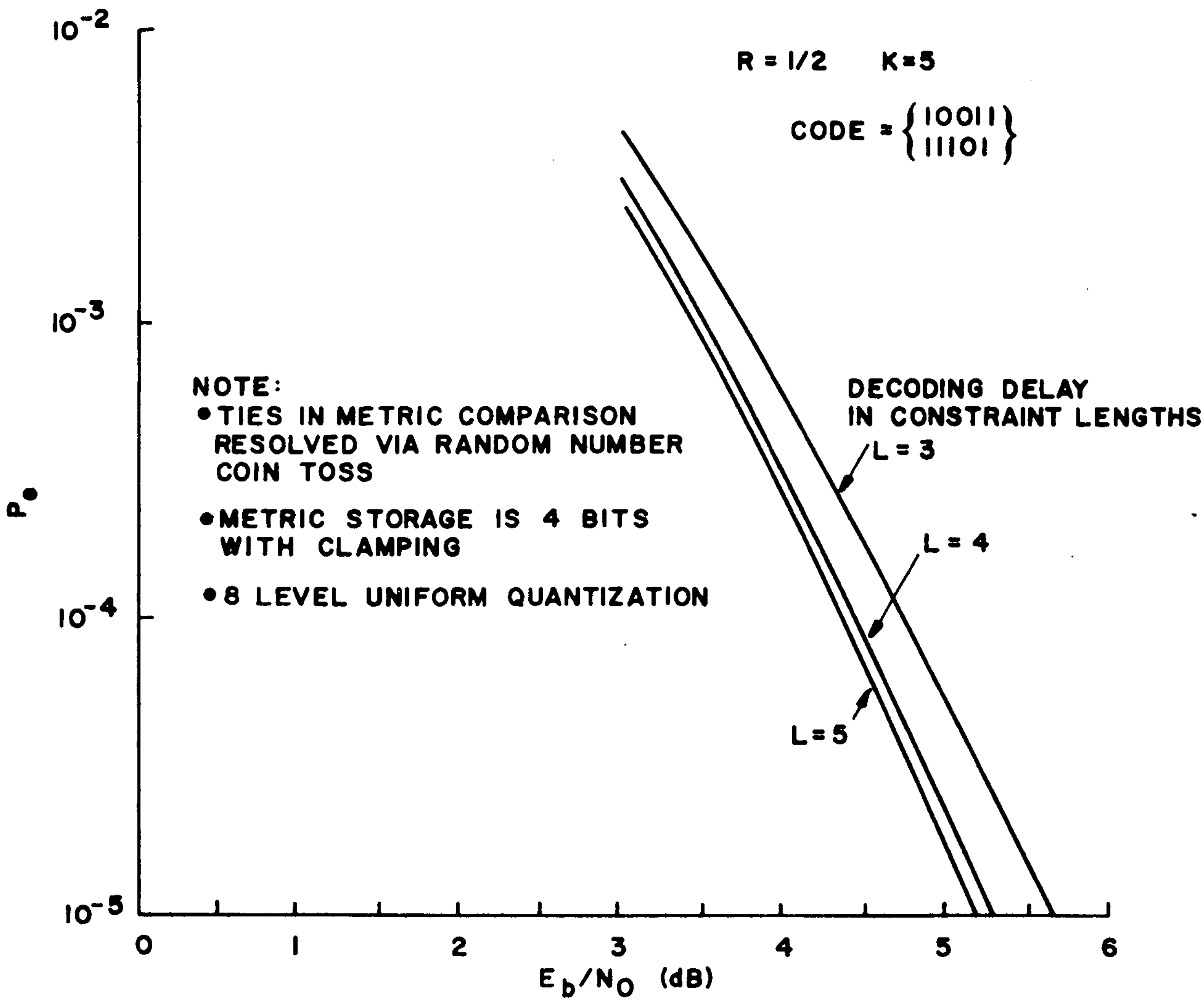

FIGURE C-2. SOFT DECISION MAXIMUM LIKELIHOOD (K=5) DECODER PERFORMANCE USING MOST LIKELY PATH DECISION RULE ON THE ADDITIVE GAUSSIAN CHANNEL

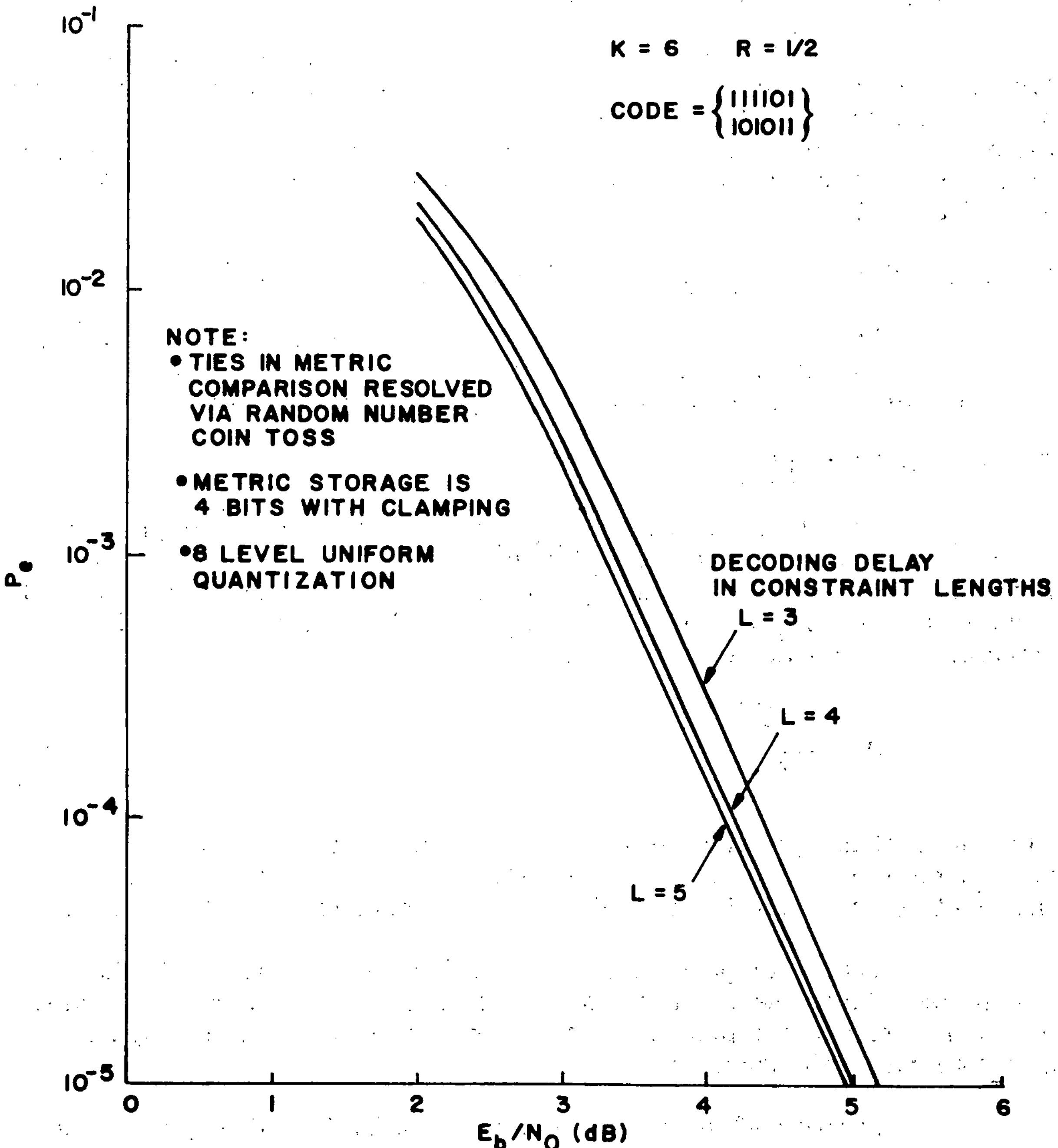

FIGURE C-3. SOFT DECISION MAXIMUM LIKELIHOOD (K=6) DECODER PERFORMANCE USING MOST LIKELY PATH DECISION RULE ON THE ADDITIVE GAUSSIAN NOISE CHANNEL decoding delay of five constraint lengths, the constraint length six soft-decision maximum likelihood decoder achieves an average error rate of 10^{-5} at E_b/N_O = 4.9 dB for a coding gain of 4.75 dB over ideal two-phase PSK.

Figure C-4 shows the result of the simulation of the constraint length seven soft-decision maximum likelihood decoder with variable decoding delay. To attain a coding gain of 5.0 dB or greater over ideal two-phase PSK, the decoding delay necessary is five constraint lengths. For a decoding delay of five constraint lengths, the system performs at E_b/N_O = 4.4 dB for a coding gain of 5.25 dB.

C.1.2 SYSTEM INTERFACE CONSIDERATIONS

The influence of the system interface on the coder/decoder is summarized in the synchronization, inversions due to phase slips in the PSK demodulator, and the quantization of the inputs to the decoder.

C.1.3 QUANTIZATION

One of the most critical parts of the system is the quantization of the soft decisions. The equally spaced quantizer is shown in Figure C-5 for eight levels. The input analog voltage is limited to a maximum signal excursion of $\pm K\sqrt{E_S}$ where $\sqrt{E_S}$ is the mean value of magnitude of the received waveform. The spacing between the levels is given by

$$Q = \frac{2K\sqrt{E_S}}{N} \tag{C-1}$$

The output of the quantizer is a three-bit binary number. The sign bit represents the hard decision on the received channel symbol and the remaining two bits represent the magnitude of the associated confidence level.

The error performance of the decoder is sensitive to the spacing Q selected. Since the level of quantization is fixed to be three bits and to be uniformly spaced, the problem is the selection of the optimum signal excursion as input to the quantizer.

Figure C-6 shows the results of the optimization procedure simulated on a computer for a constraint length 5 optimal code with soft-decision inputs of 3 bits, 4 bits and 5 bits of quantization.

In the case of 3-bit quantization and constraint length of 5, both of which will be assigned parameters of the error correction unit, the optimal level of the quantization setting is $1.8\sqrt{E_S}$.

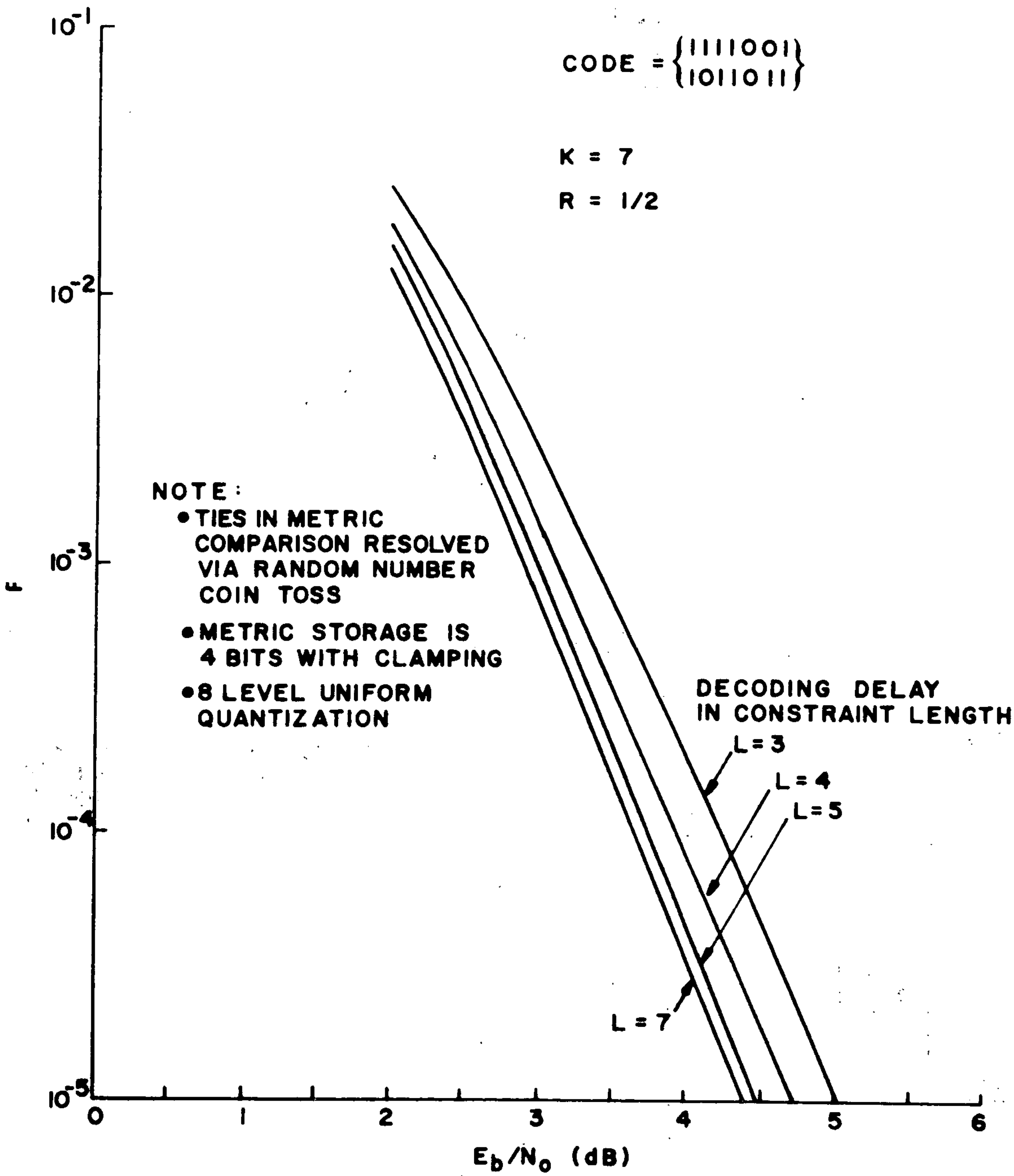

FIGURE C-4. SOFT DECISION MAXIMUM LIKELIHOOD (K=7) DECODER PERFORMANCE USING MOST LIKELY PATH DECISION STATISTIC ON THE ADDITIVE GAUSSIAN NOISE CHANNEL

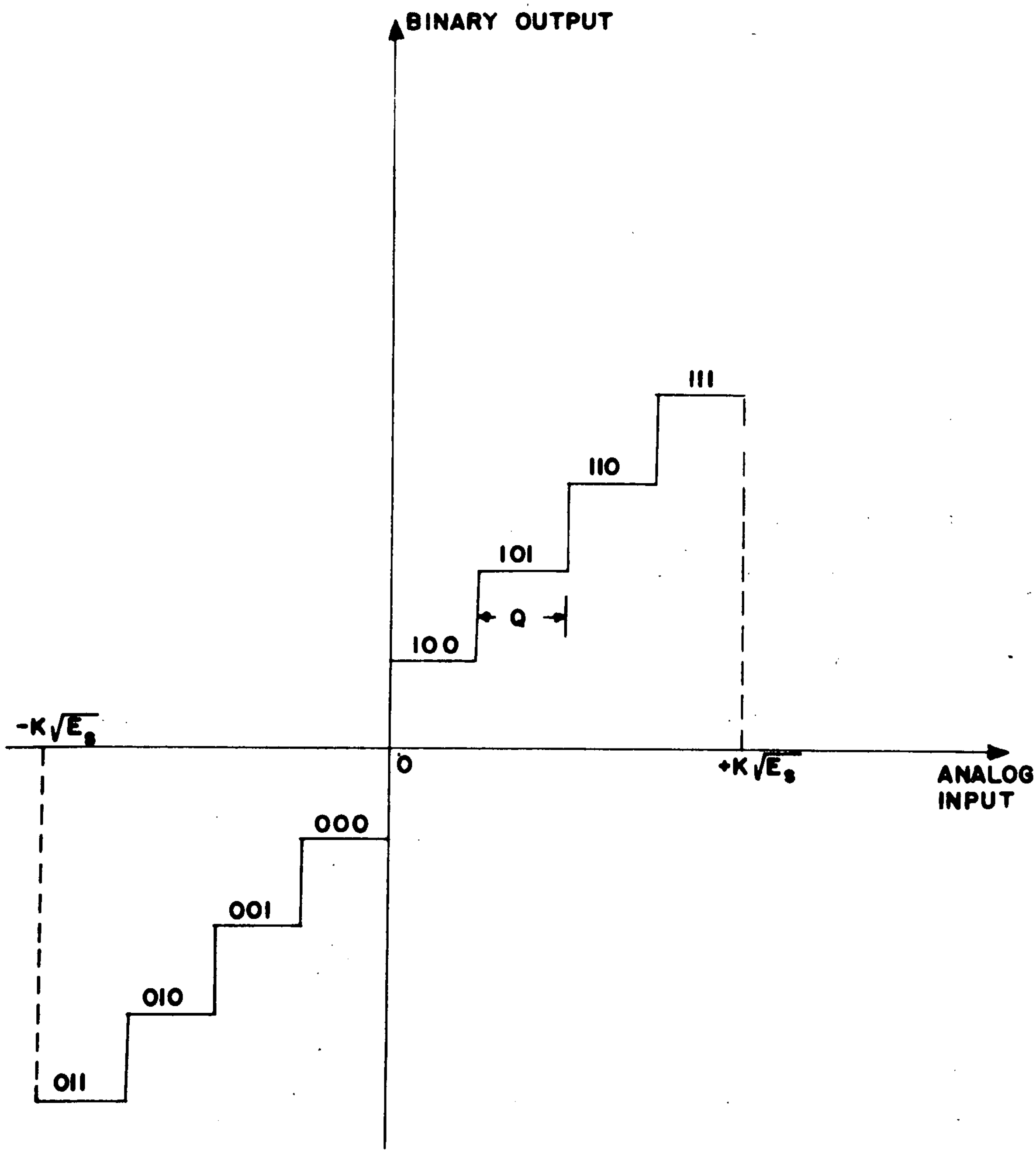

FIGURE C-5. UNIFORM QUANTIZER (N=8 LEVELS)

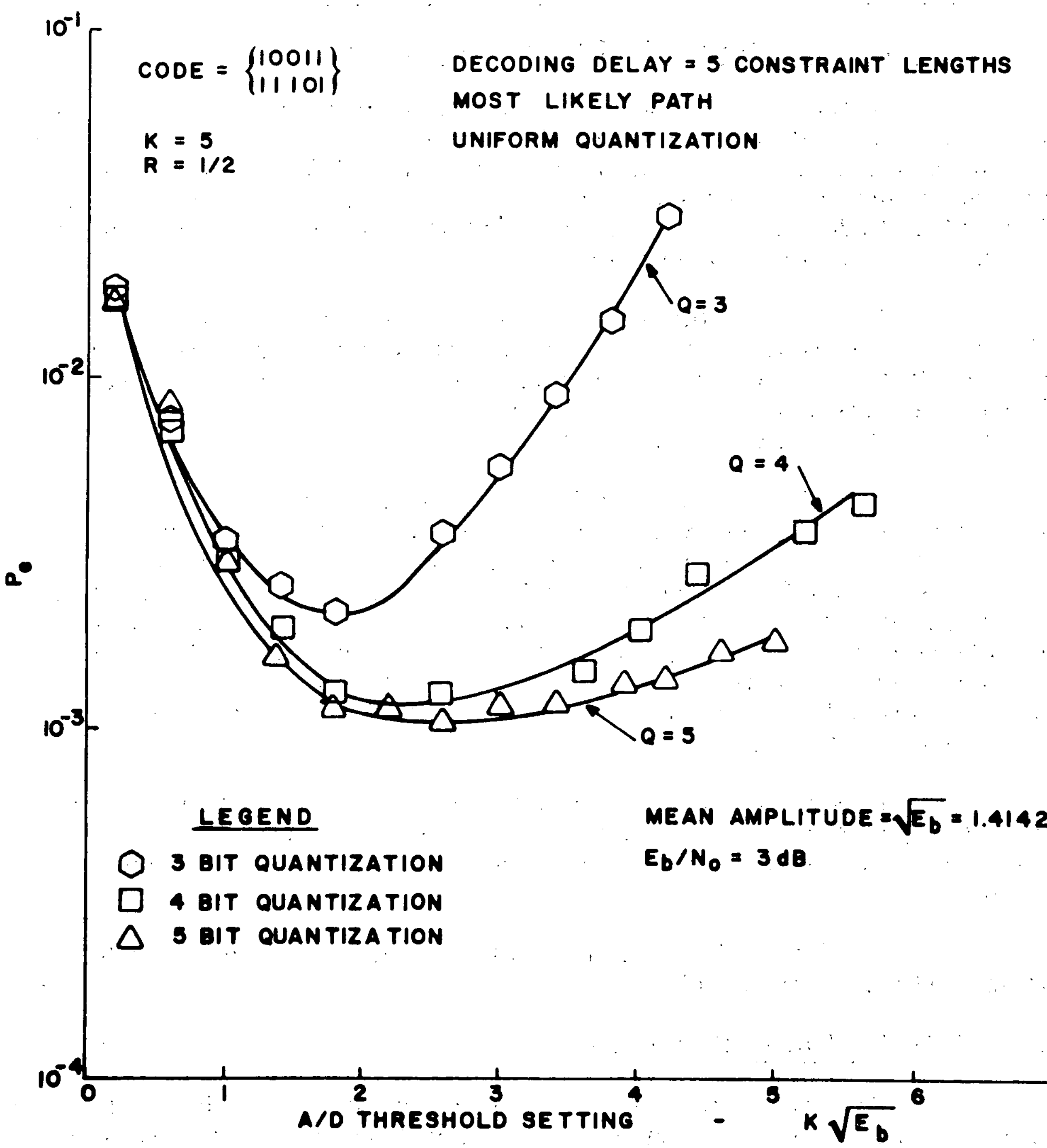

FIGURE C-6. COMPARISON OF THE EFFECTS OF SOFT-DECISION QUANTIZATION ON A/D THRESHOLD SETTING AND ERROR PERFORMANCE OF A MAXIMUM LIKELIHOOD DECODER

C.1.4 NODE SYNCHRONIZATION AND PHASE-FLIPS

C.1.4.1 NODE SYNCHRONIZATION ON CODE SYMBOLS

At the receiver (decoder) two timing problems arise during the transmission of data by a two-phase PSK system. They are the initial node synchronization on the code symbol pair of the rate 1/2 convolutional code and the monitoring of this synchronization. To monitor sync, the receiver is required to detect when a sync error (bit slip) has occurred and initiate action to recover from the sync error.

Code symbol (node) synchronization within a branch is necessary. Clearly, if the wrong decision of code symbol pairs is made, the decoder will constantly make errors thereafter. This situation can be detected because the mismatch of code symbols will cause all path metrics to be large since there will be correct paths.

A method of detecting this condition is to count the number of metric resets that occur over a specific time interval. If the resets occur too frequently, the decoder can assume that it is out of sync and initiate the appropriate action for correction. Figure C-7 shows the number of metric resets per bit as a function of E_b/N_o for the constraint length 5 maximum likelihood decoder operating in the out-of-sync and in-sync modes. Note that the number of metric resets is essentially constant for the out-of-sync mode, whereas for the in-sync mode, the number of resets decreases as E_b/N_o increases.

Another method of detecting sync errors is to accumulate the most likely path metric and compare it to a threshold after a specific time interval. This can be considered as a finer grating of the reset counting method.

The first of these methods was chosen for the initial synchronization and monitoring problems.

C.1.4.2 DETECTION OF PHASE FLIPS

The carrier tracking loop introduces another problem. The tracking loop is subject to phase flips which will cause the decisions to be inverted. That is, a zero will be decoded as a one and a one as a zero.

If the convolutional code is transparent, however, the bit inversions can be compensated by the DCPSK encoding. The maximum likelihood decoder will operate without loss of performance after the phase flip has occurred and the decoding memory associated with the actual time of the phase flip is shifted out.

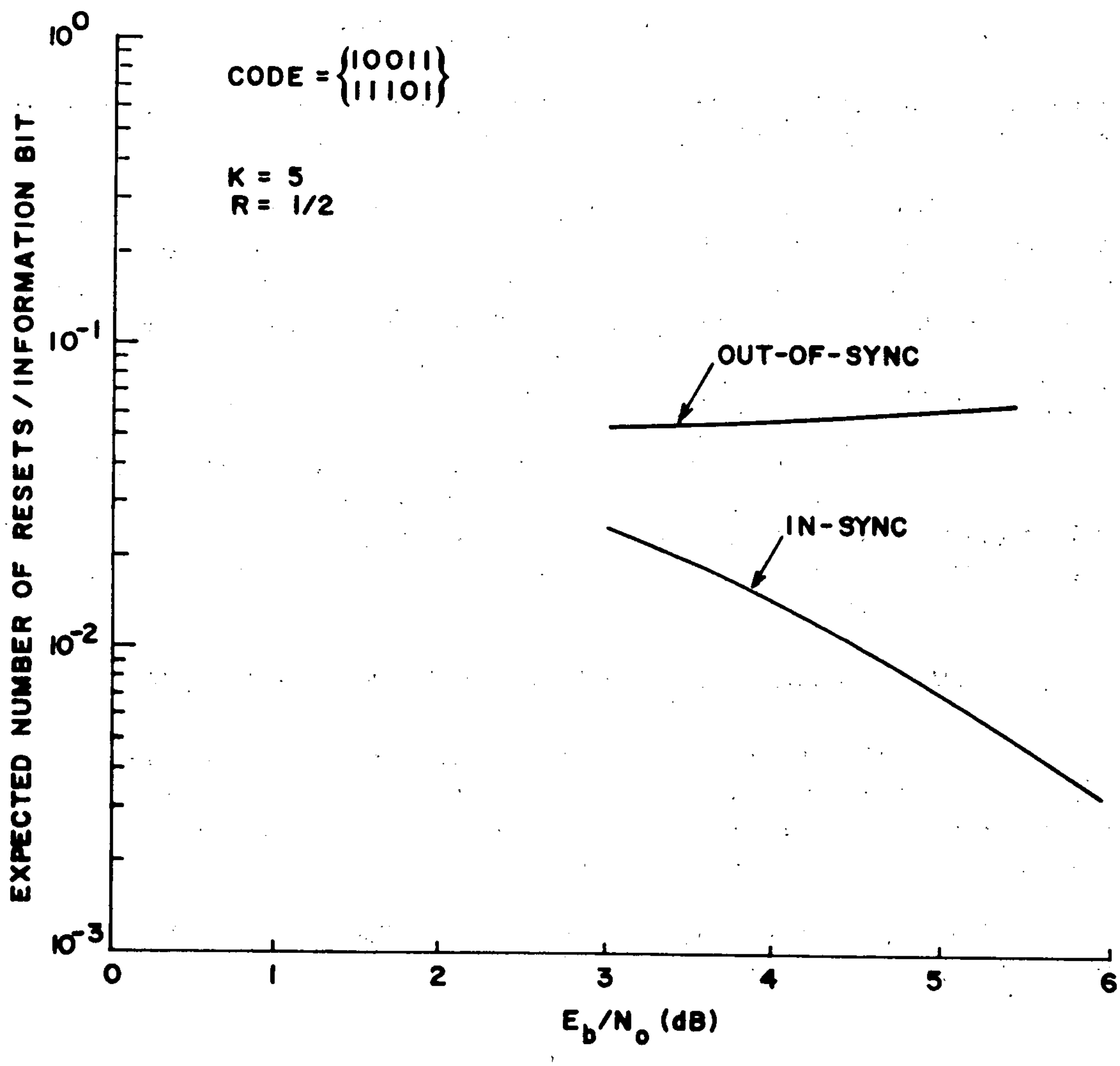

FIGURE C-7. EXPECTED NUMBER OF RESETS PER INFORMATION BIT VS. E_b/N_o FOR MOST LIKELY PATH DETECTION RULE USING 4 BITS OF METRIC STORAGE WITH CLAMPING AND 3-BIT UNIFORM QUANTIZATION

Note that if a nontransparent code was used, the maximum likelihood decoder will act as if it were out of sync. Errors will propagate and the metric will grow rapidly. When a nontransparent code is used, the decoder can be in the following out-of-sync modes:

1. bit inversion and in-sync
2. bit inversion and out-of-sync
3. out-of-sync without a bit inversion

Figure C-8 applies for conditions 1. and 2. above. Therefore, it may take longer for the decoder to obtain synchronization.

It should be noted also that when a nontransparent code is used bit inversions due to phase flips must be accounted for within the decoder. On the other hand, when a transparent code is used bit inversions are accounted for by external sources, i.e., with DCPSK coding and decoding.

C.1.5 ERROR RATE INDICATOR

In order to obtain an indication of the decoded error rate (error rate after decoding), one can just integrate the average number of resets which is monotonically related to the probability of an error. Figure C-9 illustrates this point by plotting the probability of an error as a function of the average of resets.

C.2 DESIGN OF A CODER/DECODER WITH CONSTRAINT LENGTH 5

The following section describes the design and development of a convolutional encoder and decoder for transmission of data at rates up to 1 Mb/s, and therefore will more than accommodate the data rates for both the forward and the return links of the TDRS System.

The parameters chosen are given in Table C-3. The input to the coder will be assumed to be differentially coded.

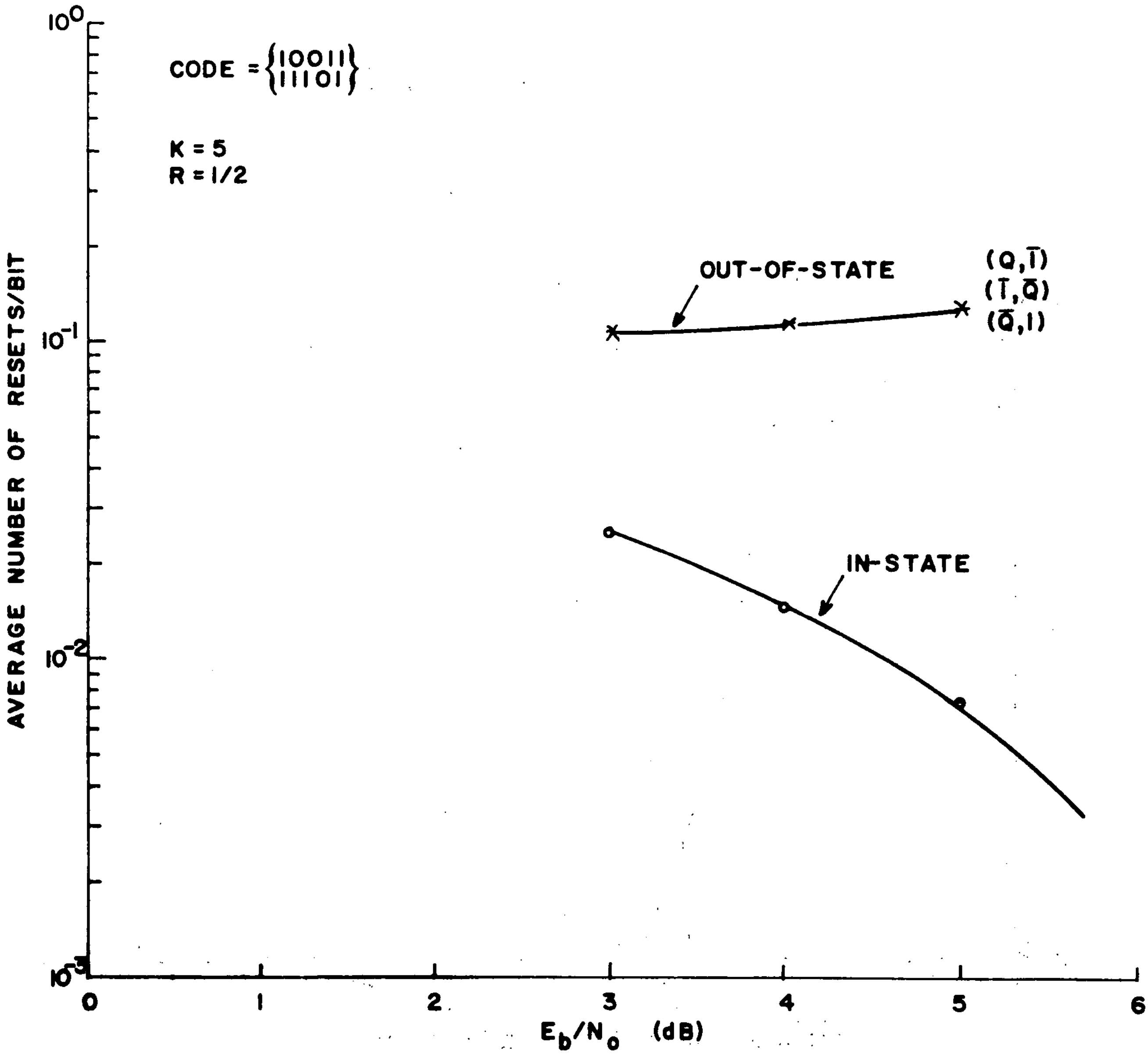

FIGURE C-8. EXPECTED NUMBER OF METRIC RESETS PER INFORMATION DIGIT VS. E_b/N_0 FOR A MOST LIKELY PATH DETECTION RULE USING 4 BITS OF METRIC STORAGE WITH CLAMPING AND 3-BIT UNIFORM QUANTIZATION

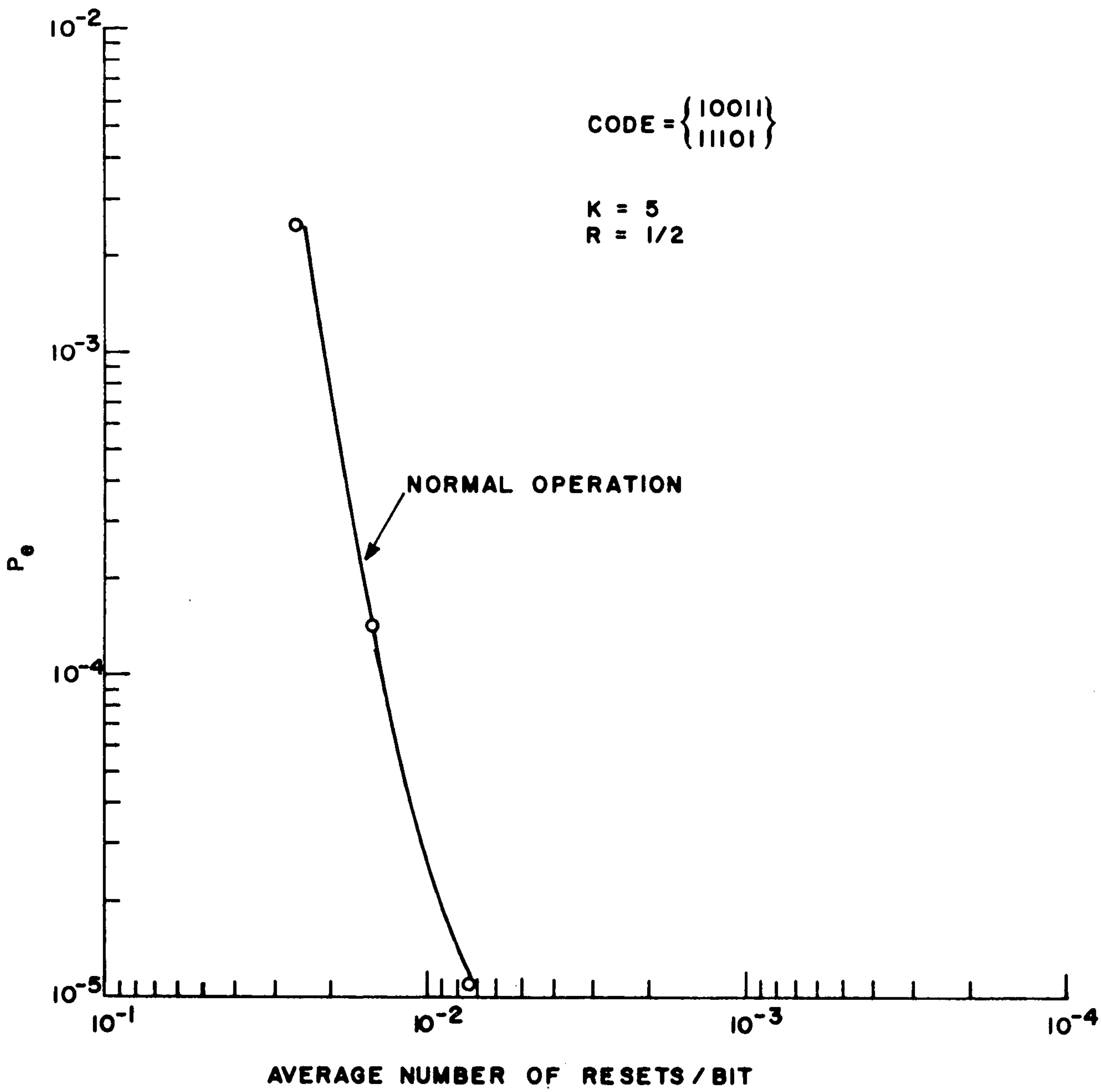

FIGURE C-9. PROBABILITY OF ERROR VS. AVERAGE NUMBER OF RESETS/INFORMATION BIT USING MOST LIKELY PATH DETECTION RULE WITH 4 BITS OF METRIC STORAGE WITH CLAMPING AND 3-BIT UNIFORM QUANTIZATION

TABLE C-3

PARAMETERS FOR CODER/DECODER DESIGN

Code Rate: 1/2

Constraint Length: K=5 bits

Decoder Input Quantization: 3 bits (8 uniformly spaced levels)

Path Delay: 5 constraint lengths

Path Selection: Most likely according to metrics

Metric Storage: 4 bits with clamping

C.2.1 SUMMARY OF THE DECODER DESIGN CONSIDERATIONS AND SYSTEM INTERFACING

The models of the maximum likelihood convolutional decoder and encoder are designed to interface with a biphase system.

Figure C-10 is a block diagram showing the inputs and outputs of the convolutional encoder. The encoder operates synchronously with the channel modulator. Clock signals at rates R and 2R are derived from a local reference. Data at clock rate R are fed into the convolutional encoder from a synchronous source. The output of the convolutional encoder is the serial coded data sequence at a clock rate 2R. Output from the convolutional encoder switch in response to positive-going edges of the clock.

Figure C-11 shows the configuration for the maximum likelihood convolutional decoder. Three-bit soft-decision statistics are brought from the quantizing unit on three parallel lines, one line for each bit of the soft-decisions. The output of the convolutional decoder is the reconstructed delta coded sequence that appeared at the input to the convolutional encoder. It is then delta decoded to recover the original information. The convolutional decoder requires clocks of R and 2R. All output data move in response to the positive edge of clock R.

C.2.2 ENCODER

Figure C-12 shows the rate = 1/2, K=5 convolutional encoder. Data bits are entered serially into a five-bit shift register. For each new input data bit two coded bits are generated by the encoder. Each coded bit is generated by a modulo-2 adder that derives its input from three of the five stages of the shift register. In this manner each coded bit is a function of the new data bit in the register and the four data bits preceding the new data bit in time. The four previous data bits which occupy the last four stages of the shift register are by definition the encoder state.

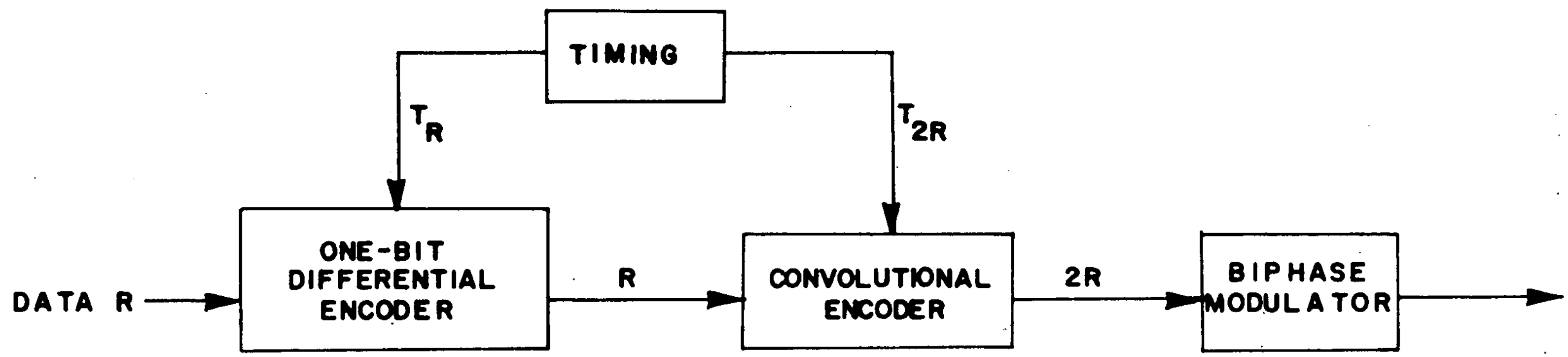

FIGURE C-10. ENCODER CONFIGURATION

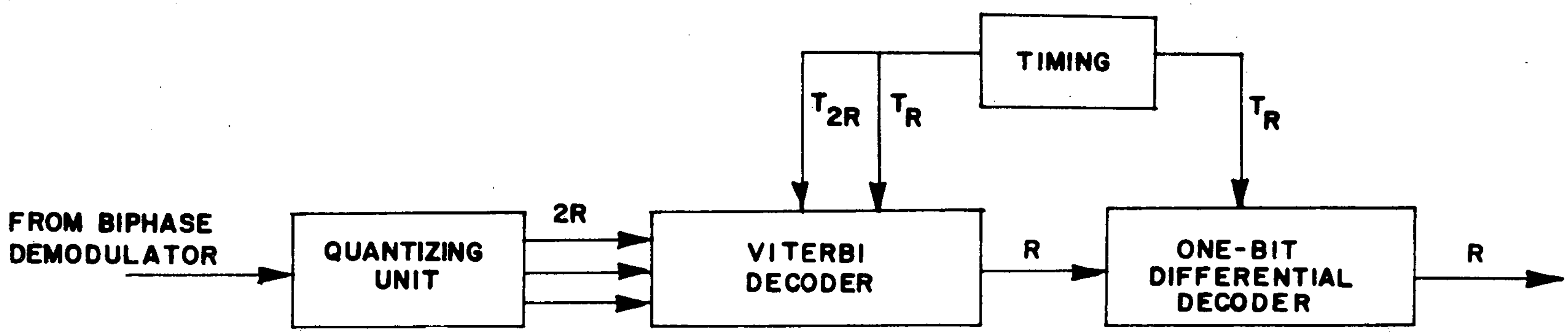

FIGURE C-11. DECODER CONFIGURATION

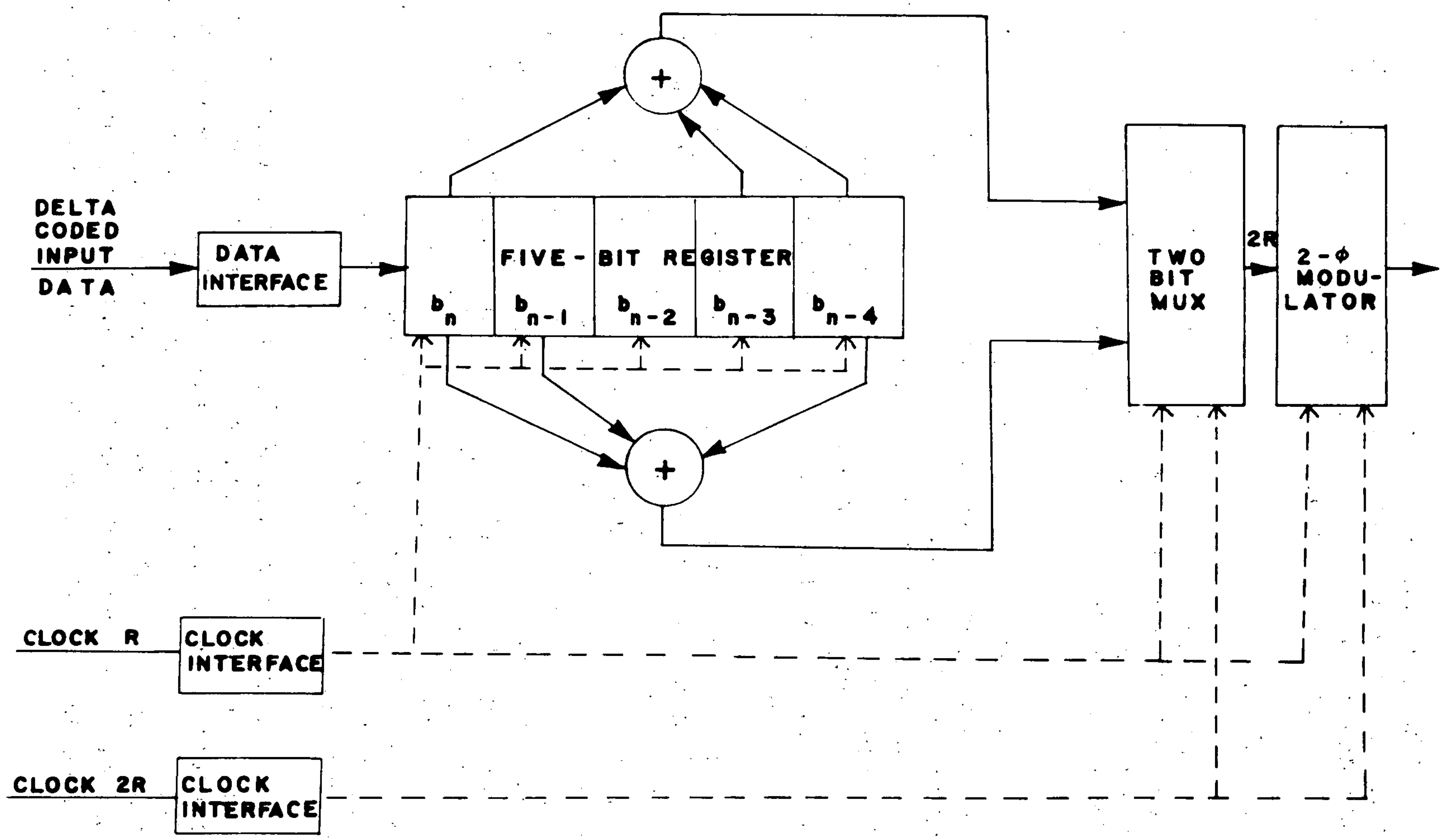

FIGURE C-12. CONVOLUTIONALENCODER R=1/2, K=5

The coder bit pair is time-multiplexed into a single line for transmission at twice the data clock rate. Interface circuits are provided at the input and output of the encoder to translate the logic levels of interfacing equipment to that of the logic used in this system.

C.2.3. DECODER

The function of the decoder is to reconstruct the data stream fed into the encoder from the soft-decision coded bit pair that the decoder receives from the channel modem. The reconstruction is achieved by determining the most probable sequence of states progressed through by the encoder.

Figure C-13 is the block diagram for the decoder. The three-bit soft-decisions from the quantizing unit are received serially by the Metric Transition Generator at twice the data rate (i.e., at the coded data rate 2R). From each pair of soft-decisions received the Metric Transition Generator calculates a probability measure on each of the four possible magnitude values (i.e., 00, 01, 10, 11) of the associated bit pair. The metric transitions are combined with the metrics from the past in order to generate a new set of metrics.

Each metric indicates the reliability of the most probable data path ending in a specific encoder state. Since for a K=5 code there are 16 encoder states, it is necessary to generate 16 metrics. From an examination of the possible sequence of encoder states, it is known that for each new data bit an encoder state can go to only one of two encoder states and conversely any encoder state can be accessed from only two encoder states. From each metric two new metrics are generated, corresponding to the two new states the old state could have progressed to with a new data bit. The two new metrics are calculated from the old metric in the following manner. The first new metric is the old metric added to the metric transition resulting from a zero (0) as the new data bit in the encoder register. The second new metric is the old metric added to the metric transition if a one (1) had been the new data bit.

In this manner a set of 32 new metrics is generated by the adders. The metric comparators perform pairwise comparisons for each pair of metrics leading to the same states. Each comparison selects the most probable of the two metrics. The most probable metric is indicated by the metric of minimum numerical value. The surviving metrics are then placed via the metric multiplex in the position assigned to their states in the metric storage. The surviving metrics now become the old metrics for the next coded bit pair received.

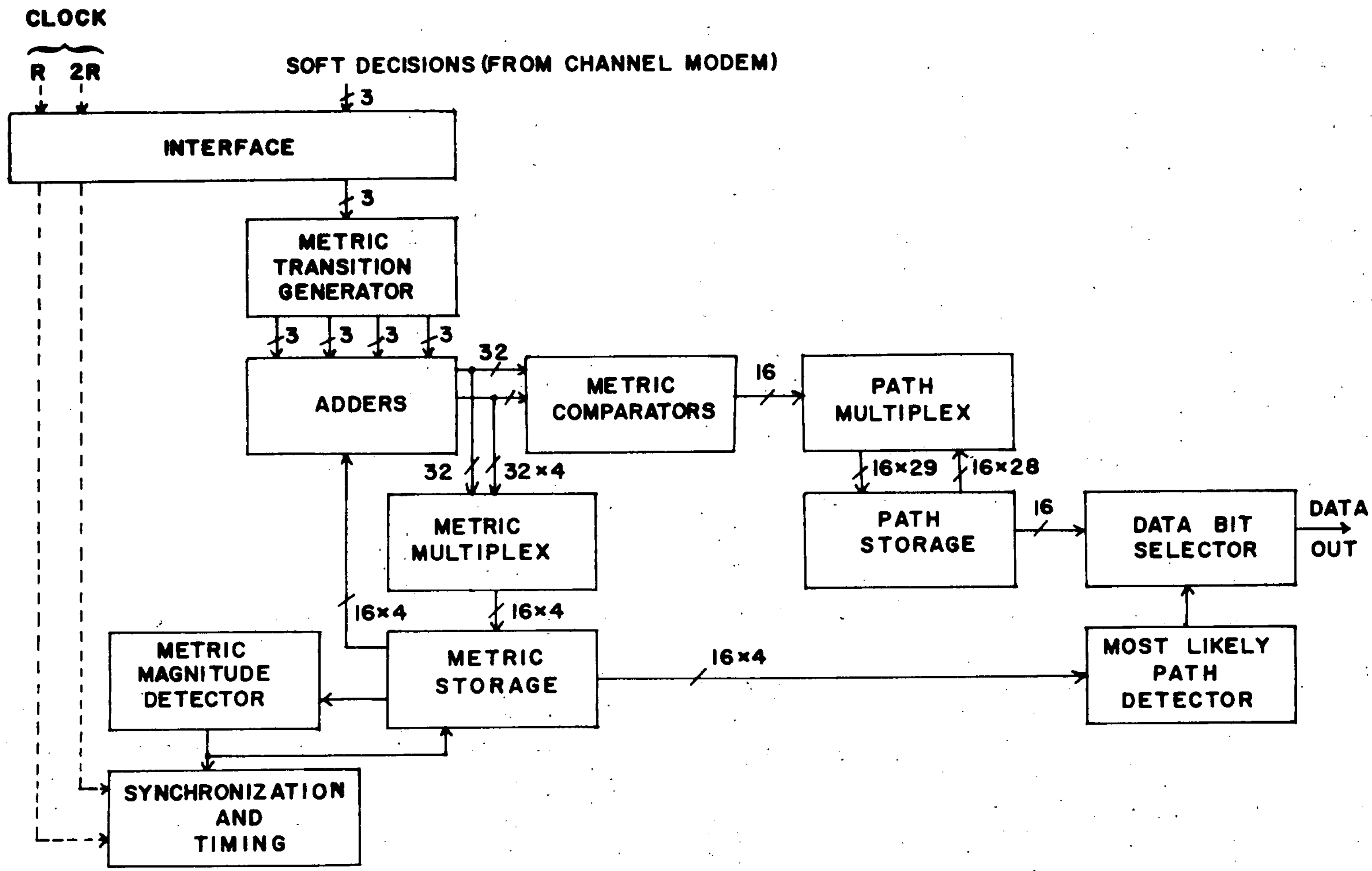

FIGURE C-13. MAXIMUM LIKELIHOOD CONVOLUTIONAL DECODER

In addition to determining the surviving metric, each comparison makes a decision on the most delayed bit in the encoder register. This is due to the fact that the states from which each set of parallel paths is derived can only differ in the most delayed bit position. Therefore, each path bit decision from the metric comparisons is, in effect, delayed four bit intervals from the time it was estimated to enter the encoder register. The bits resulting from each comparison are stored in the path storage.

The path storage corresponds to the sequence of bits or path leading to the state of its corresponding metric. When a comparison and a new bit decision are made, that bit must be added to the path associated with the metric from which the new metric was derived. The new path must then be placed in the storage position associated with the new metric. The result is that the path storage will contain 16 paths, each of which is associated with a metric in the metric storage. The paths are allowed to accumulate for a number of bits equivalent to five constraint lengths of delay from the new data bit positions of the encoder. For any one path five constraint lengths of delay is equivalent to twenty-five minus the four bits represented by the state number itself. Therefore, each path consists of 21 bits.

At each data bit interval the most probable metric in the metric storage is detected. The most delayed bit (i.e., the 21st bit) in the path associated with that metric is chosen as the decoded data bit by the data bit selector.

C.2.3.1 METRIC TRANSITION GENERATION

The metric transition generator calculates a probability measure on the received coded bit pair for each of the four possible received sequences. The four metric transitions ($L_{C_1C_2}$) are defined as follows:

Let r_1, r_2 be the quantized soft decisions on received bits #1 and #2, respectively, of the coded bit pair. Let

$$C_1 = \begin{cases} 0 \text{ if } r_1<0 \\ 1 \text{ if } r_1>0 \end{cases} \qquad \text{(C-2)}$$

$$C_2 = \begin{cases} 0 \text{ if } r_2<0 \\ 1 \text{ if } r_2>0 \end{cases} \qquad \text{(C-3)}$$

and

$$\alpha_1 = |r_1| \qquad \text{(C-4)}$$

$$\alpha_2 = |r_2| \qquad \text{(C-5)}$$

Then the metric transitions are given by

$$L_{00} = \alpha_1 \cdot C_1 + \alpha_2 \cdot C_2 \tag{C-6}$$

$$L_{11} = \alpha_1 \cdot \bar{C}_1 + \alpha_2 \cdot \bar{C}_2 \tag{C-7}$$

$$L_{10} = \alpha_1 \cdot \bar{C}_1 + \alpha_2 \cdot C_2 \tag{C-8}$$

$$L_{01} = \alpha_1 \cdot C_1 + \alpha_2 \cdot \bar{C}_2 \tag{C-9}$$

where a bar over a number indicates its complement and α_1 is the absolute value of the soft decision. These metric transitions are chosen such that a positive soft decision represents a received 1 and a negative soft decision represents a 0.

Figure C-14 is the block diagram for the metric transition generator. The three-bit soft decisions are entered into the r_2 and r_1 registers at the coded data rate. The least significant two bits of r_1 and r_2 (i.e., α_1, α_2) are fed to the inputs of four two-bit adders via NAND gates. The NAND gates will allow or block the adder inputs as dictated by the most significant bit (i.e., sign) of r_1 and r_2, and the metric transition equations. Four three-bit registers store the adder results for each metric transition. These registers are loaded at one-half the coded data rate.

C.2.3.2 METRIC GENERATION

Each data bit interval, a new set of 16 metrics is generated from the new set of 4 metric transitions and the 16 metrics of the past bit interval. There are 16 metric computation cells, one for each encoder state. The cells differ only in the inputs required to calculate the different metrics. Two four-bit metrics of the old states and the three-bit metric transitions associated with the transition to the new states are added together in two four-bit adders.

The output of each adder is a four-bit number representing one of two metrics leading to the same state. The two metrics together with the carries of the adders are compared in a five-bit comparator. The comparator indicates the lesser of the two metrics since that metric indicates the more probable of the two parallel paths. The comparator output also causes a multiplexer to select the indicated metric as the new metric. The output of the multiplexer, which is the new metric, is placed in the four-bit storage register of the new state.

Since both the metrics and the metric transitions are positive numbers, care must be taken to avoid destruction of the metrics by overflows in the addition process. Overflow protection is provided by two mechanisms in the metric calculation. One mechanism detects that both adder outputs being compared have overflowed and causes a maximum number to be

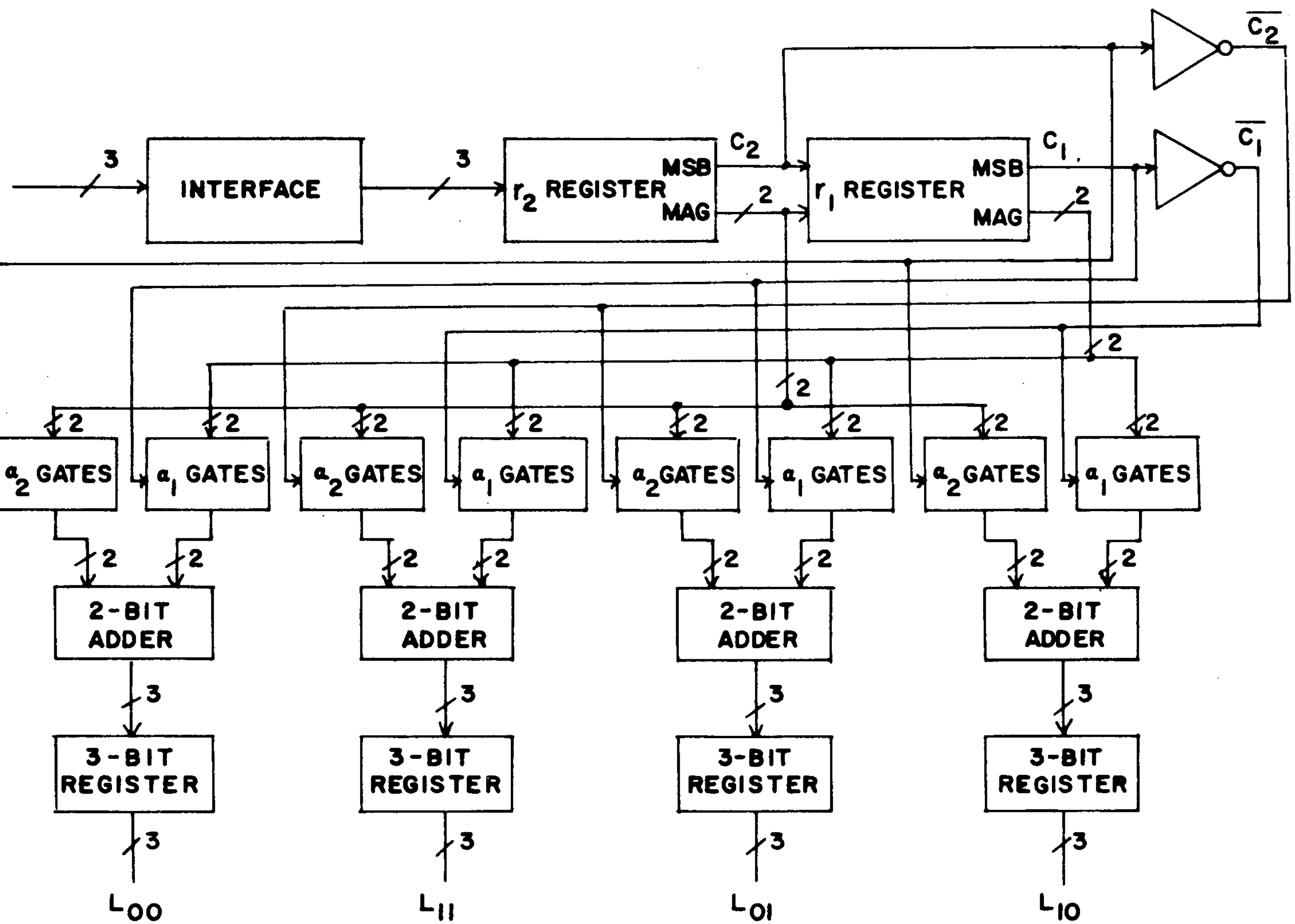

FIGURE C-14. METRIC TRANSITION GENERATOR substituted for the winning metric in the metric register. This is accomplished by detecting that both adders have produced carries and causing the multiplexer to give a maximum output in place of either of its two inputs. The second overflow mechanism detects when all the new metrics are greater in magnitude than one-half the maximum number capable of being stored in a four-bit register. When this condition is satisfied a constant equal to one-half the maximum number for a four-bit register is subtracted from all the metrics. This is accomplished by detecting when the most significant bits of all metrics are set and then causing them all to reset at the same time. By employing these overflow protections it is possible to limit the metric storage for each metric to four bits and still obtain a performance which is essentially the same as that possible for infinite metric storage.

C.2.3.4 PATH GENERATION

For each encoder state, there is a path memory which contains a record of the surviving path leading to that encoder state. Hence there are 16 path memories for the constraint length 5 code. Each comparison between parallel paths leading to the new encoder state, in addition to determining the surviving metric, also determines the surviving path and the newest member of that path. The newest member is a zero or a one depending on the old state from which the transition occurred. If the old state has a zero (one) in the most delayed state position, a zero (one) is appended to the surviving path. The choice of the most probable of the two metric states dictates the selection of this bit.

The path delay or word length is 2 bits long. The select signal from the comparator associated with the path is inserted in the first bit position and is the new member of the path.

C.2.3.5 MOST LIKELY PATH SELECTION

Selection of the most likely path is made on the basis of comparison of magnitudes of the metrics; the path register output corresponding to the lowest metric is selected. The most likely path is obtained by selecting one of eight paths, using three levels of comparison and selection logic; then extending this comparison to one more level, selection of one of 16 paths is accomplished. The path register final stage outputs are connected to the first level inputs. A typical "cell" contains a four-bit parallel comparator, a two-input four-bit multiplexer and an AND-OR-INVERT circuit which serves as a two-input one-bit multiplexer. Each comparator produces a logic ONE at the output opposite the input of smaller magnitude, or at both outputs if the inputs are equal. The smaller metric input to each cell thus appears at an input to the following cell. At the same time, the path register output corresponding to the smaller metric is routed through the AND-OR-INVERT circuit. Note that if the metric inputs to a cell are equal, both path register outputs are OR-gated; a logic ONE at either produces a ONE at the output, an acceptable situation since either choice of path is equally good in such a case.

C.2.3.6 SYNCHRONIZATION

The input to the decoder consists of a succession of soft decision A/D outputs following the integrate-and-dump circuit of the quantization unit. The soft decisions are processed in pairs, each pair corresponding to the pair of coded bits generated by the encoder each time a new data bit is generated.

Computer simulation shows that the metric corresponding to the most likely path increases in magnitude considerably faster when out of synchronization than when properly synchronized. Therefore, resetting of the most significant bit of all metrics (which occurs at times determined the smallest metric, which is by definition that corresponding to the most likely path) occurs more frequently in the out-of-synchronization condition.

C.2.3.7 PHASE SLIPS

As seen in Section C.1.4.2 a phase reversal can cause node synchronization difficulty; the decoder must monitor the metrics to recoginize a two-correct-out-of-four condition. Table C-4 summarizes this. The metric resets provide the clue to any difficulty.

TABLE C-4

SYNCHRONIZATION AND PHASE REFERENCE MODES

SYNC	PHASE SLIP	CORRECT/INCORRECT
in	0 radians	correct
out	0 radians	incorrect
in	π radians	correct
out	π radians	incorrect

APPENDIX D
TDRS GROUND STATION FUNCTIONAL DESIGN

A functional block diagram is presented in Figure D-1 of the TDRS Ground Station. The ground station as presented is essentially the baseline design. The purpose of this appendix is not to offer the figure as a final design, but merely to indicate the functional aspects of the TDRS Ground Station.

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APPENDIX E

S-BAND MULTIPLE ACCESS AGIPA ARRAY

1.0 GENERAL

This appendix describes a 31 element S-band array which provides a multiple access capability to support LDR users with return bit rates up to 20 kbps. The discussion herein is limited to the hardware impact on the TDRS relay spacecraft, and to the impact on the pseudo-noise (PN) modulation format previously selected for the space-to-space links. The hardware design requirements for this S-band array approach were established by link budget and operational analysis performed by the NASA-TDRS Program Office.

Since this S-band approach must support (up to 20 simultaneous users per TDRS Satellite) multiple users in the return link and faces potential interference signals from external RFI emitters as well as interferences from multipath and other in-band users, it has been designed as an Adaptive Ground Implemented Phased Array (AGIPA). AGIPA performs all beam forming, beam steering and signal processing on the ground as described previously in Part I Baseline LDR AGIPA system. Consequently the data form all return link channels must be relayed to the ground.

On transmit, this S-band array approach utilizes a separate array element to provide a fixed wide angle coverage of the entire 30° field-of-view. Two such elements plus their transmitters are used to provide 100% functional redundancy in the forward link.

This S-band Multiple Access AGIPA array replaces the VHF-UHF LDR Transponder used in the Part II Uprated Delta Configuration (Figure 3-1).

2.0 HARDWARE IMPACT ON TDRS RELAY SPACECRAFT

Array Design

The S-band array requires a circularly polarized elements, each of which covers the 30 degree field-of-view. Two element designs have been considered, viz. 1) a helix operating in the axial mode, and 2) a planar array of crossed dipoles. The helix has the relative simplicity in construction as well as low weight; however the planar array can be designed to provide greater aperture efficiency and hence greater element gain. In a recent planar array design for the Skylab radiometer experiment, AIL developed a highly efficient 8 x 8 maxtrix array of linear dipole elements operating at approximately 1600 MHz. This array is fed with an low loss strip line feed network which introduced less than 0.5 dB of insertion loss. The resultant effective aperture efficiency of this array as computed from measured data is better than 82.5 percent. It is expected that the planar array design can provide 1 to 1.5 dB greater element gain than the helix approach; however it is estimated that the planar array design will weigh approximately 4 to 5 times more than the helix design. Final selection of the element design depends on the ultimate need to provide greater gain or to conserve array weight. This array gain could also be exchanged for reduced transmit power to conserve prime power, if required. Furthermore, it may be desirable to use the helix design on receive, and the planar array design on transmit. These tradeoffs should be performed but has not been included in this appendix.

Characteristics of the candidate designs are shown in Table E-1. It is estimated that the 31 element helix array weighs 2.5 kg as compared to 14.2 kg for the planar array design.

Transceiver Design

A block diagram of the 29 channel S-band AGIPA receiver is shown in Figure E-1. The receiver is comprised of the S-band Module Assembly, the FDM Module, and the Local Frequency Reference.

The S-band Module Assembly includes the 29 receive channels. Each channel uses a low noise transistor preamplifier, and doubly converts the S-band signal down to the HF band. The 29 HF signals are then combined and weighted in the FDM Module, and subsequently sent to the TDRS/GS Transceiver for transmission to the ground station.

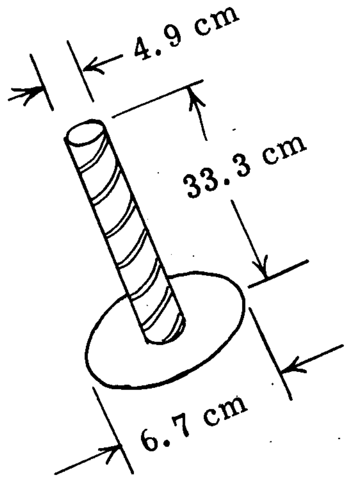
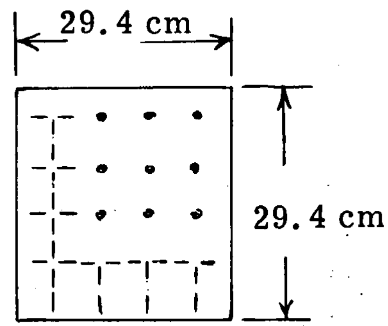
Figure E-2 shows the transmitter design providing an transmit RF power output of 15.0 watts. Two identical channels are used; however only one active channel is used at a time - the second provides 100 percent functional redundancy. Each channel feeds a separate antenna element; therefore 100 percent redundancy exists in the entire transmit channel.

The Local Frequency Reference generates all of the coherent local oscillator signals for the 29 receive channels and 1 active transmit channel. The Local Frequency Reference obtains its coherent reference signal from the main Frequency Source on-board the TDRS Spacecraft.

All active components of the FDM Module and Local Frequency Reference are designed with 100 percent block redundancy; however no redundancy is included in the S-band Module Assembly, but rather graceful

TABLE E-1

SUMMARY COMPARISON OF CANDIDATE ARRAY DESIGNS

Parameter	Helix Array	Planar Array of Crossed Dipoles
1. Transmit Element		
• HPBW°	27.3 (1)	26.0
• Gain (peak), dBi	15.7	17.0
(26°FOV), dBi	13.0	14.0
2. Receive Array		
• HPBW°	5.1	5.1
• Gain (peak), dBi	30.3	31.6
(26°FOV), dBi	27.6	28.6
• Weight, kg	2.5	14.2

NOTE: 1. HPBW was increased to 27.3° in order to limit helix length to approximately 2.5
2. Element shape can be varied to optimally fit within available space.

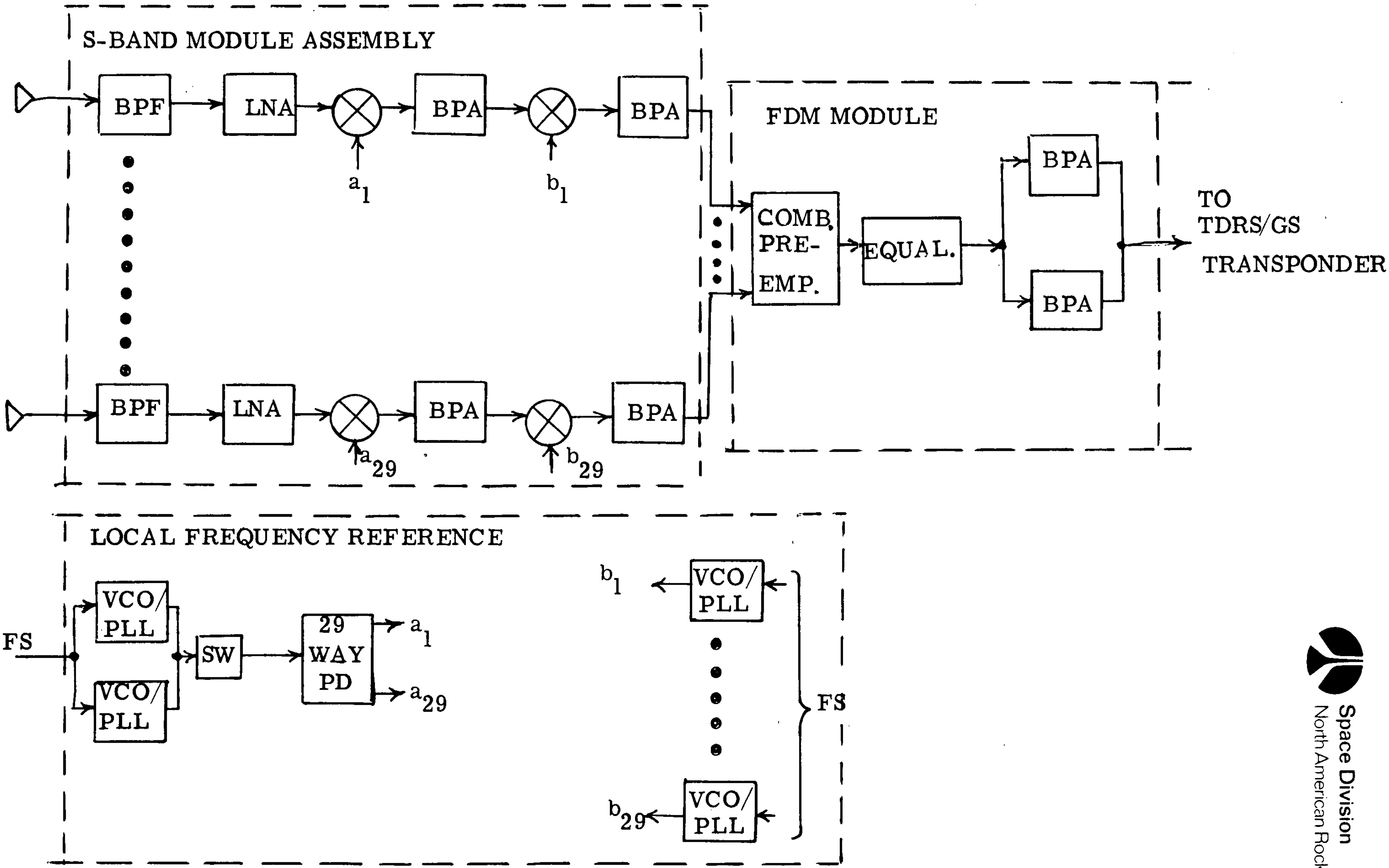


Figure E-1 S-Band Multiple -Access AGIPA Receiver

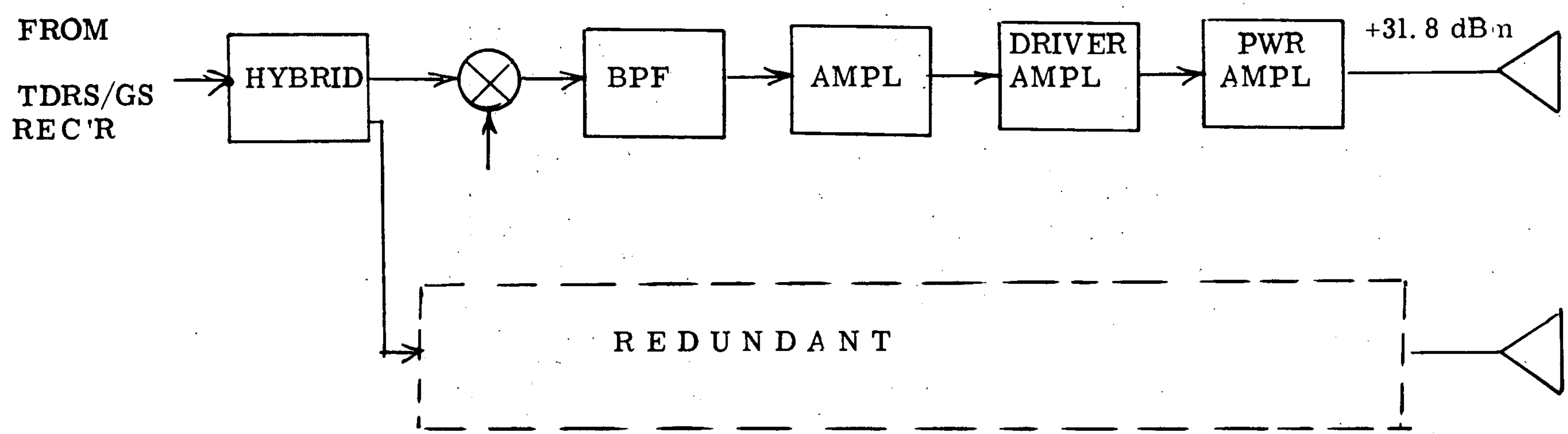


Figure E-2 S-Band Fixed Field-of-View Transmitter degradation of the 29 channels has been assumed permissable.

Weight and Power Summary

Table E-2 summarizes the weight and prime power requirements for the S-band Multiple Access AGIPA Array on the TDRS Relay Spacecraft. For this summary the 31 element helix array is shown.

3.0 SIMULTANEOUS MULTIPLE ACCESS BY THE USER SPACECRAFT

To achieve a multiple access capability in the return link several candidate techniques are realizable. Most multiple access techniques fall into three generic categories: namely, time-, frequency-, and code-division multiple access; in addition, any combination of the aforementioned is possible. Each of these techniques in themselves would be suitable for establishing the link between the user spacecraft and TDRSS; however, selection of the optimum system in terms of communication efficiency, interference rejection, and multiple access capability is subject to further analysis. The optimum system may very well be a combination of several of the aforementioned techniques. It is not the intent of this section to evaluate the various multiple access techniques but merely to present a typical example of accessing.

Consider for example, a code division multiple access (CDMA) system. The use of CDMA for the return link S-band phased array concept is in many ways similar to that proposed for the LDR user operating at VHF. Some of the inherent features of a CDMA system are a reduction of co-channel interference, establishment of a means of orthogonality for each of the S-band array beams, and a built-in ranging mechanism via the PN code rate. In addition to these features the PN code modulation results in a spectrum spreading; thereby reducing the signal power density in a given bandwidth and enabling one to meet the IRAC requirement for minimizing the power flux density at the earth's surface.

TABLE E-2

S-BAND MULTIPLE ACCESS AGIPA WEIGHT & POWER SUMMARY

Component	Weight - Kg		Prime Power - Watts
	Per Unit	Total	
Array			
• 31 elements	78.5 gms	2.44	
Receiver			
• S-band Module Assy.	0.58	16.80	41.5
• FDM Module		0.64	1.0
Local Freq. Reference(1)		7.40	31.0
Transmitter			
• 2 channels (1)	3.200	6.40	47.5
TOTAL		33.68	121.0

NOTE:

(1) Includes 100% redundancies.

As an example of postulated data rates achievable on the return link through the S-band phased array, the following table is presented:

No. of Other Users Illuminated By Same Beam	Achievable Bit Rate (KBps)	
	Chip rate = 1000 $\frac{\text{kilochips}}{\text{sec}}$	Chip rate = 2000 $\frac{\text{kilochips}}{\text{sec}}$
1	34.2	41.0
2	25.6	34.2
3	20.5	29.3
4	17.1	25.6
5	14.6	22.8
6	12.8	20.0
7	11.4	18.6
8	10.2	17.1
9	7.3	15.8
10	6.7	14.6

The table clearly illustrates that as the number of users illuminated by one beam increases, the performance of that link degrades. This due to an increase in the number of interferring users in the beam. One way to overcome this degradation is by increasing the code rate as shown in the third column of the table. The assumptions for the estimates above are as follows:

Carrier Freq = 2000 MHz
User EIRP = 48 dBm
TDRS Ant. Gain = 31 dBi
Noise Density = -170 dBm/Hz

This preliminary estimate indicates that the CDMA approach can be designed to provide the necessary performance for the return link. As mentioned, however, a detailed analysis and investigation of the S-band multiple access approach as applied to LDR users is required so that the optimum multiple access technique can be defined in detail.