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30/20 GIGAHERTZ TRANSPONDER STUDY

FINAL REPORT

DECEMBER 1980

NAS 3-21508

Prepared By

SPACE AND COMMUNICATIONS GROUP
HUGHES AIRCRAFT COMPANY
EL SEGUNDO, CALIFORNIA

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LEWIS RESEARCH CENTER, NASA
CLEVELAND, OHIO

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Hughes Ref No. E5894 • SCG 00504R

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Acknowledgement

The 30/20 GHz Transponder Study encompasses many areas of technology, and is the product of data contributed by many communication specialists.

Mr. Ralph Cager formulated the early transponder conceptual designs and Mr. Mike Fashano quantitatively evaluated each design.

Mr. Sidney Bergman, Dr. Russell Gaspari, and Mr. John Heney supplied data for the technology sections dealing with the input circuits, switching circuits, and TWT output circuits.

Mr. Ray Miyakawa performed the analysis of the transponder thermal control when mounted in typical spacecraft.

Finally, special recognition is given to Mr. Sidney Bergman for his extensive contribution to the final design concepts and for writing this final report.

CONTENTS

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1. INTRODUCTION AND SUMMARY

This study report describes the design and performance parameters of three types of wideband multiple-channel satellite transponders for use in a 30/20 GHz communications satellite, which provides high data rate trunking services to ten ground station terminals. The three types of transponder are frequency-division multiplex (FDM), time-division multiplex (TDM), and a hybrid transponder using a combination of FDM and TDM techniques.

Figure 1-1 illustrates the wideband multiple-beam trunking concept. Table 1-1 gives the traffic distribution between the trunking terminals as required by the study statement of work. The detailed system design constraints are given in Section 2 of this report.

Section 3 of the report describes the design features and performance common to all three transponders, such as the receiver front end design, the frequency conversion scheme, and the local oscillator design. In addition, Section 3.7 describes the thermal interface between the transponders and the satellite.

Sections 4, 5, and 6 contain the design trades, detailed baseline design descriptions and performance parameters of the FDM, TDM, and FDM/TDM transponders. Section 7 of the report provides a comparison of the three designs, including performance, weight, power, cost and initial technology.

Figures 1-2, 1-3, and 1-4 are simplified block diagrams of the baseline FDM, TDM, and FDM/TDM transponder designs. Table 1-2 summarizes the principal design features of each transponder.

Table 1-3 is a comparison of the two most important performance parameters for the three transponders. Bit error rate (BER) degradation performance is roughly the same for the TDM and FDM/TDM transponders. FDM transponder degradation is somewhat greater because of the intermodulation distortion occurring in the power amplifiers due to the use of one amplifier for each three FDM channels. The second column of Table 1-3 is a normalized figure of merit, derived by dividing the power delivered to the feed by the data rate and comparing to a basic standard of 1 watt at a data rate of 137 Mbps. The normalized power-to-data-rate ratios for the FDM/TDM transponders are somewhat low for the 10 watt TWT TDM transmitter and somewhat high for the 4 watt solid-state FDM transmitter. If these ratios are adjusted to 1.0 by increasing the TWT power level to 12 watts and reducing the solid-state power level to 3.2 watts to equalize the performance of this transponder, the net effect on power required for this transponder is a decrease of less than 2 percent. Therefore, the power and weight comparisons presented later are still considered valid even though the FDM/TDM power is based upon a 10 watt TWT and a 4 watt solid state amplifier.

TABLE 1-1. TRAFFIC DISTRIBUTION BY COMMUNITY OF INTEREST BY CITY

Terminating Mbps / Mbps Originating	New York	Chicago	Los Angeles	San Francisco	Washington, DC	Dallas	Houston	Minn/St. Paul	Atlanta	Denver	Total Originating
New York		578	574	231	274	196	213	158	156	120	2500
Chicago	578		403	162	145	142	152	128	107	89	1906
Los Angeles	574	403		286	141	166	179	121	107	114	2090
San Francisco	231	162	286		57	65	70	49	42	45	1006
Washington, DC	274	145	140	56		48	53	39	40	29	827
Dallas	196	143	166	65	49		78	41	40	36	811
Houston	213	152	179	70	53	78		44	44	37	870
Minn/St. Paul	158	128	121	49	39	41	44		29	28	637
Atlanta	156	107	107	42	40	39	44	29		23	588
Denver	120	88	114	45	29	36	37	28	23		521
Total Terminating	2500	1906	2090	1006	827	811	870	637	588	521	

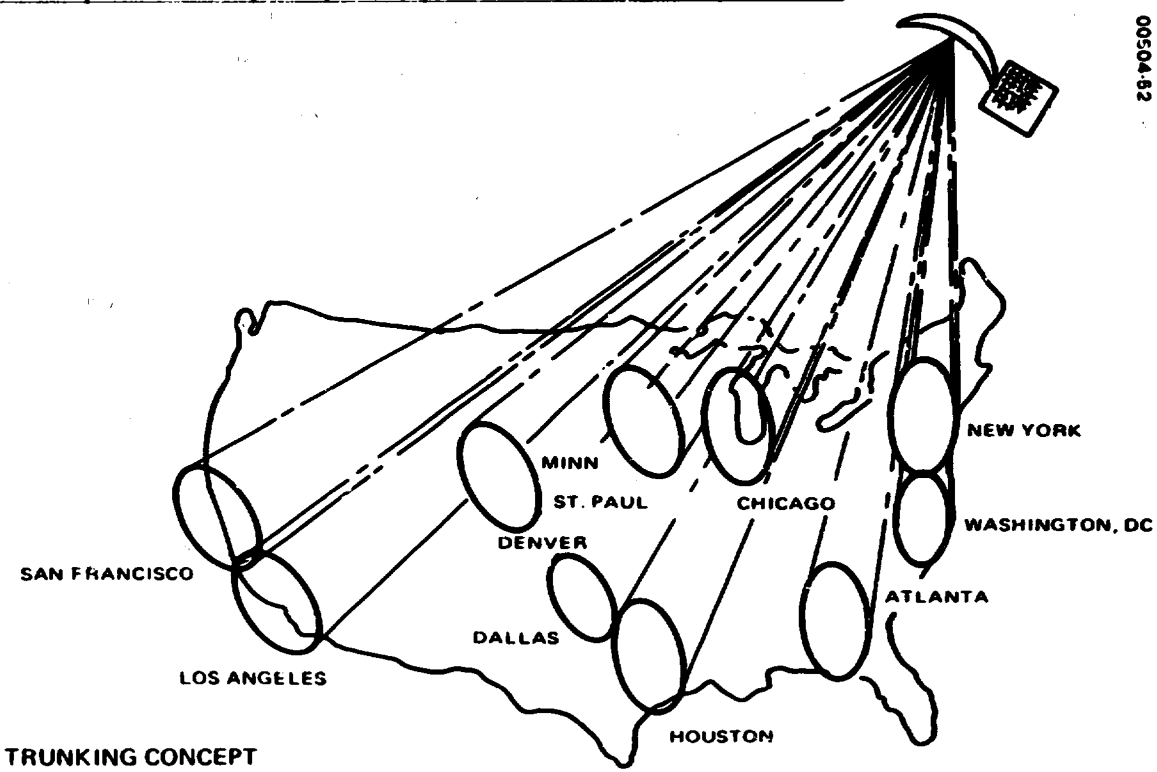


FIGURE 1-1. WIDEBAND MULTIBEAM TRUNKING CONCEPT

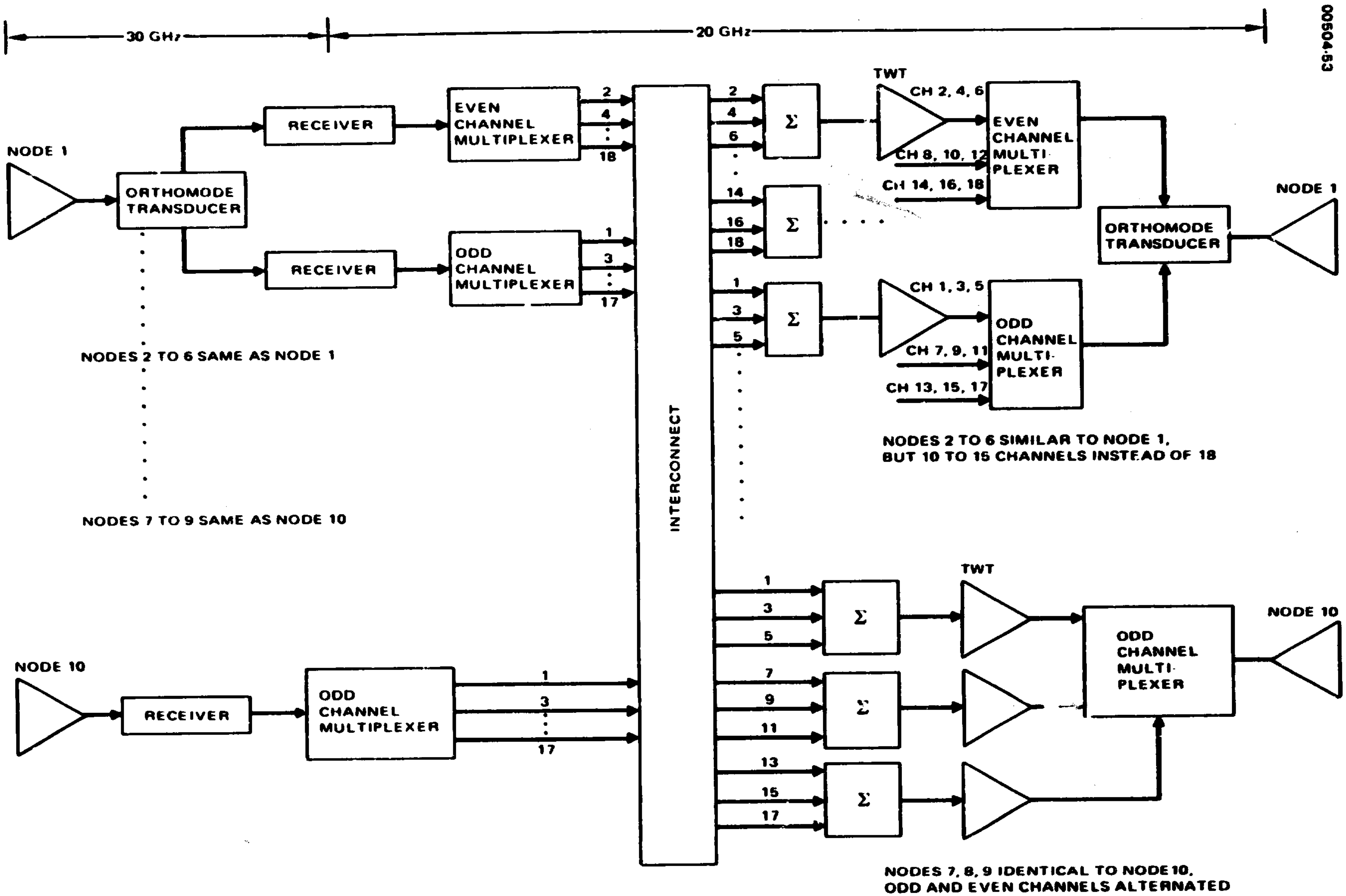


FIGURE 1-2. FDM TRANSPONDER BASELINE DESIGN

TABLE 1-2. TRANSPONDER DESIGN FEATURES

Transponder Type	Input Circuits	Channelization and Data Rates	Interconnect Switch	Output Circuits	Polarization
FDM	FET preamplifier, dual conversion receiver	137 MHz bandwidth uniform channels, 21 to 196 Mbps data rates	Interconnect at 20 GHz	Three channels/ TWT, 10 W TWT with 2.5 dB output backoff	Six highest data rate nodes, use dual polarization for alternate channels
TDM	FET preamplifier single conversion receiver	2.5 Gbps burst rate for all nodes	IF switch matrix at 5.23 GHz center frequency	Upconverter and 25 W TWT	Dual polarization not used
FDM/TDM	FET preamplifier, dual conversion receiver provides both 20 GHz and IF output	TDM – burst rate 1.096 Gbps	TDM – IF switch matrix 3.98 to 5.08 GHz	TDM – upconverter and 10 W TWT	Dual polarization for three highest data rate nodes
		FDM – 274 MHz bandwidth uniform channels, data rate 274 to 288 Mbps	FDM – interconnect at 20 GHz	FDM – solid-state 4 W power amplifier	

TABLE 1-3. TRANSPONDER PERFORMANCE COMPARISON

Transponder Configuration	BER Degradation, dB (10^{-6} BER)	Normalized Power Output to Data Rate Ratio (1.0 = 1 W/137 Mbps)
FDM	2 to 3	1.05
TDM	1.8	0.95
FDM/TDM		
FDM channel	<1.8	1.25
TDM channel	<1.8	0.83

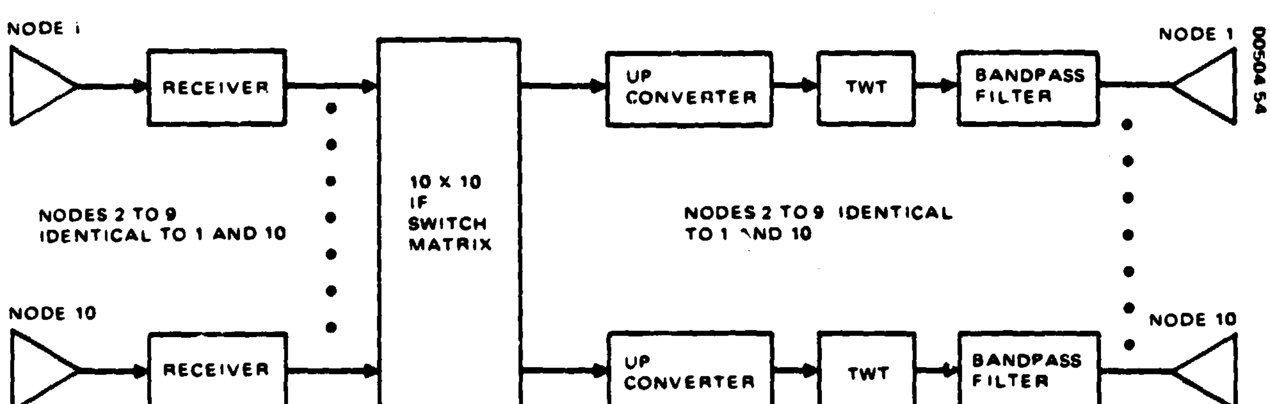


FIGURE 1-3. TDM TRANSPONDER BASELINE DESIGN

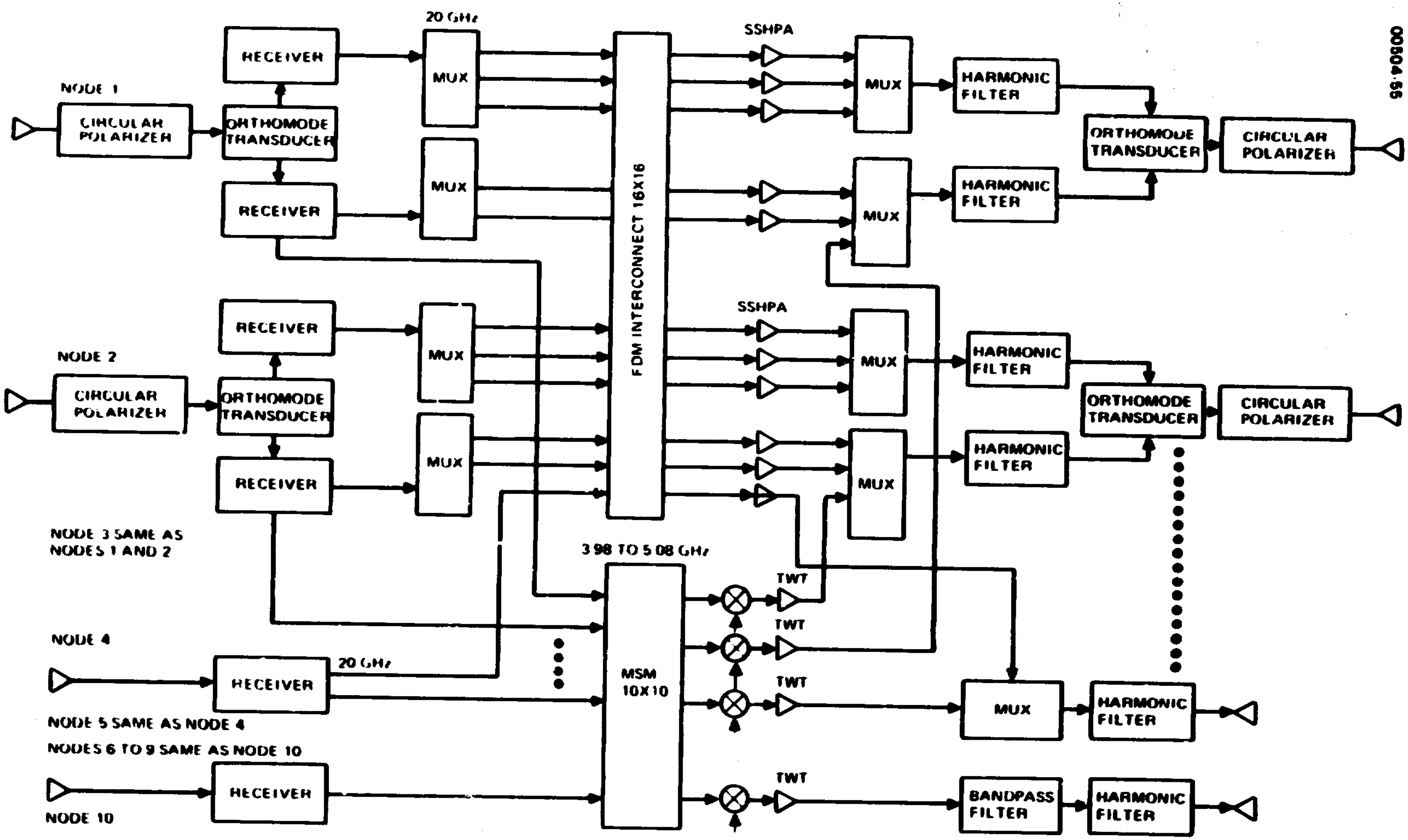


FIGURE 1-4. FDM/TDM TRANSPONDER BASELINE DESIGN

TABLE 1-4. TRANSPONDER WEIGHT, POWER, AND COST COMPARISON

Transponder Type	Weight, lb	Power, W	Cost, $M		
			Nonrecurring	Recurring	Total
FDM	587	1204	8.1	9.8	17.9
TDM	241	651	7.7	6.2	13.9
FDM/TDM	357	712	10.0	7.8	17.8

The TDM and FDM/TDM transponders have some advantage in flexibility since they can reconfigure terminal-to-terminal data rates by altering the IF switch control program stored in the switch control unit memory. The FDM repeater would require interconnect switches and subdivision of channels to achieve equivalent capability.

Table 1-4 compares the weight, power, and cost of the three transponder configurations. It must be emphasized that the data of Table 1-4 are not certain, but may be considered relatively valid. It appears that the TDM transponder has a considerable advantage over the other two types. However, this conclusion is somewhat unrealistic since use of this transponder requires 2.5 Gbps burst rate modems for the earth stations which have not been developed and the development of such modems may present formidable technical problems. If the TDM transponder design is based upon a lower burst data rate of about 500 Mbps the TDM and the hybrid FDM/TDM transponder weight, power, and cost would be nearly equal.

The FDM transponder does not appear to be well-suited for this application. The high total data rate of 2.5 Gbps required in this study, the use of uniform channelization to avoid more than one power amplifier type, and the large power penalty imposed by the use of three channels for each power amplifier that requires TWT backoff point this out. The advent of lightweight, efficient solid-state power amplifiers of 2, 4, and 8 watts would result in a more efficient FDM transponder design through the use of nonuniform channelization and a single channel per amplifier.

The conclusion reached in the thermal design study in Section 3.7 is that all three transponders are compatible with either a spinning satellite or a three-axis altitude-controlled spacecraft. Because of the dual power mode operation required of the power amplifiers, all three transponders require thermal regulation to compensate for the large change in dissipation which can occur as a function of the number of power amplifiers operated in the high power mode. The simplest method to provide the required thermal regulation would be to use heaters, since power would be available for the heaters when required. The heaters would be used when most of the transmitters are in the low power mode and during eclipse.

The technological development that would most benefit transponder design is the development of lightweight, high-efficiency, solid-state power amplifiers in the power output range of 2 to 25 watts. In addition, full realization of the potential advantage of TDM techniques will require the development of very high data rate burst modems in the range above 500 Mbps.

2. STATEMENT OF WORK CONSTRAINTS

This section lists the statement of work constraints which are to be satisfied in the design of the three types of repeaters.

The system constraints of the statement of work are as follows:

2.1 Uplink frequency	27.5 to 30.0 GHz
2.2 Downlink frequency	17.7 to 20.2 GHz
2.3 Uplink energy per bit to noise density ratio E_b/N_o)	23 dB
2.4 Downlink bit error rate (BER)	10^{-6} (no coding)
2.5 Inherent reliability (assume one spare spacecraft; Ref. Military Handbook 217B)	0.9999
2.6 Satellite lifetime	7 yr (operational)
2.7 Launch vehicle	Shuttle
2.8 Eclipse operation	Yes (housekeeping +200 W RF)
2.9 Traffic matrix	Interconnection of 10 cities to satisfy traffic requirements stated in Table 1-1 (10x10 switch; 10 antenna beams)
2.10 Channel capacities	See Table 1-1. Suggested channels: 548 Mb/s, 274 Mb/s, 137 Mb/s
2.11 RF power channel	137 Mbps channel ≅ 1 W 274 Mbps channel ≅ 1.75 W 548 Mbps channel ≅ 3.0 W (Power at antenna port; includes 10 dB of rain margin)
2.12 Beacon	Required; global coverage, ≅ 20 dBm at antenna port, coherent

2.13 Spacecraft received power	-70 to -100 dBW (at antenna port)
2.14 Spacecraft receive/transmit antenna	Separate (14 ft dia max)
2.15 Modulation	QPSK
2.16 Earth station G/T	≅39 dB
2.17 Spacecraft G/T	≅30 dB
2.18 Spurious signals	-50 dB within downlink channel -30 dB 14 to 30 GHz (ref. to lowest channel power)
2.19 Harmonic signals	-50 dB (ref. to lowest channel power)
2.20 Local oscillator stability	$\pm 5 \times 10^{-6}$
2.21 Input/output isolation	-30 dB
2.22 Distortion	Apportion total interference, distortion, and gain and phase characteristics of system so as to obtain BER -10^{-6} with a received $E_b/N_o \cong 15$ dB (earth station)
2.23 Carrier signal power limiting	Provision for uplink signal amplitude equalization for signals arriving to spacecraft from different uplink transmitters.

3. COMMON TRANSPONDER DESIGN FEATURES

A number of design features are common to all three types of transponders. This section of the report describes these common features.

3.1 RECEIVER FRONT END DESIGN

The original design chosen for this study was an image enhancement mixer, because of design simplicity. This design has a noise figure between 5.0 and 5.5 dB and cannot be expected to improve. A GaAs FET preamplifier can be expected to achieve a 4.5 dB noise figure based on projected data, and promises improved performance as better devices become available. This final report is based, therefore, upon the use of a low noise preamplifier instead of an image enhancement mixer. Impact of this change on repeater weight and power is negligible.

The receiver front end design which is described in this section includes only the preamplifier, the down-conversion mixer, and the first local oscillator multiplier chain. This circuitry is common to all three transponders and is a subassembly of the repeater receivers for all transponders. Individual differences in the balance of the receiver design are described in Sections 4 through 6 of this report. Figure 3-1 is a block diagram of the receiver low noise front end. The preamplifier consists of four identical GaAs FET low noise amplifiers on substrates mounted in a WR-28 waveguide housing. The mixer is also waveguide mounted, and consists of a signal bandpass filter, local oscillator bandpass filter, mixer diode, and IF bandpass filter. The local oscillator chain multiplies the local oscillator input at 1.96 GHz to 23.52 GHz. A two-stage amplifier is used to drive an X6 varactor diode frequency multiplier which is followed by an X2 varactor diode frequency multiplier.

Note that no bandpass filtering is provided in the signal path except at the mixer input. The preamplifier ia mounted in WR-28 waveguide, which has a waveguide cutoff frequency of 21.08 GHz. Any transponder transmitter direct leakage will be below cut-off, and a bandpass filter is not required to isolate the preamplifier from transmitter leakage.

3.1.1 Preamplifier Design

The preamplifier provides low noise amplification with sufficient gain to determine the transponder noise figure. The first consideration in design is to determine the number of preamplifier stages required.

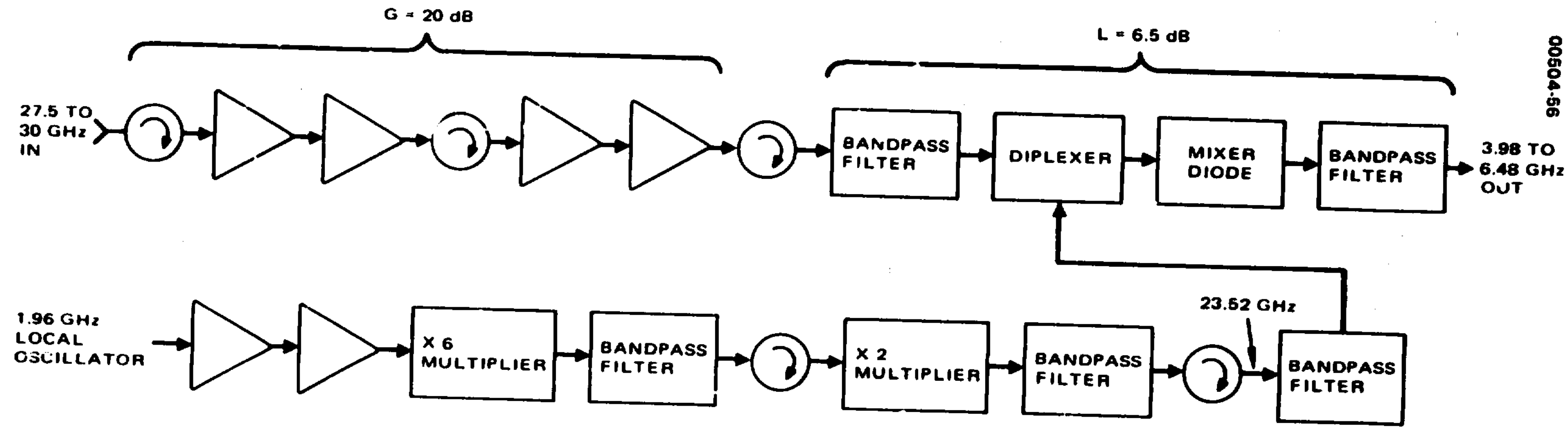


FIGURE 3-1. RECEIVER LOW NOISE FRONT END

The total receiver noise figure, NF, is given by

$$NF = L_i + 10 \log_{10} \left[F_p + \frac{\left(F_p - 1\right)\left(G^{n-1} - 1\right)}{G^{n-1}(G-1)} + \frac{F_{m-1}}{G^n} \right] \quad (1)$$

where F_p is the noise factor of each preamplifier stage, G is the power gain of each stage, n is the number of stages, F_m is the noise factor of the down conversion mixer, and L_i is the loss in dB of the isolator at the preamplifier input. The assumptions are that each preamplifier stage is nearly identical and that noise contributions after the mixer are included in the mixer noise factor.

Figure 3-2 is a family of curves which show receiver noise figure as a function of number of stages with preamplifier stage gain and noise figures as parameters. A noise figure of 8.5 dB (noise factor of 7.08) is used for the down-conversion mixer and the following IF stages. The mixer noise figure is based on a 6 dB conversion loss and a 2.5 dB IF noise figure. Isolator insertion loss is 0.2 dB.

Figure 3-2 indicates that there is little performance advantage in a five-stage preamplifier. A four-stage preamplifier does show a noticeable performance improvement over a three-stage preamplifier, however, particularly for stage gains of 5 dB or less. For this reason, a four-stage preamplifier has been chosen as the baseline design.

Figure 3-3 shows a two-stage microwave integrated circuit (MIC) GaAs FET MIC amplifier mounted in WR-28 waveguide. Two coaxial amplifiers of this configuration, separated by a WR-28 waveguide isolator is provided at both input and output of the four-stage preamplifier to ensure low VSWR at the interfaces. Figure 3-4 is a photograph of a disassembled single stage K_a-band amplifier.

Based on near term expected improvements in the performance of GaAs FET devices at 27.5 to 30.0 GHz, a single-stage amplifier should have a noise figure of about 3.25 dB and an associated gain of 5 dB at the optimum noise figure impedance match. Equation (1) gives a receiver noise figure of 4.5 dB for such a device.

3.1.2 Mixer Design

The mixer configuration is shown in Figure 3-5. The input signal is filtered through a three-section waveguide filter which has a bandwidth of 2.5 GHz. An E plane tee junction combines the local oscillator and signal and is coupled by a two-step waveguide transformer to the mixer diode. An IF bandpass filter filters the output signal. The local oscillator filter and signal bandpass filter optimize mixer conversion loss.

3.1.3 Local Oscillator Chain

The local oscillator chain multiplies the local oscillator reference frequency of 1.96 GHz to 23.52 GHz. Figure 3-6 is a block diagram of this subassembly. Figure 3-6 also gives a stage-by-stage breakdown of gain and level. Figure 3-6 also shows the estimated level of un wanted harmnics. Harmonic suppression is performed by narrowband, multiple pole filters which follow the 2 multipliers.

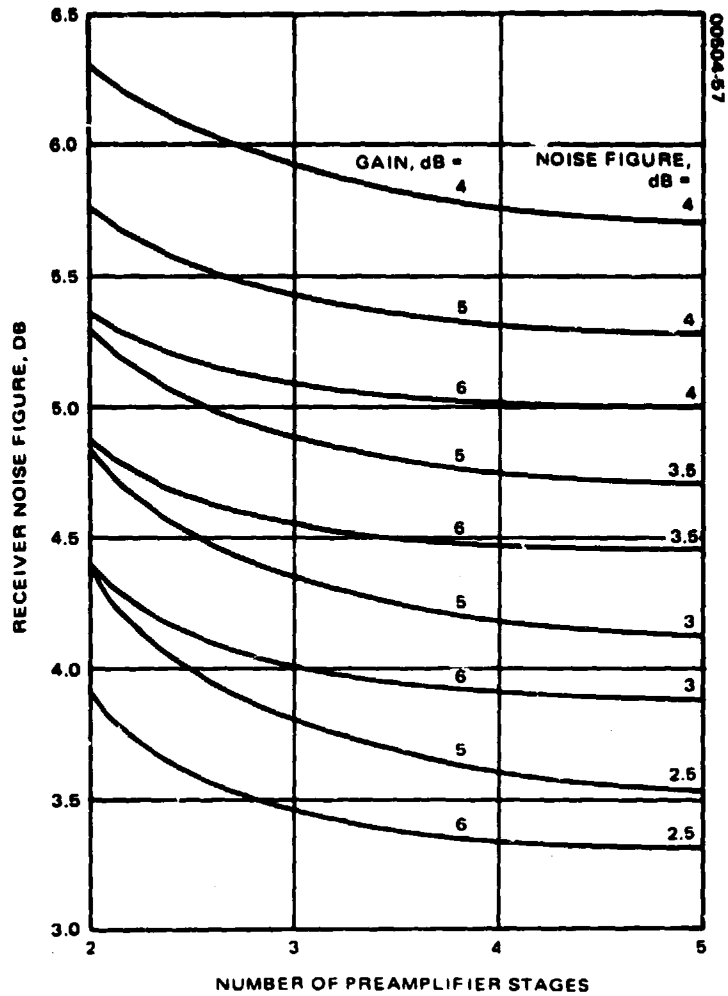


FIGURE 3-2. RECEIVER NOISE FIGURE AS FUNCTION OF PREAMPLIFIER STAGES AND STAGE GAIN AND NOISE FIGURE

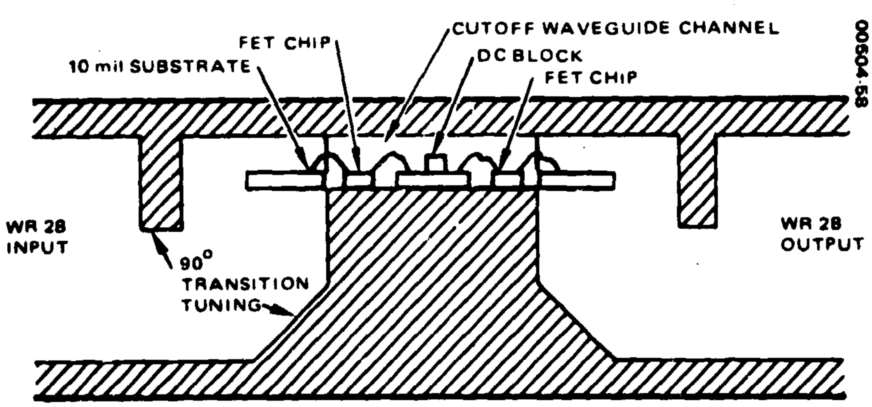


FIGURE 3-3. PREAMPLIFIER CONFIGURATION

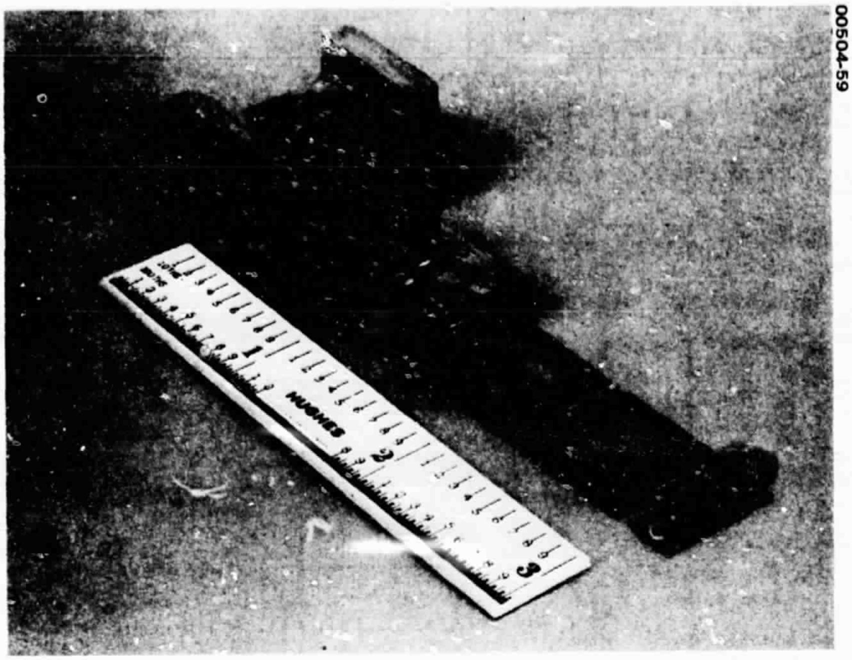


FIGURE 3-4. DISASSEMBLED K_a BAND SINGLE STAGE FET AMPLIFIER

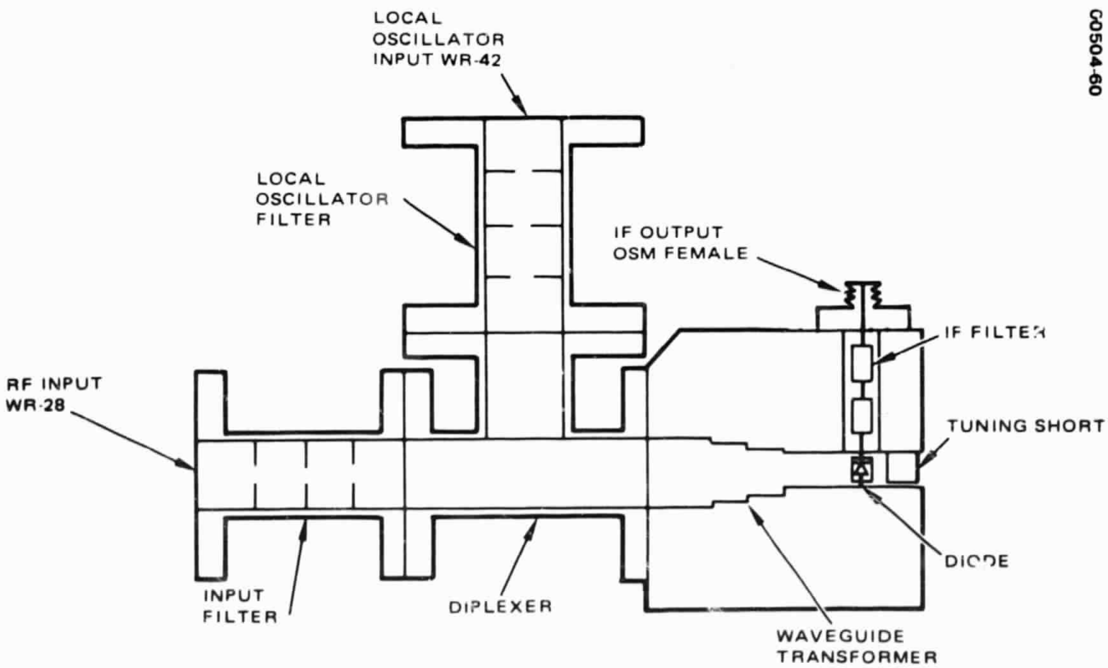
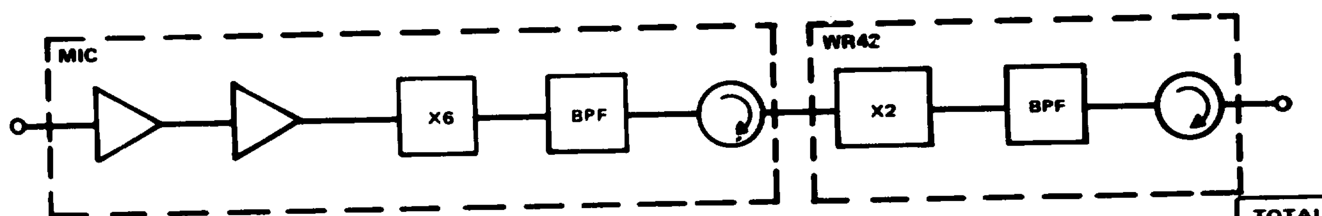


FIGURE 3-5. MIXER CONFIGURATION



																	TOTAL
FREQUENCY, GHz	1.96				1.96		11.76				11.76		23.52				23.52
GAIN, dB		13		10		−9		−1		−0.5		−4		−0.3		−0.3	7.9
POWER, dBm	2.1		15.1		25.1		16.1		15.1		14.6		10.6		10.3		10
HARMONICS, dBc									−60						−130		−130

FIGURE 3-6. LOCAL OSCILLATOR CHAIN MIC/WAVEGUIDE BLOCK DIAGRAM AND PERFORMANCE TABLE

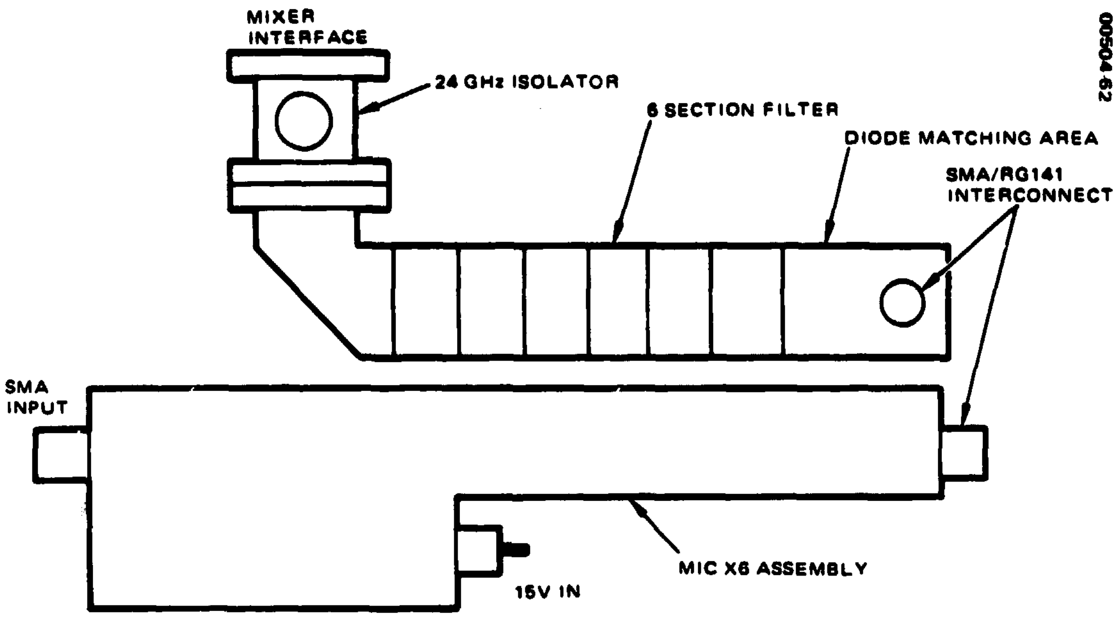


FIGURE 3-7. LOCAL OSCILLATOR LAYOUT

A two-stage transistor amplifier drives the X6 silicon varactor diode frequency multiplier. This is followed by a three-section bandpass filter and an isolator. This section of the multiplier is fabricated using a MIC. The final X2 frequency multiplier uses a GaAs varactor diode mounted in WR-42 waveguide and a six-section waveguide bandpass filter. Figure 3-7 shows the packaging concept of the local oscillator multiplier.

3.2 INPUT AND OUTPUT WAVEGUIDE SELECTION

Because of the very high frequencies and the fact that interconnection length from the repeater to the satellite feeds may be substantial, the choice of waveguide is important. Table 3-1 shows theoretical and practical losses per foot for four types of waveguide at both the upper and lower frequencies of the transponder transmit and receive bands.

The WR-42 and WR-51 size waveguides are the possible choices for the transmit bands. WR-42 has the advantage of being a standard waveguide. However, the WR-42 waveguide has much higher loss and also has a recommended lower operating frequency of 18.0 GHz, slightly above the 17.7 GHz minimum frequency of the transmit band. The only problem in using the lower-loss WR-51 waveguide is that adapters will be required in many cases to connect standard commercial test equipment. The use of WR-51 is recommended because of its substantially lower loss.

The same argument applies to the choice of receiver waveguide, and the use of the WR-34 waveguide is recommended.

TABLE 3-1. WAVEGUIDE CHOICE

Type	Band, GHz	Loss/Foot, dB	
		Theoretical	Practical
WR-42 (Common)	18 to 26.5	0.140 at 17.7 GHz 0.117 at 20.2 GHz	0.21 at 17.7 GHz 0.18 at 20.2 GHz
WR-51 (Special)	15 to 22	0.071 at 17.7 GHz 0.065 at 20.2 GHz	0.11 at 17.7 GHz 0.10 at 20.2 GHz
WR-28 (Common)	26.5 to 40	0.211 at 27.5 GHz 0.187 at 30 GHz	0.32 at 27.5 GHz 0.28 at 30 GHz
WR-34 (Special)	22 to 33	0.127 at 27.5 GHz 0.120 at 30 GHz	0.19 at 27.5 GHz 0.18 at 30 GHz

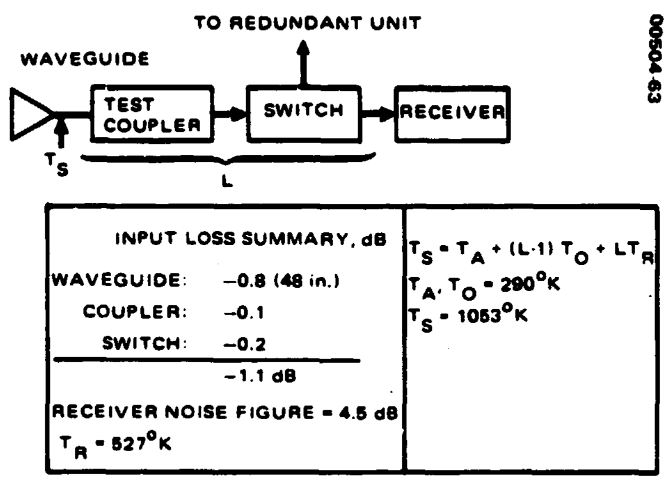


FIGURE 3-8. REPEATER RECEIVER NOISE TEMPERATURE

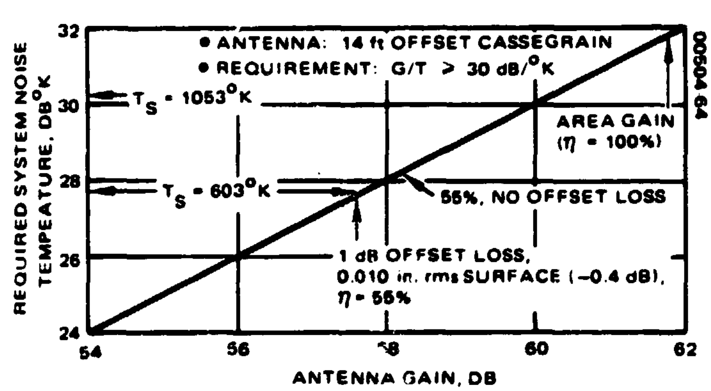


FIGURE 3-9. SYSTEM NOISE TEMPERATURE VERSUS SPACECRAFT ANTENNA GAIN (27.5 GHz)

3.3 REPEATER NOISE TEMPERATURE AND G/T

The statement of work specifies a spacecraft G/T of 30 dB/°K, based upon a 14-foot diameter antenna. The repeater receiver noise figure is 4.5 dB. Figure 3-8 shows the interface between the receiver and satellite and shows the computation of system noise temperature referenced to the satellite antenna feed. A four-foot waveguide run is assumed between the satellite antenna and the repeater receiver. Total system noise temperature is 1053°K. Figure 3-9 is a plot of the system noise temperature required to achieve a G/T of 30 dB/°K versus antenna gain. The expected antenna gain of about 57.8 dB is based upon a basic antenna efficiency of 55 percent, a surface tolerance loss of 0.4 dB, and an offset pointing loss of 1 dB. The system noise temperature required for a 30 dB/°K G/T for this case is 603°K, which is equivalent to a required repeater noise figure of 2.1 dB. The G/T with a transponder receiver noise figure of 4.5 dB is 27.6 dB/°K. The specified G/T of 30 dB/°K does not appear to be achievable with current technology.

3.4 FREQUENCY CONVERSION

All three repeater designs require frequency conversion. A dual-frequency conversion scheme is required for the TDM and TDM/FDM transponders since the microwave switch matrix implementation is not practical at the repeater output frequency range. Single conversion is also undesirable for the FDM, since low side conversion results in the second harmonic of the local oscillator appearing at 19.6 GHz in the output frequency band. A further constraint in the frequency conversion process is to avoid inversion of the frequency spectrum in the repeater.

Figure 3-10 shows three possible frequency conversion plans and Table 3-2 compares these plans. The third plan has been selected as the baseline frequency conversion scheme for all repeaters. This plan has the further advantage of using a common, local oscillator source for both the down-conversion and up-conversion local oscillator frequency multipliers, as shown in Figure 3-11.

Final local oscillator frequency multiplication is included as part of the mixer package in all repeaters, since it was considered impractical to route the final local oscillator frequencies without excessive cable loss or bulky waveguide interconnects.

3.5 LOCAL OSCILLATOR SOURCE DESIGN

Figure 3-12 is a block diagram of the local oscillator source. The crystal oscillator frequency is set relatively high to reduce the frequency multiplication ratio. This also reduces the phase noise generated in the local oscillator frequency multiplier chain. The crystal oscillator is housed in a temperature-controlled enclosure to ensure good frequency stability. The statement of work requires a local oscillator stability of x 10^{-6}. This degree of stability can be met with the baseline design with temperature control of the crystal oscillator.

The crystal oscillator is frequency-multiplied by 16 in two cascaded X4 frequency multipliers. Each X4 multiplier stage is followed by a bandpass filter which attenuates undesired harmonics generated in the multipliers.

TABLE 3-2. COMPARISON OF NONINVERTING FREQUENCY PLANS

Plan	Input Spurs	Output Spurs	Other
Noninverting, 4.9 GHz IF	4X LO – 3X signal	2X LO – 2X signal, 3X signal	2X LO in receive band
Double inversion	5X Signal – 4X LO	5X LO – 2X signal, 3X signal	Higher frequency LO signals require more power
Noninverting, 5.23 GHz IF	4X LO – 3X signal	2X LO – 2X signal, 3X signal	

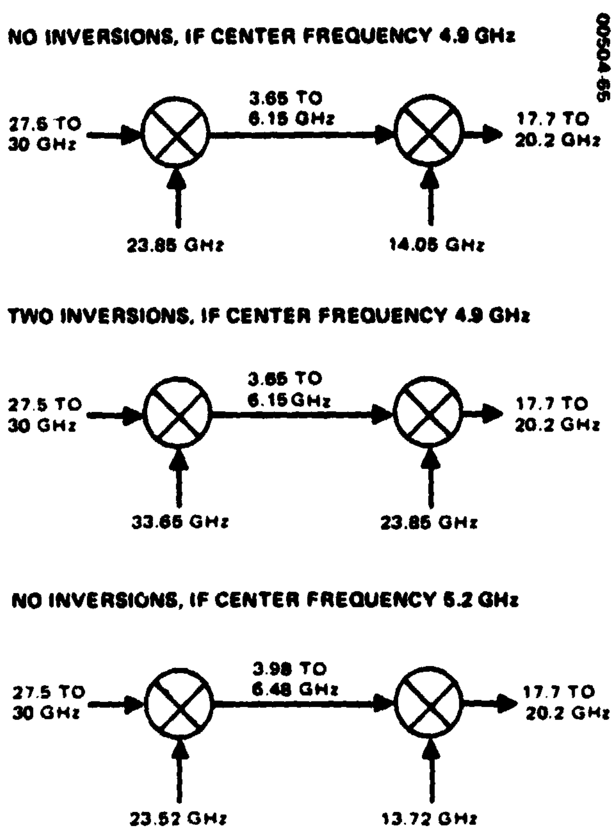
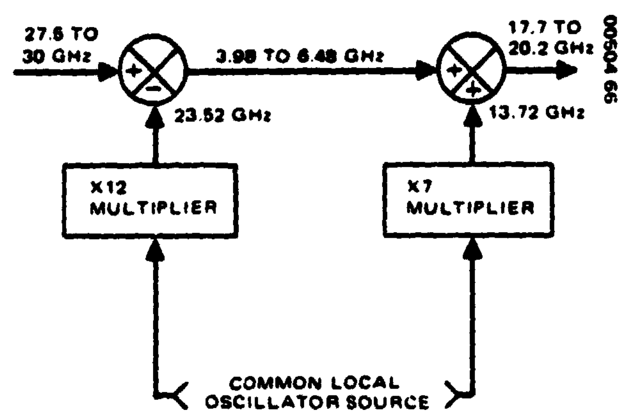


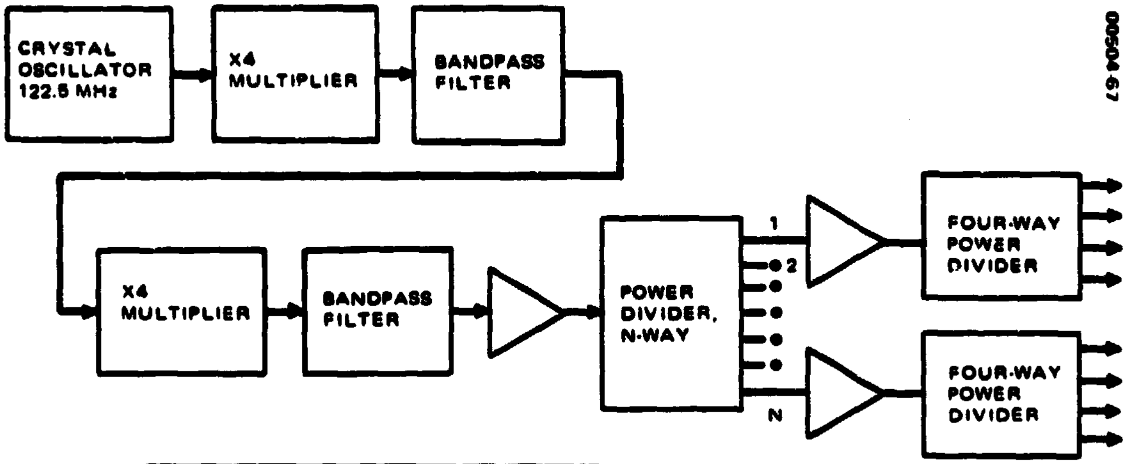
FIGURE 3-10. NONINVERTING DOUBLE CONVERSION FREQUENCY PLANS



$$f_{LO} = \frac{27.5 - 17.7}{12-7} = 1.96 \text{ GHz}$$

FIGURE 3-11. UPCONVERTER AND DOWNCONVERTER CONCEPT

The 1.96 GHz output of the X16 multiplier is then amplified and power split by a power divider. The number of power-divided outputs at this point is a function of the repeater type as shown by the table in Figure 3-12. The signal is then amplified again and power split by a four-way power divider. The local oscillator source provides 32 outputs for the FDM repeater, 36 outputs for the TDM repeater, and 42 outputs for the FDM/TDM repeater.



TRANSPONDER TYPE	N
FDM	8
TDM	9
TDM/FDM	11

FIGURE 3-12. LOCAL OSCILLATOR SOURCE BLOCK DIAGRAM

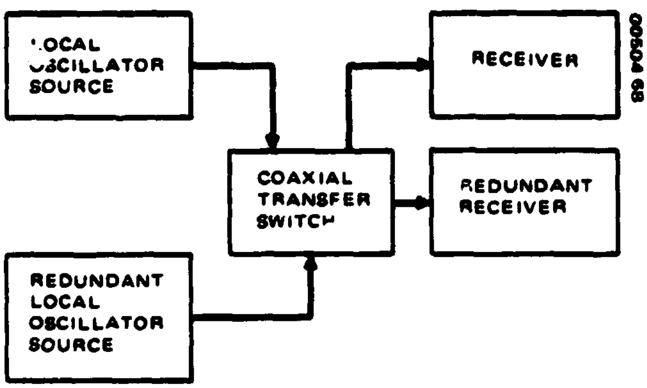


FIGURE 3-13. LOCAL OSCILLATOR/RECEIVER REDUNDANCY CONCEPT

3.6 LOCAL OSCILLATOR REDUNDANCY

Figure 3-13 shows the local oscillator and receiver redundancy concept. Either one of the two local oscillator sources can be selected to supply either of the two redundant receivers. Coaxial transfer switches are provided for each output of the local oscillator source. Both local oscillator sources always have dc power on so that individual outputs can be selected independently to achieve maximum redundancy.

3.7 THERMAL DESIGN CONSIDERATION

3.7.1 Summary

A thermal study was performed to assess the feasibility of integrating a 30/20 GHz transponder system onto typical communications satellites. The approach taken in this study was to consider various thermal design approaches from a system (macroscopic) level. Thermal integration of the following three transponder configuration was considered: (1) FDM; (2) TDM; and (3) FDM/TDM hybrid. The communications satellite configurations considered for the application were (1) 3-axis stabilized configuration (Figure 3-14) and (2) spin stabilized configuration (Figure 3-14). Both the three-axis and spinner configuration are currently being utilized for geosynchronous communications.

Results of the study indicate the thermal feasibility of integrating a 30/20 GHz transponder system onto the above mentioned communications satellite configurations. A summary of the feasibility study is shown in Table 3-3. In regards to the thermal design three basic thermal control elements are required in order to achieve feasibility. The first basic thermal control element required is a radiator of enough area to eject the heat from the transponder. The second element is a temperature regulation device. It will maintain the temperature of the transponder within operating limits, since the heat dissipation has a large variation created by its operating modes. The third element is a device consisting of conduction doublers or heat pipes used to dissipate the heat locally from under the TWT or SSPA. This would maintain the unit temperature within required operating limits.

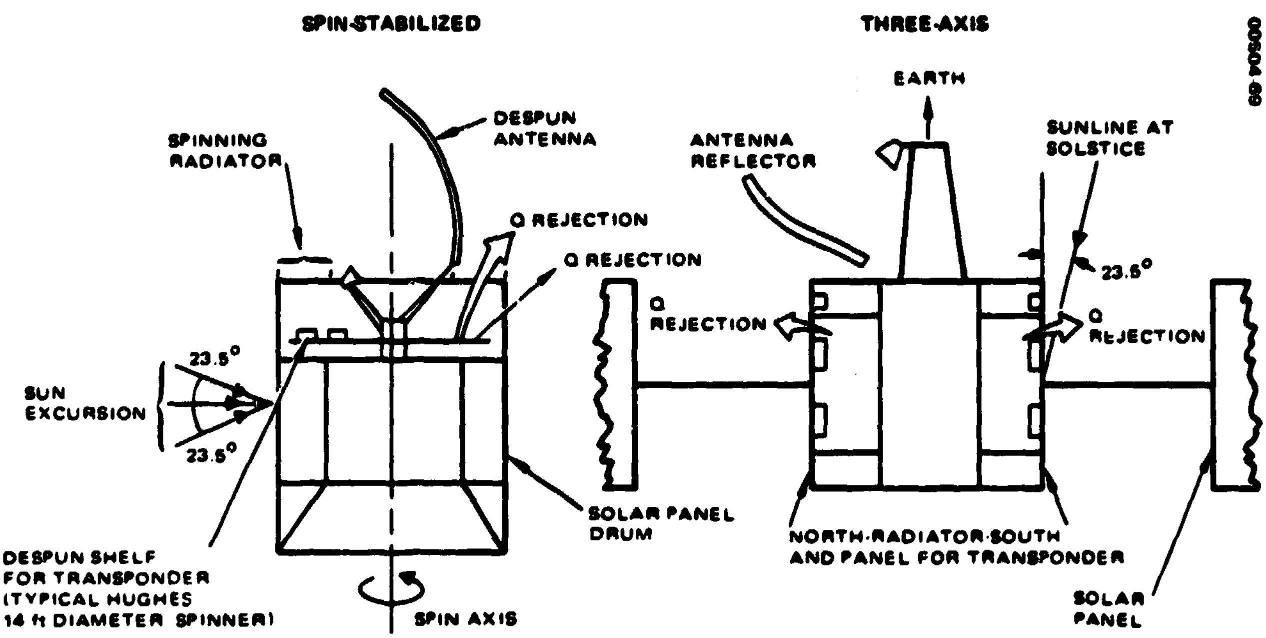


FIGURE 3-14. THERMAL INTEGRATION OF TRANSPONDER ONTO TYPICAL SPACECRAFT – SPACECRAFT TYPES AND HEAT REJECTION CONCEPT

TABLE 3-3. THERMAL CONTROL FEASIBILITY SUMMARY

Configuration	Heat Dissipation Range, W	Desired Bulk Temperature Range, °F	Heat Reject Tech, Q/A, W/ft²	Radiator Area Required ft²	Temperature Regulation Required?	Transponder Unit Thermal Control
FDM						
14 ft diameter spinner	440 to 1300 $\frac{QMAX}{QMIN} = 2.5$	80 to 60(1)	Indirect 16 W/ft² max.	81	Yes – heaters at 270 W min.	Conduction doublers
Three-axis	440 to 1300	80 to 50	Direct 29 W/ft² max.	42	Yes – Options Heater at 496 W min*. Louvers** Heat pipe	 Doubler Doubler Heat pipe
TDM						
14 ft diameter spinner	194 to 566 $\frac{QMAX}{QMIN} = 2.9$	90 to 60(1)	Indirect 19 W/ft² max.	30 (Unit size dictates)	Yes – Heaters at 104 W min*	Conduction doublers
Three-axis	194 to 566	90 to 50	Direct 31 W/ft² max.	18	Yes – Options Heaters at 194 W min* Louvers** Heat pipe	 Doubler Doubler Heat pipe
TDM/FDM						
14 ft diameter spinner	290 to 713 $\frac{QMAX}{QMIN} = 2.5$	80 to 60(1)	Indirect 16 W/ft² max.	45 (Unit size dictates)	Yes Heaters at 156 W min*	Conduction doublers
Three-axis	290 to 713	80 to 50	Direct 29 W/ft² max.	25	Yes – Options Heaters at 290 W min* Louvers** Heat pipe	 Doublers Doublers Heat pipe

*Uses of available power when minimum RF configuration is used.
**Radiator area required increases by factor of 2.
(1)Pre-eclipse

3.7.2 Considerations for Thermal Feasibility

The following considerations are necessary ingredients on a systems level to obtain thermal feasibility.

Environment

The external environment in which the transponder is exposed depends upon the satellite configuration and the orbital environment of the satellite. For this study the three-axis and spinner configurations were investigated. In addition, the orbital environment was assumed to be a geosynchronous equatorial orbit. In a geosynchronous orbit a three-axis satellite will have its antenna pointed toward earth as shown in Figure 3-14. The indicators are usually located on the north and south-pointing faces to min-

TABLE 3-4. THERMAL HEAT DISSIPATION
FDM CONFIGURATION

Solid-state high power amplifier transponder

- High power mode Q = 20 W at 2.0 W RF output
- Low power mode Q = 5 W at 0.2 W RF output

Operating modes

- Number operating: 114 channels
- Modes —Max Q condition: 60-channels high power
  54-channels low power

  -- Min Q condition: 114-channels low power

Total heat dissipation

Component	Number Channels Operating	Q = Heat Dissipation, W	
		24 at High Power/ 16 at Low Power	40 at Low Power
Traveling wave tube amplifier	40	1134	364
Receivers	10	70	70
Insertion losses at 2.05 dB		96	7
Total Q		1300	441

TABLE 3-5. THERMAL HEAT DISSIPATION
TDM CONFIGURATION

Traveling wave tube amplifier

- High power mode Q = 65 W at 25 W RF output
- Low power mode Q = 11.5 W at 2.5 W RF output

Operating modes

- Number operating: 10 channels
- Modes – Max Q condition: 6-channel high power
  4-channel low power

  – Min Q condition: 10-channel low power

Total heat dissipation

Component	Number Channels Operating	Q = Heat Dissipation, W	
		6 at High Power 4 at Low Power	10 at Low Power
Traveling wave tube amplifier	10	436	115
Receivers	10	[illegible]	70
Insertion losses at 2.5 dB		[illegible]	9
Total Q		566	194

imize solar exposure. The maximum sun angle relative to the radiation is 23.5° during the solstice seasons. The rest of the satellite body is insulated to minimize solar heat input and radiation to space. On a spinner configuration, the antenna is also pointed toward earth with the satellite spin axis along a north/south line. The solar excursion angle, as shown in Figure 3-14, is a 3.5° angle with the sun vector perpendicular to the spin axis at equinox. With this attitude relative to the sun vector, the solar panel with its radiation properties provides a rather benign "room temperature" environment for the transponder. The primary heat rejection for the transponder is provided by a radiator located on the forward end of the satellite, as shown in Figure 3-14.

The internal environment of the transponder is also an important design consideration. Fundamentally, the heat dissipation dictates the area required for the heat rejection radiator in addition to the physical layout of the transponder units. The heat dissipation is a function of the inherent characteristic of the unit and the operating mode in which the transponder is exercised. The thermal heat dissipation and corresponding operating modes assumed for this study are shown in the following tables. Table 3-4 shows the FDM configuration, Table 3-5 the FDM configuration, and Table 3-6 the hybrid configuration.

TABLE 3-6. THERMAL HEAT DISSIPATION
FDM/TDM CONFIGURATION

Traveling wave tube transponder – TDM

- High power mode Q = 30 W at 10 W RF output
- Low power mode Q = 6W at 1 W RF output

Solid-state high power amplifier – FDM

- High power mode Q = 40 W at 4 W RF output
- Low power mode Q = 9.6 W at 0.4 W RF output

Operating modes

- Number channels operating at TDM: 10
  FDM: 16
- Modes – Max Q condition TDM: 5-channel high power, 5-channel low power
  FDM: 9-channel high power, 7-channel low power
  – Min Q condition TDM: 10-channel low power
  FDM: 16-channel low power

Component	Number Channels Operating	Q = Heat Dissipation, W	
		Max Q	Min Q
TDM			
• Traveling wave tube amplifier	10	180	60
• Insertion losses		21	4
FDM			
• Solid-state high power amplifier	16	427	154
• Insertion losses		15	2
Receiver	10	70	70
Total Q		713	290

Temperature Requirements

In order to fabricate a viable thermal design, transponder unit temperature goals must be established. Table 3-7 gives an indication of typical unit temperature goals. Since this study was based on a systems level approach, these unit temperatures were translated into an overall unit mounting shelf (bulk) temperature. The bulk temperatures selected were based on past experience. They differ slightly, depending on satellite configuration and whether the transponder has TWTs or SSPA, as shown in Table 3-3.

Transponder Arrangement

The physical arrangement of the transponder is a thermal design consideration. It must provide enough local area on the mounting shelf to dissipate the heat away from each unit. Obviously this may affect the minimization of RF loss in waveguides or coax. An exact tradeoff of thermal control weight and RF loss can only be accomplished in a detailed design study. Consequently it was not addressed in this study.

Heat Rejection Techniques

The heat that is dissipated by the transponder unit must be rejected to space in order to maintain proper temperatures. On a systems level there are two fundamental

TABLE 3-7. THERMAL REQUIREMENTS ON MAJOR TRANSPONDER COMPONENTS

Component	Temperature Design Goals, °F	
	Minimum	Maximum
Traveling wave tube collector	20	150
Electronic power converter	0	110
Solid-state high power amplifier	20	113
Receivers	60	121
Output Filters	20	130

TABLE 3-8. CONSIDERATIONS FOR THERMAL FEASIBILITY HEAT REJECTION TECHNIQUES

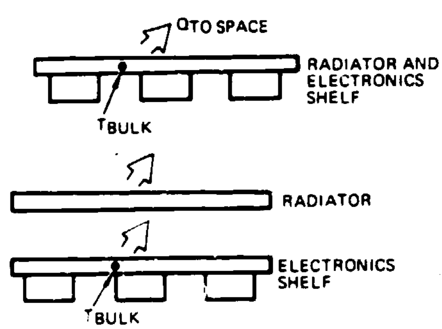
System level (bulk)

Direct radiator

- Heat rejection capability 20 to 30 W/ft^2
- Typical three-axis spacecraft

Indirect radiator

- Heat rejection capability 10 to 20 W/ft^2
- Typical spinner spacecraft



approaches, as shown in Table 3-8. In the direct radiator approach, the units reject their heat directly to space. This is the more efficient approach because it requires less radiator area. This method is commonly used on the three-axis configuration shown in Figure 3-14. However, the direct radiator is sensitive to external and internal environmental changes from a temperature standpoint.

The indirect radiator approach shown in Table 3-8 is commonly used in the spinner configuration. An intermediate media between the unit mounting shelf and the radiator makes this approach less efficient. However, the unit temperature swing is less severe, since it is dampened by the media.

Once a radiator approach is selected, most likely from the satellite configurations, the reduction of the hot spot under the unit must be considered. Two approaches to reduction were considered in this study.

The first was a conduction doubler. The conduction doubler is a plate of adequate thickness under the unit which acts as a conduction fin to dissipate unit heat. The advantages of a conduction doubler are (1) reliability, (2) passivity, and (3) simplicity. The disadvantage is weight compared to heat pipes for units that dissipate more than 100 watts.

The second approach uses heat pipes to dissipate unit heat. Their advantage is high thermal conductivity. With variable conductance heat pipes the transponder unit can be maintained at a relatively constant temperature, although heat dissipation may vary as a function of operating mode. The disadvantage is the complex integration of heat pipes as opposed to the simplicity of a conduction doubler.

Temperature Regulation Devices

The transponders ability to operate in a high or low power mode can produce a large swing in heat dissipation. It is necessary to compensate for this swing by means of a temperature regulation device. Under consideration are the three devices shown in Table 3-9. A temperature regulation device is also required to prevent violation of the lower temperature limit of the transponder units.

The first device consists of resistance heaters which compensate for lowered dissipation. This device is low in cost and weight. There is satellite power available since the heaters compensate for the reduction in power required for the transponder.

The second device consists of shield louvers on the radiator. The blades of the louvers open or close as a function of temperature. The closing of the blades decreases the heat ejection capability of the radiator. Louvers are usually used on direct radiator applications such as the three-axis satellite configuration. Because of the mechanical actuation of the blades, reliability becomes a design parameter. In addition, the radiator area must be approximately doubled because the radiator with shielded louvers behaves like a indirect radiator. A shield must be provided over the louver blade assembly in order to prevent sunlight from entering the assembly.

The third device uses variable conductance heat pipes to regulate heat dissipation. The condenser length of the heat pipe that controls the amount of heat ejected by the radiator varies by means of a feedback loop to the transponder temperature. A synopsis of the advantages and disadvantages of these three temperature regulation devices is shown in Table 3-9.

• MAKEUP HEATERS REQUIRED TO MAINTAIN BULK TEMPERATURE BETWEEN 60 TO 80°F

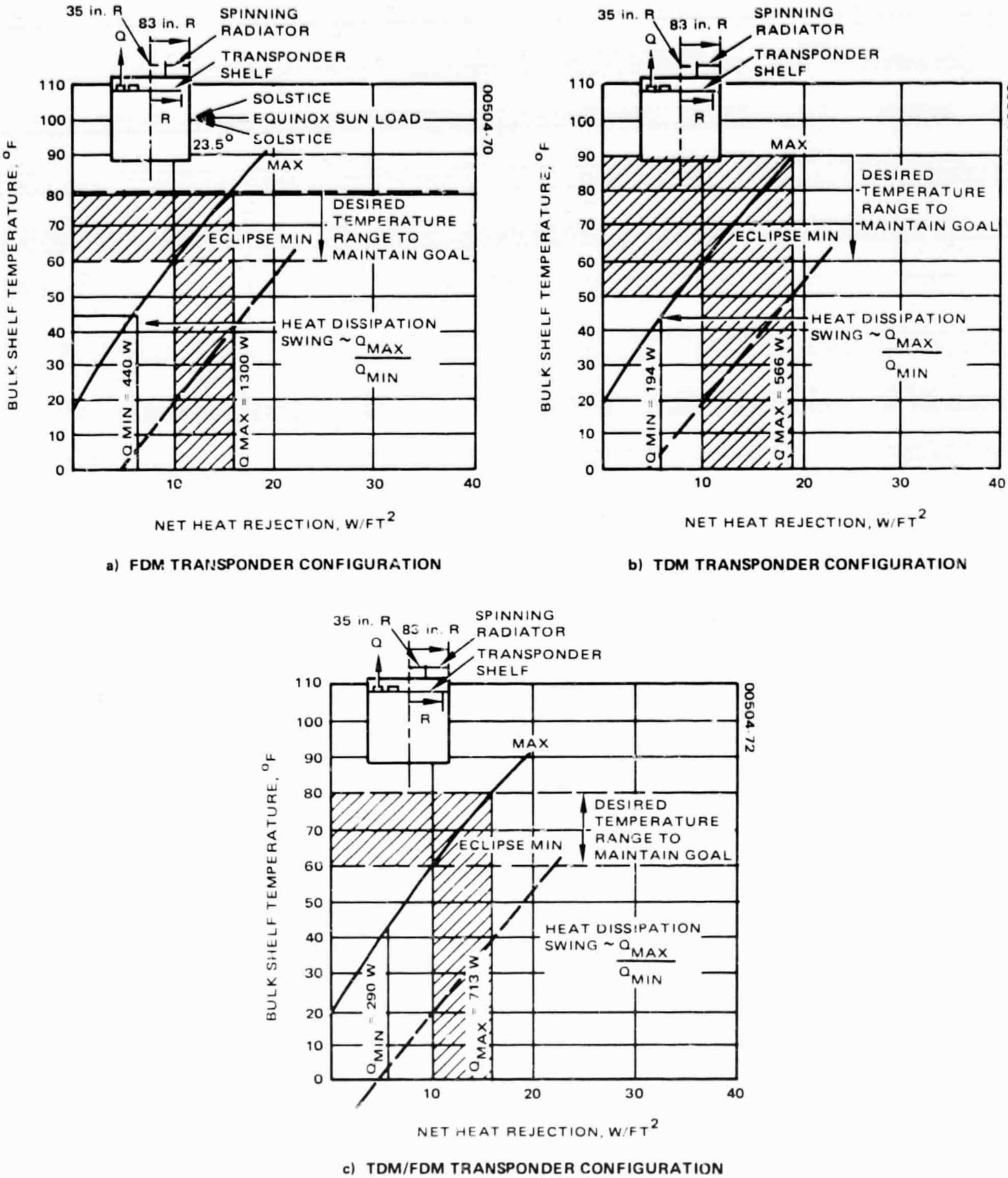


FIGURE 3-15. TYPICAL 14-FOOT DIAMETER SPINNER SPACECRAFT THERMAL PERFORMANCE

TABLE 3-9. TEMPERATURE REGULATION DEVICES

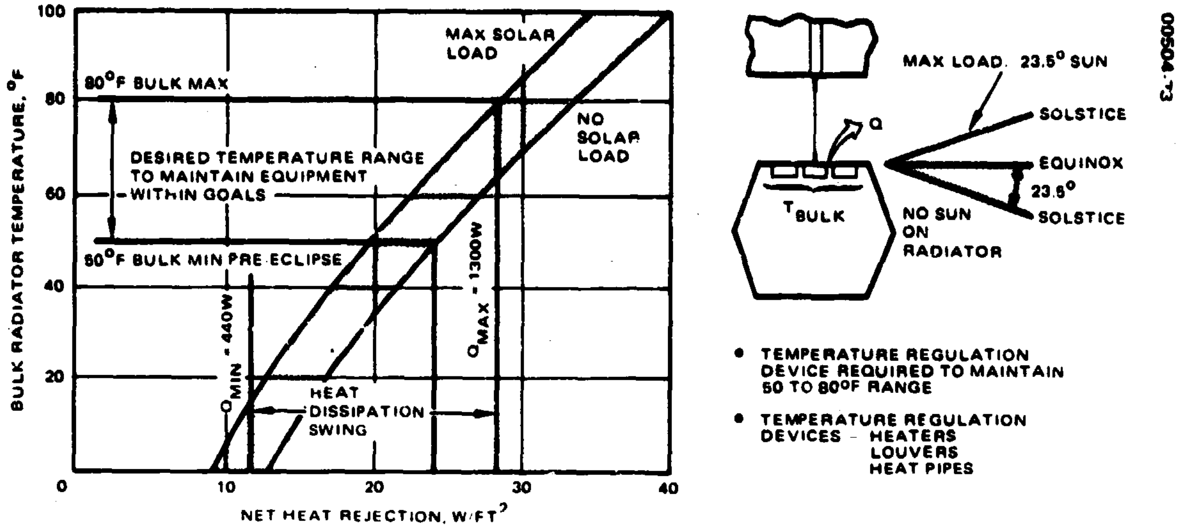
- To maintain temperature within desired range with large heat dissipation swing

	Make-Up Heaters	Shielded Louvers	Variable Conductance Heat Pipes
Function	Maintain near thermal dissipation	Decouple radiators from space as temperature drops	Conduct unit thermal dissipation to radiators and reduce radiator area as power dissipation decreases
Primary use	To keep available solar panel power in satellite	Medium low power density temperature regulation	High power density applications in place of thermal doublers
Advantages	Least costly method of temperature regulation: power available in low power modes	Simple; reliable; easily adapted to design; decouples power and thermal; 0.10 lb/W regulation	Combines high thermal conductivity with temperature regulation
Disadvantages	Makes thermal control dependent on power system and switching logic	Must be shielded so maximum radiation capability = 14 W/ft ^{2}	Complex integration into satellite structural design

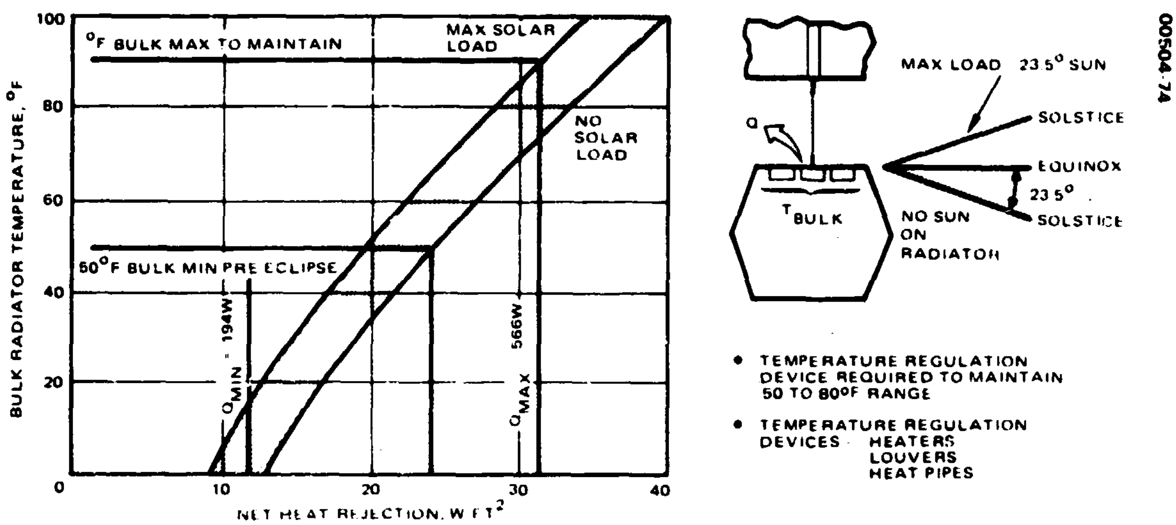
3.7.3 Thermal Performance

The systems level (bulk) performance of the transponder configurations (FDM, TDM, and hybrid) were considered for a three-axis and a spinner satellite configuration. The thermal performance was based on the bulk temperature of a shelf in which transponders are typically mounted as a function of heat dissipation (watts) per square foot of shelf area. For the three-axis configuration, because the shelf and radiator (direct) are physically the same, a determination of the radiator area required can be accomplished for given heat dissipations. For the spinner the radiator area is fixed as shown in Figure 3-15a and only a variation in shelf area as a function of heat dissipation can be determined. The maximum shelf area for this particular spinner configuration is the diameter of the solar panel (14 feet). In reference to Figure 3-15a, the crosshatched area between 60°F and 80°F represents the desired boundaries of bulk temperature. By sizing the shelf area to accommodate the maximum heat dissipation at the high power mode of the FDM transponder configuration, the low power mode heat dissipation violates the lower bulk limit of 60°F. Consequently, make-up heaters or a similar temperature regulation device would be required. The reason for the 60°F low temperature limit is that during an equinox eclipse of 72 minutes the transponder unit will cool, changing temperature by as much as 40°F, as shown in Figure 3-15a. In order to maintain minimum transponder unit temperatures above the required temperature limit, a minimum of 60°F prior to the eclipse is required. The conclusive performance curves for the TDM and the hybrid are shown in Figures 3-15b and 3-15c, respectively. It should be noted that the upper bulk temperature limit for the TDM transponder configuration is 90°F rather than 80°F, because the TWT transponder can tolerate higher temperatures than the SSPA transponder.

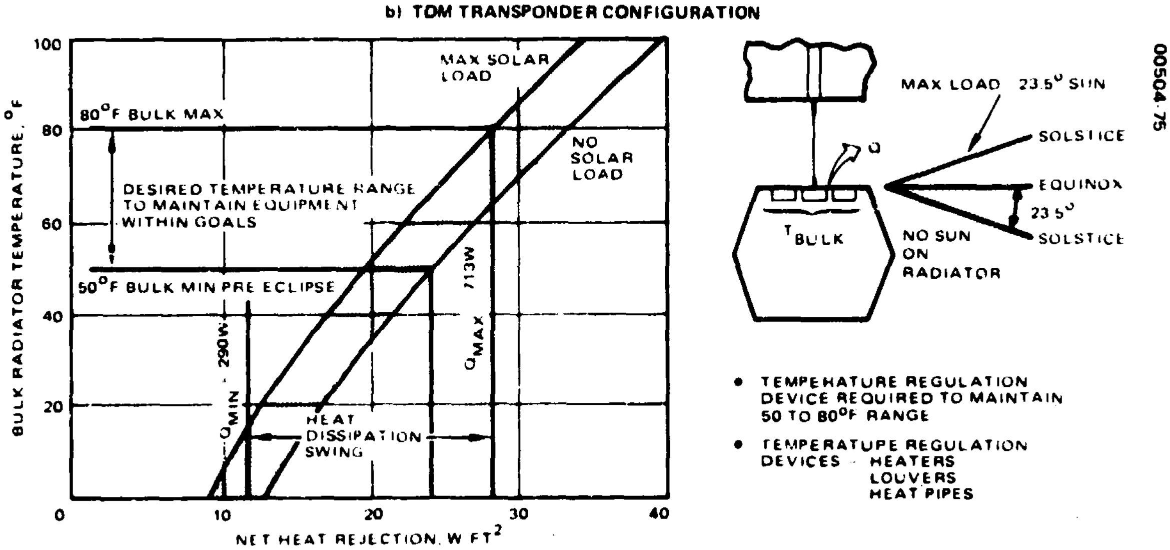
For the three-axis satellite configuration the thermal performance is also generalized by temperature characteristics of the radiator/transponder shelf as a function of heat dissipation density. The FDM transponder configuration thermal perfor-



a) FDM TRANSPONDER CONFIGURATION



b) TDM TRANSPONDER CONFIGURATION



c) TDM/FDM TRANSPONDER CONFIGURATION

FIGURE 3-16. TYPICAL THREE-AXIS SPACECRAFT THERMAL PERFORMANCE

mance is shown in Figure 3-16a. In the spinner configuration, the transponder temperature range is established as shown in Figure 3-16a. Because of the large heat dissipation swing of the transponder from high power to low power, a temperature regulation device is required. The viable temperature regulation devices are heaters, louvers, and heat pipes. The performance profiles for the TDM and hybrid are shown in Figures 3-16b and 3-16c, respectively.

4. FDM REPEATER DESIGN

This section of the report describes the design trades, the baseline design of the FDM repeater, and the performance parameters of the FDM transponder.

4.1 TRANSPONDER DESIGN TRADES

Table 4-1 lists the major constraints which influence the design of the FDM repeater. The traffic matrix shown in Table 4-2 and the modulation format must be considered in determining channel bandwidth and interconnection of channels. The RF power per channel requirement determines the output power level of the power amplifiers used in the repeater. Degradation of BER relative to the theoretical due to repeater linear and nonlinear distortion must be considered in selecting channel bandwidth, design of the individual channels, the number of channels per power amplifier, and the degree of power amplifier backoff.

A dual-mode high power amplifier (HPA) is required to provide 10 dB of link margin for conditions of heavy rainfall attenuation on the downlink. A repeater beacon transmitter is also required by the statement of work and the frequency of this transmitter must be coherent with the repeater local oscillator.

4.1.1 Repeater Channelization Trade

The FDM repeater basic design concept is shown in Figure 4-1. The repeater receives signals from and transmits signals to ten spatially isolated nodes cities. Spatial isolation permits frequency reuse on all nodes. The number of channels provided for each node is a function of the amount of data originating from or terminating at the particular node. Table 4-2 gives a breakdown of the data originating from and terminating at the ten nodes.

The channel capacities suggested in the study statement of work are 137 Mbps, 274 Mbps, and 548 Mbps. The modulation format is specified to be QPSK. The symbol rates are therefore one-half of the data rate. A basic channel bandwidth of 137 MHz is compatible with a data rate up to 200 Mbps, as discussed in 4.3.3.

Two approaches were considered in determining the FDM repeater channelization. One uses three different channel bandwidths, 137 MHz, 274 MHz, and 548 MHz. The other uses only a 137 MHz channel bandwidth. Table 4-3 shows the channel bandwidth allocations for the nonuniform channel bandwidth configuration, and Table 4-4 shows the channel allocations for the single bandwidth ca . In Table 4-3 the channel bandwidth allocated to a particular city to city link is sized to accommodate the

TABLE 4-1. FDM REPEATER SYSTEM CONSTRAINTS

Statement of Work
Traffic matrix
RF power/channel
Modulation format
Repeater noise temperature extrapolation
G/T
Antenna diameter
Repeater BER degradation allocation
Dual mode high power amplifier
Transponder beacon
Auxiliary
NASA-LeRC link budgets

TABLE 4-2. TRAFFIC DISTRIBUTION BY COMMUNITY OF INTEREST BY CITY

Terminating Mbps / Mbps Originating	New York	Chicago	Los Angeles	San Francisco	Washington, DC	Dallas	Houston	Minn/St Paul	Atlanta	Denver	Total Originating
New York		578	574	231	274	196	213	158	156	120	2500
Chicago	578		403	162	145	142	152	128	107	89	1906
Los Angeles	574	403		286	141	166	179	121	107	114	2090
San Francisco	231	162	286		57	65	70	49	42	45	1006
Washington, DC	274	145	140	56		48	53	39	40	29	827
Dallas	196	143	166	65	49		78	41	40	36	811
Houston	213	152	179	70	53	78		44	44	37	870
Minn.St Paul	158	128	121	49	39	41	44		29	28	637
Atlanta	156	107	107	42	40	39	44	29		23	588
Denver	120	88	114	45	29	36	37	28	23		521
Total Terminating	2500	1906	2090	1006	827	811	870	637	588	521	

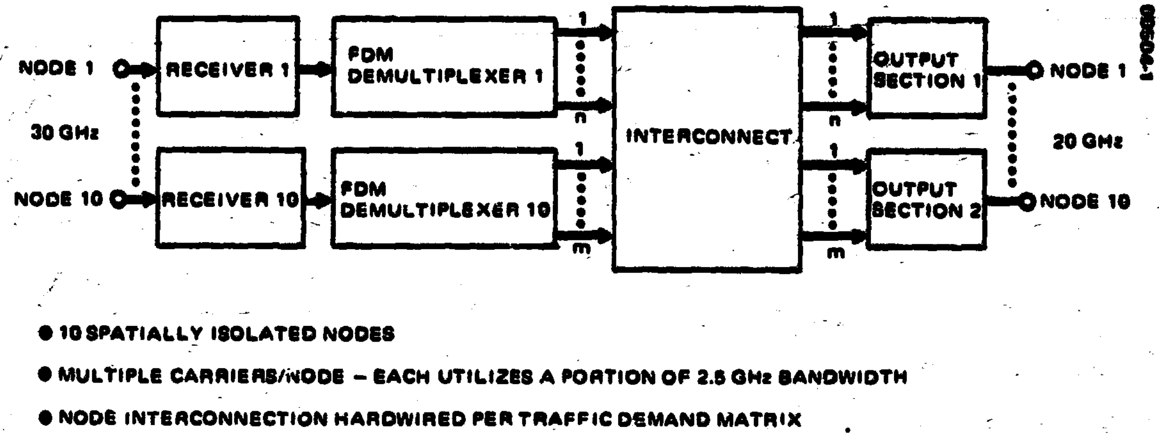


FIGURE 4-1. FDM REPEATER CONCEPT

data rate required for that link. In Table 4-4 the number of 137 MHz channels allocated to a particular link is determined by the data rate. Both configurations provide the same data capacity for each link and both meet the traffic distribution requirements.

Table 4-5 is a comparison of the features of the two configurations. The most significant difference is that the nonuniform configuration requires three different high power amplifier configurations, which significantly increases the cost of this design. Technical risk of this design is also considered to be much greater since high power amplifier design is often a problem area. The uniform channelization configuration has been chosen for the baseline design for these reasons.

Figure 4-2 shows the FDM repeater baseline channel assignment for the uniform channel bandwidth configuration. Note that all cities requiring only nine channels use only the alternate channels. The reason for this will be discussed in Paragraph 4.1.2.

4.1.2 Dual Polarization and Multiplexing Requirements

The design of the FDM repeater is dictated by the data rate requirements of New York, Los Angeles, and Chicago. New York uses 18 channels, each having a 137 MHz bandwidth, resulting in a total bandwidth of 2466 MHz. Obviously very little spectrum is left for a guard band between channels. This poses a severe input spectrum multiplexing requirement, since a channel bandwidth of at least 137 MHz is desired to provide for data rates up to 196 Mbps.

Figure 4-3 illustrates the basic parameters of the demultiplexer design. For contiguous bandpass filters the crossover level is 3 dB. However, contiguous bandpass filters are generally limited to a relatively small number of channels and present formidable design problems for the 18-channel multiplexer required for the New York application.

Two typical multiplexer concepts are shown in Figure 4-4. In Figure 4-4a, energy enters the common waveguide manifold from the left and is coupled through a directional slot into a filter composed of several cascaded circular resonators. Energy within the passband of filter F1 is coupled into that filter. Energy outside the passband

TABLE 4-3. FDM REPEATER NONUNIFORM CHANNEL ALLOCATION
NOMINAL CHANNEL BANDWIDTH, MHz

From \ To	New York	Chicago	Los Angeles	San Francisco	Washington, DC	Dallas	Houston	Minneapolis/St Paul	Atlanta	Denver	No. Filters Rqd
New York		548	548	274	274	137	274	137	137	137	9
Chicago	548		548	137	137	137	137	137	137	137	9
Los Angeles	548	548		274	137	137	137	137	137	137	9
San Francisco	274	137	274		137	137	137	137	137	137	9
Washington, DC	274	137	137	137		137	137	137	137	137	9
Dallas	137	137	137	137	137		137	137	137	137	9
Houston	274	137	137	137	137	137		137	137	137	9
Minneapolis St Paul	137	137	137	137	137	137	137		137	137	9
Atlanta	137	137	137	137	137	137	137	137		137	9
Denver	137	137	137	137	137	137	137	137	137		9
No. Filters Rqd	9	9	9	9	9	9	9	9	9	9	90

TABLE 4-4. FDM REPEATER UNIFORM CHANNEL ALLOCATION
DEDICATED BANDS, 137 MHz

From \ To	New York	Chicago	Los Angeles	San Francisco	Washington, DC	Dallas	Houston	Minneapolis/St Paul	Atlanta	Denver	No. Filters Rqd.
New York		4	4	2	2	1	2	1	1	1	18
Chicago	4		3	1	1	1	1	1	1	1	14
Los Angeles	4	3		2	1	1	1	1	1	1	15
San Francisco	2	1	2		1	1	1	1	1	1	11
Washington, DC	2	1	1	1		1	1	1	1	1	10
Dallas	1	1	1	1	1		1	1	1	1	9
Houston	2	1	1	1	1	1		1	1	1	10
Minneapolis St Paul	1	1	1	1	1	1	1		1	1	9
Atlanta	1	1	1	1	1	1	1	1		1	9
Denver	1	1	1	1	1	1	1	1	1		9
No. Filters Rqd.	18	14	15	11	10	9	10	9	9	9	114

TABLE 4-5. FDM REPEATER CHANNELIZATION COMPARISON

	Uniform Channelization (137 MHz)	Nonuniform Chanelization (137, 274, 548 MHz)
Earth station impact	Requires 114 total uplink and downlink carrier channels, more uniform design	Requires 80 total uplink and downlink carrier channels
Repeater impact		
Multiplexer design		More difficult
Power amplifier implementation	Same design for all nodes	Requires three different power amplifier configurations
Transponder weight and power	5% greater	
Nonrecurring cost		40% greater
Recurring cost		10% greater
Technical risk		Much greater

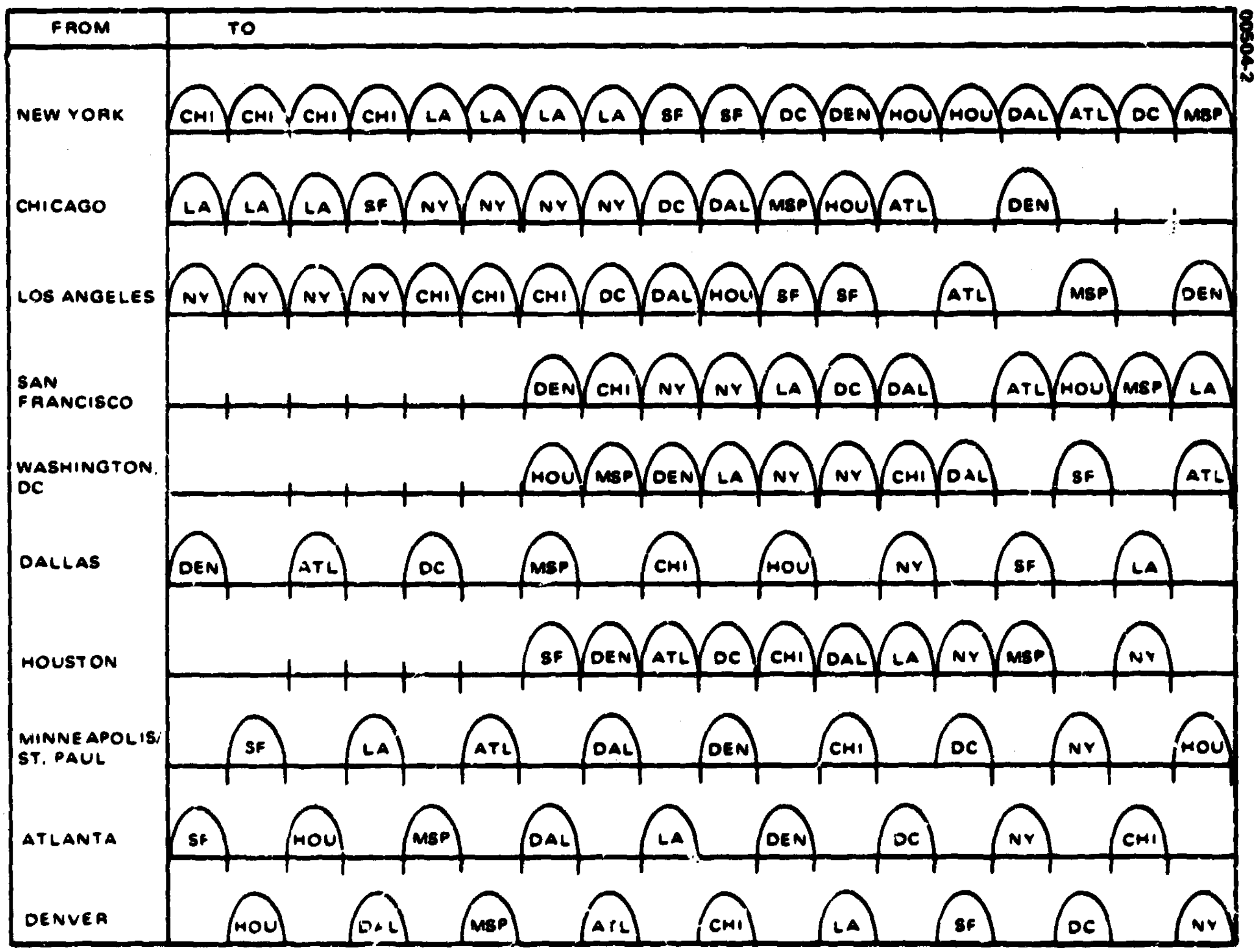
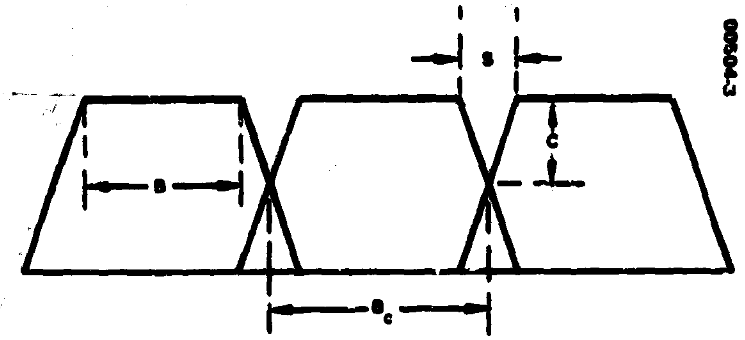


FIGURE 4-2. FDM REPEATER BASELINE CHANNEL ASSIGNMENT



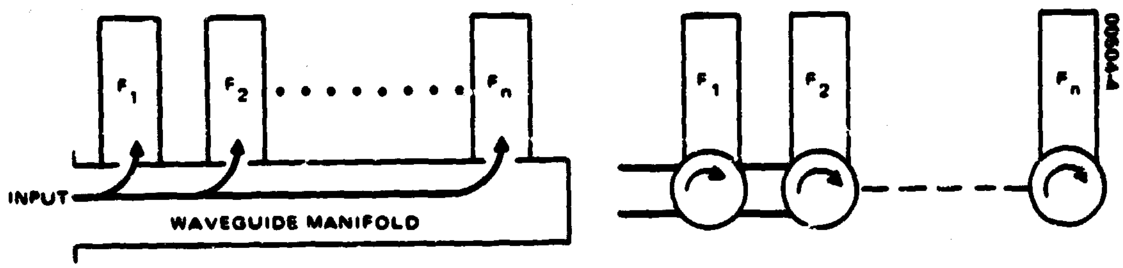
B = CHANNEL FILTER BANDWIDTH

B_c = FILTER BANDWIDTH AT CROSSOVER LEVEL

C = CROSSOVER LEVEL, dB

S = SPACING BETWEEN ADJACENT FILTER PASSBANDS

FIGURE 4-3. MULTIPLEXER CONSIDERATIONS



a) COMMON MANIFOLD

b) CIRCULATOR COUPLED FILTERS

FIGURE 4-4. DEMULTIPLEXER CONCEPTS

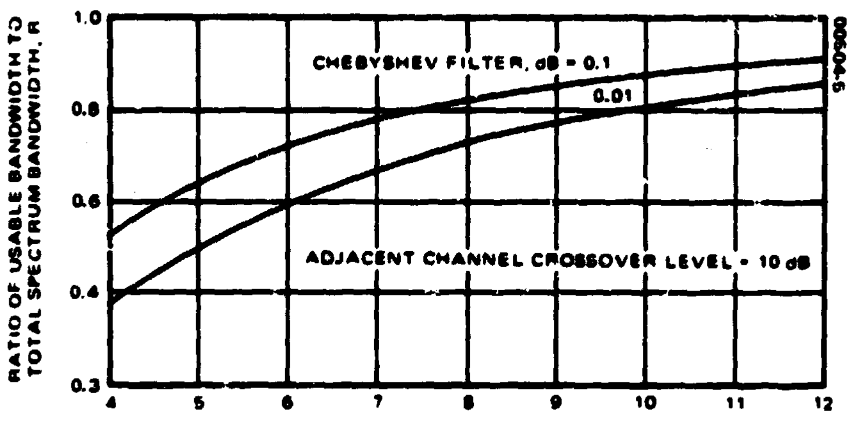


FIGURE 4-5. RATIO OF USABLE BANDWIDTH TO TOTAL SPECTRUM BANDWIDTH

of filter F1 is reflected and continues on to filter F2, and so on. This type of multiplexer is used where channel bandwidth is small, and low, nearly equal insertion loss is desired for all channels. The circulator-coupled filter shown in Figure 4-4b is a simpler design, and is used when low and equal insertion loss in each channel is unimportant. The three-port circulator acts to direct energy within a filter passband into the filter and couples reflected energy outside a filter passband to succeeding filters. Insertion loss of this multiplexer is substantially greater for the nth filter when compared to the first filter, since the energy arriving at the nth filter must propagate through n-1 circulators.

In both multiplexer types, filter F1 affects one side of the passband response of filter F2, and F2 affects one side of the passband response of filter F3, etc. These effects can be compensated fairly easily in design if the crossover level indicated in Figure 4-2 is at least 10 dB.

Figure 4-3 shows a channel spacing, S, between the passbands of adjacent filters. This spacing is a function of the filter bandwidth at the crossover level, B_C, and the filter passband, B. S is related to B_C by

$$S = 2(B_C/B-1)$$

The total bandwidth in an n-channel multiplexer that lies within the passband of all the channel filters is a function of S. The ratio (R) of this bandwidth to the total spectrum bandwidth is a function of S, which is in turn a function of B and B_C. The relation is

$$R = \frac{1}{1+2\,(B_C/B-1)}$$

This relation is plotted in Figure 4-5 as a function of the number of poles of each channel filter for both 0.01 and 0.1 dB ripple Chebyshev filters. These are typical of the filters used in this type of application. The adjacent channel crossover used in deriving Figure 4-5 is 10 dB.

It is clear from inspection of Figure 4-5 that a large number of poles are required to achieve a high ratio of usable bandwidth. For example, a 10-pole 0.1 dB ripple filter provides a total usable bandwidth of 21.975 GHz, or 122 MHz per channel in the case of New York.

The previous discussion illustrates the difficulty and complexity of providing channel filters with nearly contiguous passbands for the high data rate cities. An alternate solution that would greatly simplify the multiplexing problem is to split the uplink and downlink spectrums into odd and even channels, using orthogonal linear or orthogonal circular polarizations to isolate these channels. This approach permits a substantial relaxation in the channel filter requirements and also permits full utilization of channel bandwidth. Polarization isolation typically is at least 20 dB, which reduces adjacent channel interference to a second-order design consideration.

The penalty for using dual polarization is that two receivers are required for each node of the transponder which uses dual polarization. As shown in Figure 4-2, six

TABLE 4-6. COMPARISON OF FDM TRANSPONDER WITH AND WITHOUT DUAL POLARIZATIONS

Parameter	With Dual Polarization	Without Dual Polarization
Channel bandwidth, MHz	> 137	122
Demultiplexer weight, lb	28.5 (4 pole filters)	70 (10-pole filters)
Number of receivers (including redundancy)	32	20
Receiver total weight, lb	64	40
Receiver total power, W	72	45
Polarization component weight, lb	4	
BER distortion degradation	Less	

cities use dual polarization while four do not require it. Twelve receivers must be added for a fully redundant system. The number of channel filters required is not affected by the use of dual polarization. Each node which uses dual polarization must provide an orthomode transducer to generate orthogonal linear polarizations. If orthogonal circular polarizations are required a circular polarizer must be added. These components are a very minor penalty since they are simple and very low-loss. Dual polarization of the FDM repeater output is also necessary in order to preserve the isolation between alternating odd and even channels. This requirement has no significant impact on output circuit design except when a common amplifier is used to amplify all individual FDM channels for any node. It will be shown in 4.1.4 that this particular design is undesirable.

Table 4-6 presents a tradeoff summary, comparing features of the FDM repeater with and without the use of dual polarizations. The system weight is slightly less and the amount of dc power required slightly greater with dual polarization. The channel bandwidth available is substantially improved with dual polarization and BER distortion degradation is less for the dual polarization system because of the larger filter bandwidth and the use of fewer poles in the channel filters. These advantages are sufficient to justify the use of dual polarization.

4.1.3 Input Circuit Trades

The remaining tradeoff to be considered in input circuit design is the frequency at which the input signal spectrum is separated into the individual FDM channels. Three possible design concepts are illustrated in Figure 4-6.

Each of the concepts in Figures 4-6a and 4-6b has a significant penalty. Both require a large number of local oscillator signals, equal to or greater than the total number of 114 individual FDM channels.

The design concept of Figure 4-6a uses single conversion in the repeater, which requires at least two different local oscillator frequencies in order to avoid interference problems due to the second harmonic of the local oscillator frequency. The advantage of this concept is that the local oscillator frequencies are lower and thus require less frequency multiplication.

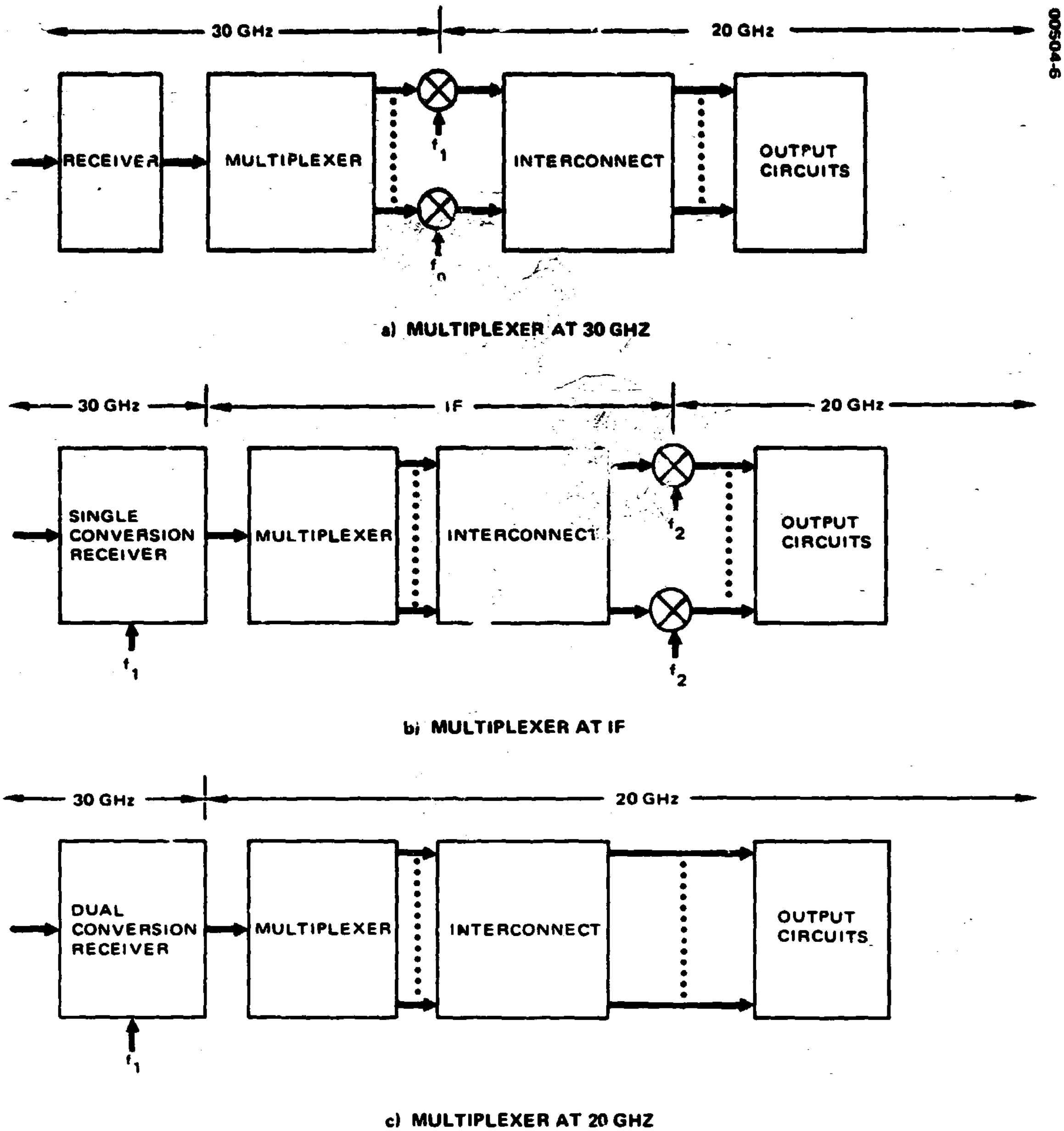


FIGURE 4-6. FDM REPEATER CHANNELIZATION ALTERNATES

The advantage of the concept in Figure 4-6b is that the interconnect is at IF and can be implemented with cables rather than waveguide. However, the multiplexer is at IF and the resonators are much larger compared to the resonators at 20 GHz.

The only disadvantage of the design concept of Figure 4-6c is that the interconnect must be implemented in waveguide. The 20 GHz multiplexers are much smaller and lighter than IF multiplexers and have lower loss than a 30 GHz multiplexer. The design concept of Figure 4-6c has been chosen for the baseline design.

TABLE 4-7. POSSIBLE FDM OUTPUT CIRCUIT CONCEPTS

Concept	Channels/ Amplifier	Power Amplifiers Required (Without Redundancy)	Amplifier Saturated Power, W	Amplifier Backoff, dB		Dual Polarization Output
				Input	Output	
1	18	6	150	12	7	No
		4	75	12	7	
2	9	16	75	12	7	Yes
3	3	40	10	6.5	2.5	Yes
4	1	114	2	0	0	Yes

4.1.4 Output Circuit Design Trades

The major tradeoff issue in output circuit design is to determine the number of channels per output amplifier. The primary comparison criteria are weight and power required. Performance comparison is based principally on BER degradation. This can be made approximately the same for each design compared by using a sufficiently large power amplifier backoff, which does have significant impact on power required. Consequently, performance will not be considered directly in the comparison criteria.

Table 4-7 gives the basic parameters of four possible output circuit designs. The first of these concepts was dropped from further consideration because of the 150 watt saturated power capability required for six nodes, and the fact that more stringent requirements would be imposed on the input circuit channel filters. The remaining three design concepts were compared in more detail.

The amplifier saturated power levels given in Table 4-7 for the various configurations were generated from the statement of work requirement of 1 watt per 137 Mbps channel delivered to the antenna feed. Circuit losses including a 6 foot waveguide run to the antenna were estimated for each configuration. The power required at the power amplifier output was then multiplied by the number of channels, increased by the amount of TWT backoff required, and increased by an additional small loss due to power lost in intermodulation products. The final result is the saturated power capability of the power amplifier.

Figure 4-7 shows the implementation of design concept 4 of Table 4-7 for the New York application. Each channel has a separate solid state power amplifier with a power output of 2 watts. Two nine-channel multiplexers, one for the odd channels and one for the even channels, sum the outputs of the two sets of nine channel amplifiers. The two multiplexers are essentially identical to the input circuit multiplexers for this design. An orthomode transducer sums the output of the two multiplexers to produce orthogonal linear polarizations for the antenna feed. A solid state transmitter was selected for this design concept because of the weight advantage, which more than offsets the slight power efficiency advantage of a low power TWT.

Figure 4-8 shows the implementation of design concept 3 for the New York application. The odd and even channel implementations are identical. Individual

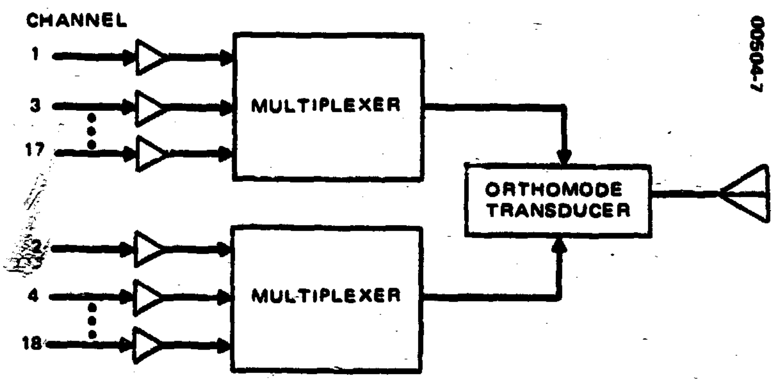


FIGURE 4-7. SINGLE CHANNEL AMPLIFIER FDM OUTPUT CIRCUITS

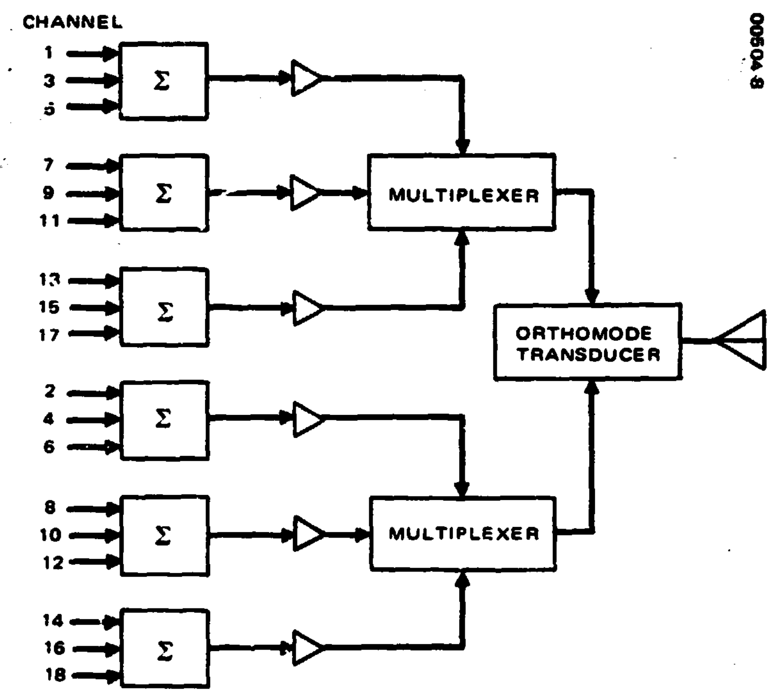
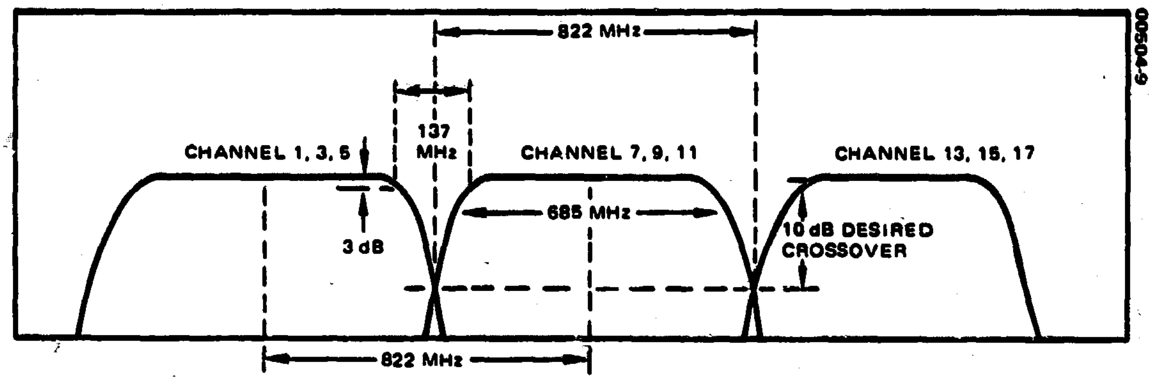


FIGURE 4-8. THREE-CHANNEL AMPLIFIER FDM OUTPUT CIRCUIT CONCEPT

channels are summed in sequential groups of three, amplified in a 10 watt TWT, and then multiplexed in a three-channel multiplexer. A summing circuit rather than a multiplexer is used at the power amplifier input, since channel separation is accomplished in the demultiplexer of the repeater input circuits. The output multiplexers for this design are quite different from the input circuit multiplexer, since each filter must pass the equivalent of five channels instead of one and the guard band between channels is still only one channel bandwidth. Figure 4-9 shows the frequency relations for the output multiplexer. Each channel filter has a passband of equal ripple filter bandwidth of 5 x 137 or 685 MHz. Spacing between the band edges of adjacent filters is 137 MHz. The



RIPPLE, dB	NUMBER OF POLES REQUIRED
0.01	8
0.1	7

FIGURE 4-9. THREE-CHANNEL/AMPLIFIER OUTPUT MULTIPLEXING FREQUENCY DIAGRAM

table in the lower portion of Figure 4-9 indicates the number of poles required for 0.01 dB ripple and 0.1 dB ripple Chebyshev filters. Although a large number of poles are required, the filter insertion loss is still quite reasonable since the percentage bandwidth (3.5 percent) is relatively high and high Q cylindrical resonators are used. Figure 4-10 shows the implementation of design concept 2 for the New York application. The nine odd or even channels are summed in a hybrid network and amplified in a 40 watt TWT. An output multiplexer is not required in this case. A very significant factor in comparing design concepts 2 and 3 is the amount of TWT backoff required to achieve an acceptable BER degradation due to TWT intermodulation products. Figures 4-11 and 4-12 are plots of the results of computer simulations which show BER degradation as a function of TWT backoff. These results were used to select the TWT backoffs given in Table 4-7. Figure 4-13 shows the TWT nonlinearity model used in the computer simulation. The results of Figures 4-11 and 4-12 must be considered on a relative basis since the simulation is for the total system which includes a basic distortion of about 1 dB.

The statement of work constraint on distortion requires that total BER degradation be such that the E_b/N_0 required for 10^{-6} BER at the earth station is 15 dB. This corresponds to a total BER degradation of 4.4 dB. Figure 4-11 shows that the TWT input backoff must be about 10 dB to achieve this amount of degradation for the nine-channel-per-amplifier implementation in the worst case center channel. The three-channel-per-amplifier results of Figure 4-12 indicate that this level is met at an input backoff of about 4 dB. At the selected input backoff levels of 12 dB and 6.5 dB for the nine-channel and three-channel designs, the degradation for the two worst case center channels is 3.4 and 3.2 dB, respectively.

Table 4-8 compares the power and weight of the three candidate output circuit concepts. The third column is an estimate of the total weight impact on the spacecraft and includes the weight of the solar array and power conditioning equipment to provide the power required by the FDM repeater. The power required for the three-channel-per-amplifier concept is much lower than for the other two concepts and weight is somewhat less. This concept has been chosen for the baseline design of the FDM system.

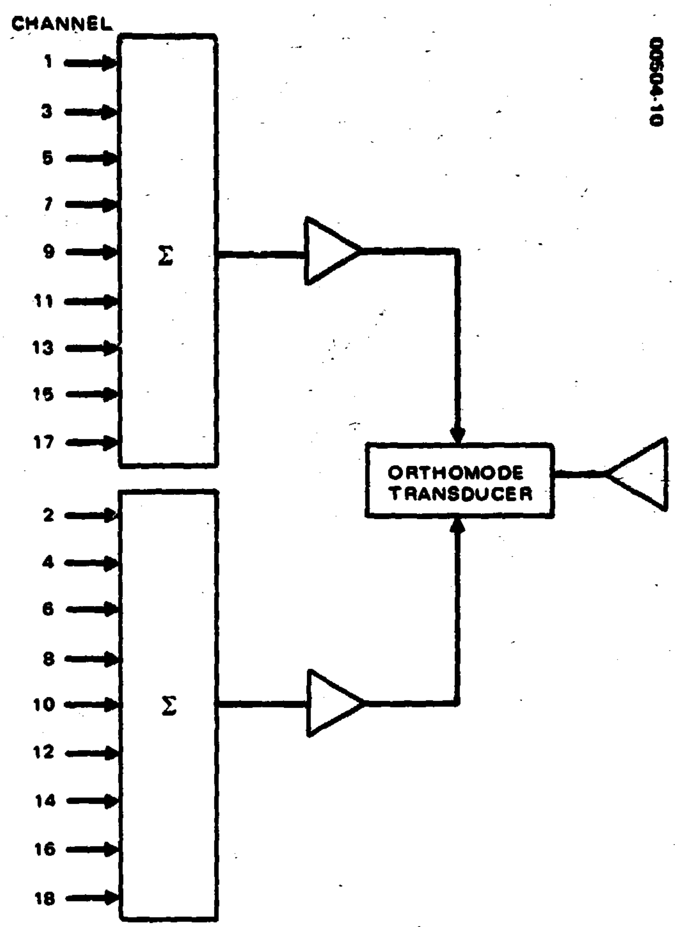


FIGURE 4-10. NINE-CHANNEL/AMPLIFIER FDM OUTPUT CIRCUIT CONCEPT

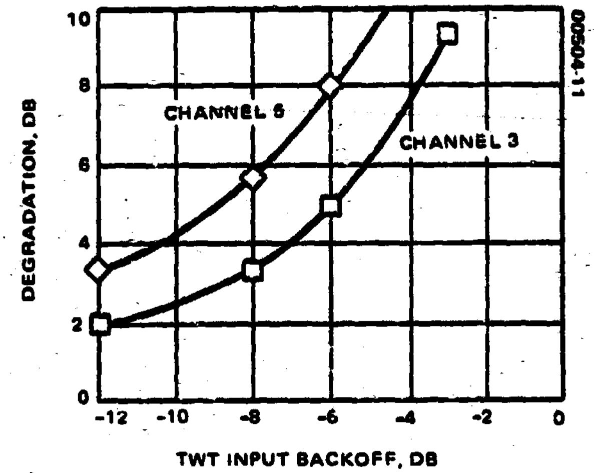


FIGURE 4-11. DISTORTION DEGRADATION, NINE CHANNELS

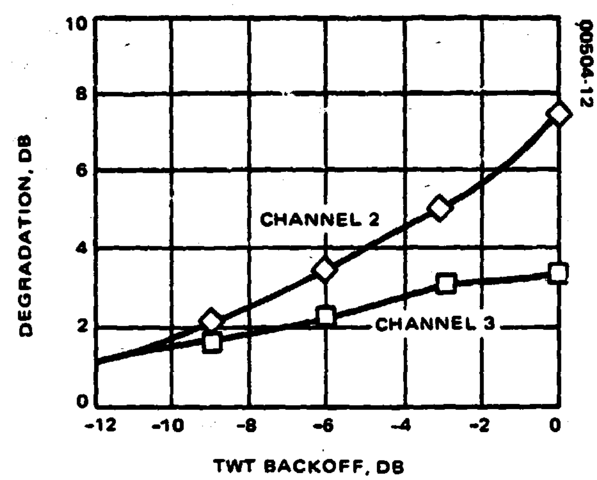


FIGURE 4-12. DISTORTION DEGRADATION, THREE CHANNELS

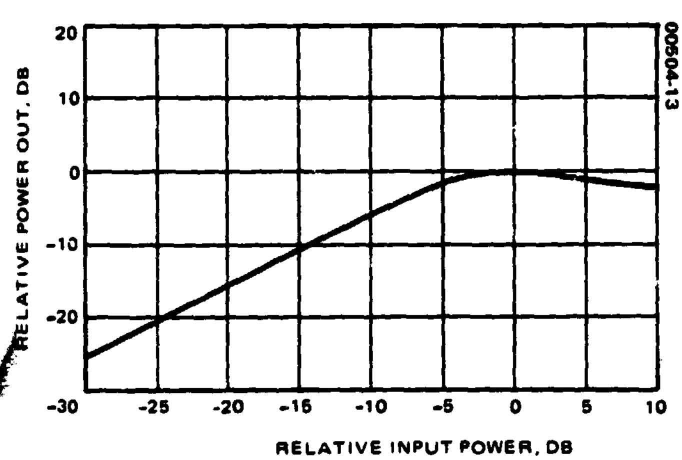


FIGURE 4-13. TWT POWER NONLINEARITY CHARACTERISTIC

TABLE 4-8. POWER/WEIGHT COMPARISON OF FDM CONCEPTS

Concept	Transponder Weight, lb	Transponder Power, W*	Spacecraft Weight Impact**, lb
9 channels/amplifier	620	1502	1221
3 channels/amplifier	587	1204	1062
Single channel amplifier	686	1600	1326

*60 channels high power, 54 low power
**Includes spacecraft solar array and power conditioning weight, 0.4 lb/W

4.1.5 Beacon Transmitter Design Considerations

The statement of work requires that a beacon transmitter be provided. Power level required is +20 dBm at the antenna feed and global coverage is required. It is further required that the carrier frequency be coherent with the repeater local oscillator frequency. No modulation is required on the beacon carrier.

Because the downlink spectrum is fully occupied in the New York case, the beacon must be located somewhere in the modulated data spectrum. Channel assignments for other cities are selected to reduce interference, but it is not possible to eliminate the interference entirely. Two interference problems must be considered in implementation of the beacon. The first and most important is the effect of the modulated signals on beacon reception at the earth stations. The second is the effect of the beacon carrier on data reception at the earth stations. The latter is not expected to be significant.

It is helpful to select a tentative beacon frequency before the interference problems are analyzed. A convenient beacon frequency selection is 10 times the frequency of the local oscillator supplied to the FDM receiver at 1.96 GHz. This frequency is 19.6 GHz and the two closest FDM channels nominally would be at 20.2 GHz - 5 x 137 MHz + 68.5 MHz or 19.5835 GHz and at 20.2 GHz -4 x 137 MHz + 68.5 MHz or 19.7205 GHz. However, some latitude is permitted in selecting channel frequencies since 2.5 GHz is available and 18 x 137 MHz is 34 MHz less than the total available. Therefore these channels will be located at 19.550 and 19.687 GHz to increase the offset from the 19.6 GHz beacon frequency.

The effective radiated power (ERP) of the beacon carrier is the required +20 dBm of the transmitter plus the beacon antenna gain. The beacon antenna provides global coverage which is consistent with a peak gain of about 19 dBi. Assuming that the beacon antenna pattern is centered on the equator, the effective gain over the earth station region will be about 18 dBi. ERP in the direction of the earth stations is then +38 dBm.

The gain of the 14 foot spacecraft antenna is about 56 dBi and the power level for each channel is 1 watt. The ERP for the data signal in the direction of the earth station is +86 dBm, which is 48 dB above the beacon ERP. The data channels are always modulated, however, and it is the power of the spectrum of the modulated signal in the vicinity of the beacon frequency that is of concern. Table 4-9 was generated from the data given in Figure 4-2 and Table 4-2 and shows the various links and data rates by city

TABLE 4-9. CITY TO CITY LINKS AND DATA RATES ADJACENT TO BEACON FREQUENCY

Originating City	Destination City	Carrier Frequency, GHz	Data Rate, Mbps	Power Density at 19.6 GHz Relative to Carrier, dBc/Hz	Discrete Spectral Line Power at 19.6 GHz Relative to Carrier, dBc
New York	Houston	19.550	111.5	-96.4	-64.1
New York	Dallas	19.687	196	-98.1	-63.3
Chicago	Denver	19.687	89	-109.3	-78.0
Los Angeles	Atlanta	19.550	107	-100.4	-68.3
San Francisco	Atlanta	19.687	42	-102.8	-74.7
Washington, DC	Dallas	19.550	48	-101.9	-73.2
Dallas	San Francisco	19.687	65	-95.0	-65.1
Houston	New York	19.550	111.5	-96.4	-64.1
Houston	Minneapolis/ St Paul	19.687	44	-112.2	-84.0
Minneapolis/St Paul	Washington, DC	19.550	39	-91.2	-63.5
Atlanta	New York	19.687	156	-98.8	-65.1
Denver	San Francisco	19.550	45	-94.2	-65.9

for the channels at 19.555 and 19.687 GHz. Table 4-9 also shows the power density and discrete spectral line power at 19.6 GHz relative to total data carrier power. Derivation of the values of these two columns will now be shown.

The power density relative to total carrier power of a carrier QPSK modulated by an NRZ random data bit sequence of infinite length is given by

$$\text{Power density} = \frac{P_c}{\text{Symbol rate}} \left(\frac{\text{Sin}\,\pi f T}{\pi f T} \right)^2 \tag{1}$$

where P_c is carrier power, f is frequency offset from the carrier, and T is symbol period. This equation is the basis for the power density column of Table 4-9.

Equation 1 represents the envelope of power density as a function of symbol rate and frequency, f, offset from the carrier. For an infinite sequence of truly random data bits it represents power density in watts/Hz. For a pseudorandom data sequence of length n bits which is transmitted repetitively, discrete spectral lines appear which have a frequency separation, f, given by

$$\Delta f = \frac{\text{Symbol rate}}{n} \tag{2}$$

The power of each spectral line relative to total carrier power at any frequency, f, from the carrier can be derived from the consideration that each spectral line contains the total power for a frequency region of Δf Hertz. For this particular case, the power in any spectral line, P_r, at a frequency f from the carrier which is integrally divisible by f is derived by combining Equations 1 and 2 to give

$$P_r = \frac{P_c}{n} \left(\frac{\sin \pi f T}{\pi f T}\right)^2 \tag{3}$$

where n is the number of bits in the sequence.

Digital data is usually formatted in a finite bit length frame with a certain number of bits repeated in each frame to identify frame occurrence. Common practice is to divide a frame into a number of words, each having the same number of bits, with one word for frame synchronization. The remaining words are used for data and can be considered essentially random. This format is similar to the previously described pseudorandom repetitive bit sequence. The difference is that the repetitive sequence is now one word in m data words and the remaining words are random data bits. Therefore, Equation 3 is modified by dividing by the number of words per frame, m, to give the power of spectral lines caused by the repetitive frame synchronization word. The resulting expression for the power in any spectral line, P_s, relative to carrier power is

$$\frac{P_s}{P_c} = \frac{1}{mn} \left(\frac{\sin \pi f T}{\pi f T}\right)^2 \tag{4}$$

Equation 4 is the basis for the last column of Table 4-9, which gives the power of the nearest discrete spectral line relative to total data carrier power. Interference levels will be evaluated based on a conservative assumption of 512 data bits per frame and a synchronization word length of 8 bits and 63 data words per frame. It was shown previously that the data ERP relative to beacon ERP is +48 dB. Referring back to Table 4-9, the worstcase spectral line interference occurs for the New York to Dallas and Minneapolis/St. Paul to Washington, D.C. channels and is +48 -63.5, or 15.5 dB below the beacon carrier.

It is assumed that the primary purpose of the repeater beacon is to monitor the frequency and quality of the repeater local oscillator. A simple second-order phase lock receiver at the earth station can recover the beacon carrier even in the presence of discrete interference if the signal to interference ratio is ≥ 10 dB. However, the recovered signal may be phase-modulated at a low index at a rate equal to the frequency separation of the interference and the beacon frequency. This modulation will be very low and of no consequence if the interfering spectral line lies outside of the tracking loop response of the earth station phase lock loop. Assuming that the earth station phase lock receiver single-sided loop noise bandwidth is on the order of 1000 Hz and has a damping ratio of 0.7, the phase lock loop response is 10 dB down at a frequency 1.4 kHz away from the beacon carrier frequency. In Equation 2 the frequency spacing of the interfering spectral lines ranges from 63 kHz to 109 kHz for all cases where signal to interference ratio is less than 20 dB. If interference should become a problem, a small change in ground transmitter frequency can be used to avoid the problem. For example, a shift of 1 part in 10^6 will shift the frequency of the interfering line by 19.6 kHz.

The purpose of the foregoing discussion has been to establish that the use of the beacon is not impaired substantially by data interference for the particular beacon frequency selected. This has been shown to be the case, with the possible reservation that minor adjustment of earth station data carrier frequencies might occasionally be necessary. The ratio of data carrier power to beacon carrier power is 45 dB, sufficiently large that the possibility of beacon interference with data recovery is negligible.

4.2 FDM REPEATER DESIGN

This section describes the final baseline design of the FDM repeater based on the design tradeoffs of 4.1.

4.2.1 Repeater Input Circuits

Figure 4-14 is a simplified block diagram of the FDM repeater baseline design input circuits. Six of the nodes which receive more than nine channels have a circular polarizer and an orthomode transducer which separates the received dual polarization signal into a set of even channels and a set of odd channels. Separate dual conversion receivers are required to amplify and frequency convert the odd and even channel signals to the output frequency of 17.7 to 20.2 GHz. Four of the nodes, as shown in the lower portion of Figure 4-14, receive a single polarized signal and thus require only one operational receiver. The outputs of each receiver are separated into individual channels in an odd or even channel multiplexer. The multiplexer consists of up to nine-channel filters and phase equalizers isolated from the signal flow by a three-port circulator at the input of each channel filter.

Redundancy is not shown in Figure 4-14, but each receiver shown on the diagram is actually a pair of redundant receivers whose inputs and outputs can be selected by latching circulator switches. Insertion loss of the latching circulator input switch is 0.2 dB. Each latching switch is driven by an internally redundant driver. The channel separator components are not redundant since they are entirely passive.

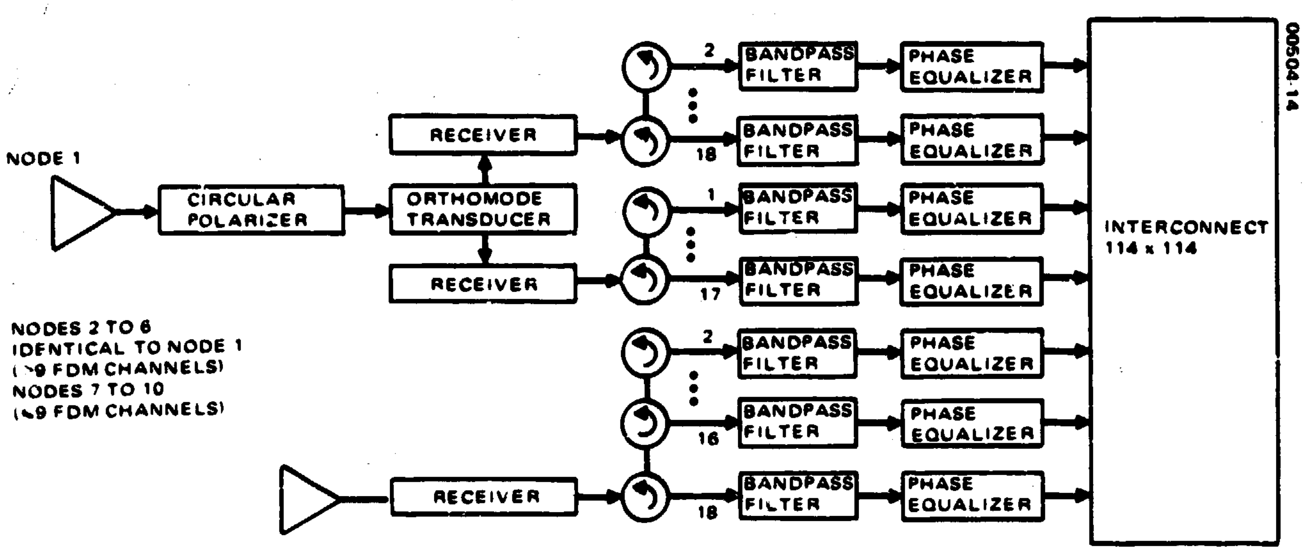


FIGURE 4-14. FDM BASELINE INPUT CIRCUITS

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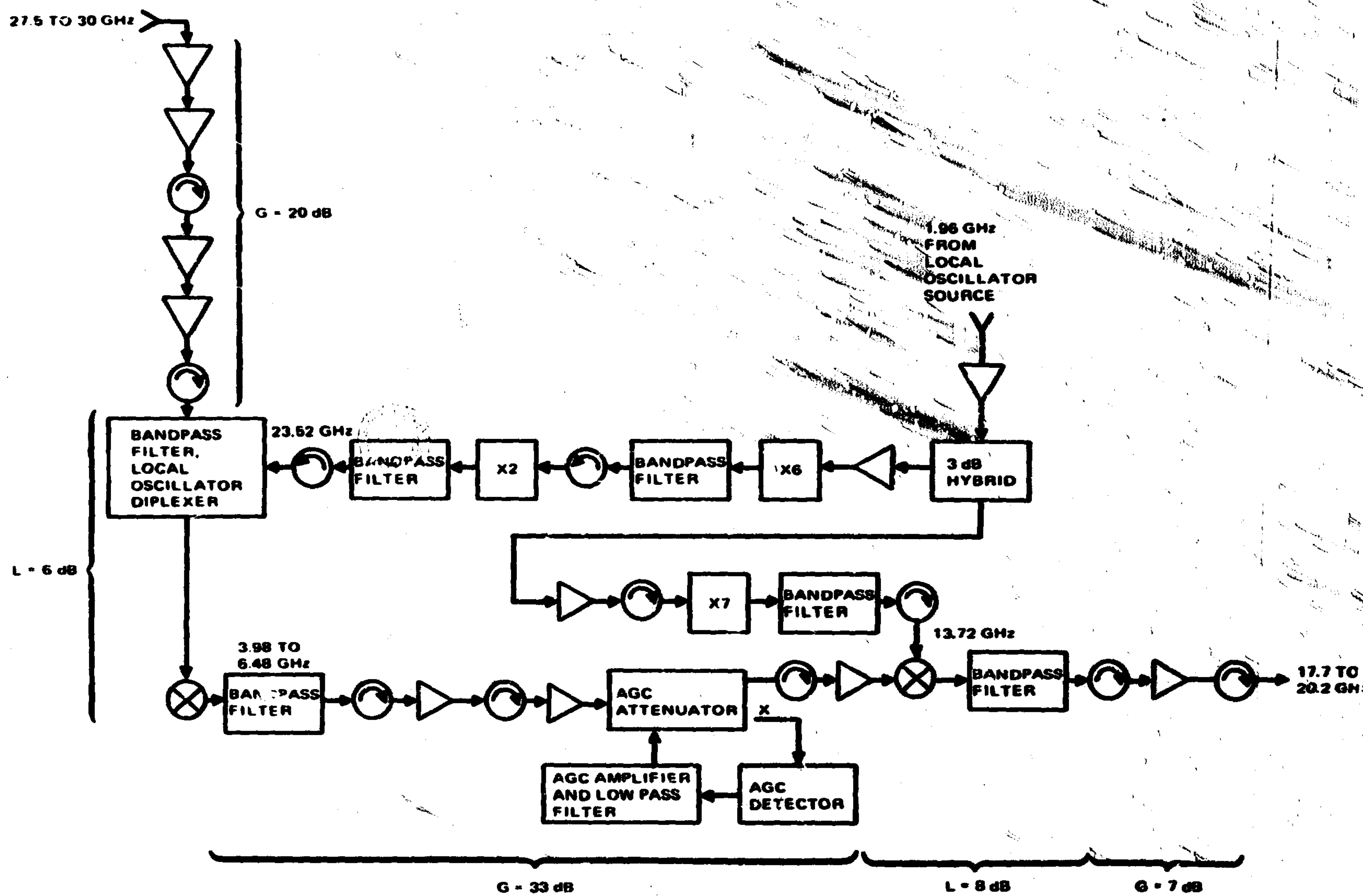


FIGURE 4-15. FDM REPEATER DUAL CONVERSION RECEIVER

4.2.1.1 Receiver Design

Figure 4-15 is a detailed block diagram of the FDM repeater dual conversion receiver. The preamplifier, first mixer, and first local oscillator chain were described in detail in 3.3 of this report. A four-stage low noise GaAs FET preamplifier amplifies the input signal and establishes the receiver noise figure. The signal is then combined with the 23.52 GHz first local oscillator in a combination BPF and local oscillator diplexer, and downconverted to IF, 3.98 to 6.48 GHz. The bandpass filter has a 2.5 GHz signal bandwidth. The first mixer is a single-ended waveguide-mounted diode.

The IF amplifier consists of three stages of wideband GaAs FET amplifiers constructed in microwave integrated circuit (MIC) form. An AGC loop is also provided to maintain a constant signal level at the receiver output. The signal is then upconverted in a balanced MIC mixer, filtered in a bandpass filter, and amplified in a single stage output amplifier.

The first and second local oscillator frequencies are generated by frequency-multiplying the 1.96 GHz reference frequency supplied by the local oscillator source described in 3.5. The first local oscillator multiplier chain has two frequency multipliers, each of which is followed by a bandpass filter to reject unwanted harmonies. Spurious outputs of the multipliers are at least 80 dB below the desired frequency. The second local oscillator multiplier is very similar in design and construction to the first local oscillator multiplier discussed in 3.3, except that only one frequency multiplier, a x 7 multiplier, is required. The most important receiver performance parameters are noise figure, gain variation over the passband, gain slope, and spurious responses. Sufficient gain is provided in the four-stage low noise preamplifier to ensure that post-preamplifier noise contribution is negligible.

All signal amplifiers in the receiver are designed with the input and output matching networks optimized in the upper frequency region of the desired bandpass. This technique compensates for the inherent gain decrease with increasing frequency of the GaAs FET amplifiers. Isolators are used extensively to minimize interstage VSWR, which can cause ripple in the bandpass response. Gain variation will be ± 1 dB over the 2.5 GHz passband, and gain variation over any 137 MHz channel bandwidth segment will be less than 1 dB peak to peak.

Spurious responses are minimized by selection of the first and second local oscillator frequencies and provision of adequate filtering in the local oscillator multiplier chain. Gain distribution in the receiver is designed to avoid excessive signal level at the input to the second mixer. Excessive signal level at this point would cause significant intermodulation products and also generate a spurious output due to the third harmonic of the IF and the spurious product created by twice the second L.O. frequency minus the signal second harmonic.

Receiver size, weight, and power are tabulated below:

Size	6 x 10 x 2 in.
Weight	3.0 lb
DC power	4.5 watts

4.2.1.2 Input Multiplexer

The input multiplexer design parameters are as follows:

Channel bandwidth	137 MHz
Channel spacing	274 MHz
Filter type	0.1 dB ripple Chebyshev
Number of poles	4

Each input multiplexer consists of a series of up to nine individual filters connected in a channel-dropping arrangement as illustrated previously in Figure 4-4b. Energy within the passband of each filter is coupled to the output of that filter, and energy outside the passband is reflected and coupled by the circulator at the filter input to the adjacent filter. This setup is duplicated for each filter down to the lowest frequency filter. The filters are arranged in order of decreasing frequency because receiver gain generally will decrease with increasing frequency and output power amplifiers will tend to have lower gain at higher frequency and finally the channel filter insertion loss tends to increase with frequency.

The circulator coupled filter arrangement was chosen for the input multiplexer in preference to the common waveguide manifold approach shown in Figure 4-4a because of the potential filter interaction problems inherent in that design.

Each channel filter is a dual mode directional circulator resonator filter and uses orthogonal TE_{111} modes. Thus only two resonators are required for a four-pole filter design since each resonator is used twice. Insertion loss per filter is approximately 0.4 dB, and the loss associated with each circulator reflection is 0.2 dB. Total insertion loss of the multiplexer ranges from 0.6 dB for the highest frequency channel to 2.2 dB for the lowest frequency in a nine-channel multiplexer.

Each filter of the multiplexer may be followed by a phase or time-delay equalizer to reduce the filter's phase distortion. The equalizer is required only for channels where the data rate is 137 Mbps or greater. Lower data rate channels do not require the equalizer. Each equalizer consists of a circulator and a resonator connected to one port of the circulator. Energy reflects with equal loss over the channel filter passband so that amplitude response is not affected, but the group delay of the reflected energy varies with frequency in such a way that the channel filter group delay variation over the passband is reduced.

All outputs of a multiplexer include a selectable attenuator pad, which is used to equalize the levels of each group of multiplexer outputs to a common, standardized level. These pads compensate for receiver passband gain variation, multiplexer filter loss variation and repeater interconnect circuit loss variation, and ensure that the level of each signal at the input to the summing circuit preceding the TWT is equal.

The theoretical response characteristics of two typical adjacent filters of a multiplexer are shown in Figure 4-16. The theoretical attenuation characteristics have been generated by transforming a low pass prototype response into the response of a narrow band waveguide filter which includes the guide wavelength variation effects of the circular resonator filter.

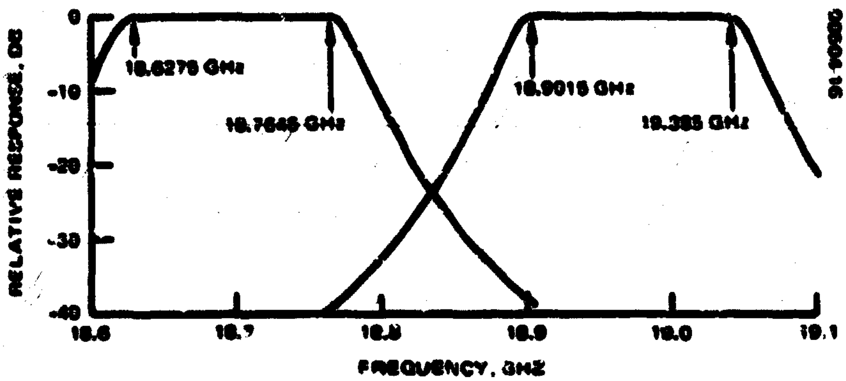


FIGURE 4-16. INPUT MULTIPLEXER FILTER RESPONSE

It is evident from inspection of Figure 4-16 that a four-pole channel filter is more than adequate for separation of alternate odd and even channels. In addition, the attenuation at the band center of the orthogonally polarized channel is more than 20 dB. This attenuation, in combination with a polarization isolation of 20 dB, reduces adjacent channel interference in the FDM repeater to a negligible level.

The multiplexer filters are made of Invar for good temperature stability. Size and weight of a channel filter including the input isolator is as follows:

Envelope size	1.1 x 2 x 1.5 in.
Weight	0.30 lb

4.2.2 Output Circuit Design

Figure 4-17 is a block diagram showing the configuration of the FDM repeater baseline design output circuits. Each group of three channels is summed in a summing circuit consisting of three WR-42 short slot hybrids. A switched attenuator controls the RF power level into the driver amplifier and TWT as required by the dual mode operation of the TWT. Crossover waveguide switches are used at the input and output of each group of attenuators, drivers, and TWTs for redundancy switching. Three for two redundancy is employed in the output circuits, and the points labeled "A" in Figure 4-16 are connected to the next group of output circuits so that each operational channel always has two adjacent spare amplifier groups to select from when required.

A power monitor is provided for status monitoring after the output crossover switch in each group. Groups of three channels each are then combined in the output multiplexer. A harmonic or low pass filter is used to attenuate the TWT harmonics and to attenuate TWT noise in the receiver band. The odd and even channels are then combined in an orthomode transducer and converted to orthogonal circular polarizations in the circular polarizer. If orthogonal linear polarizations are desired the circular polarizer may be omitted. It should be noted that the polarizer components must be located at the antenna feed and not in the repeater package. Waveguide and waveguide components use WR-42 instead of WR-51 for all components preceding the TWT because of the smaller size and lighter weight. All components except the output multiplexer following the TWT are in WR-51 waveguide to reduce insertion loss.

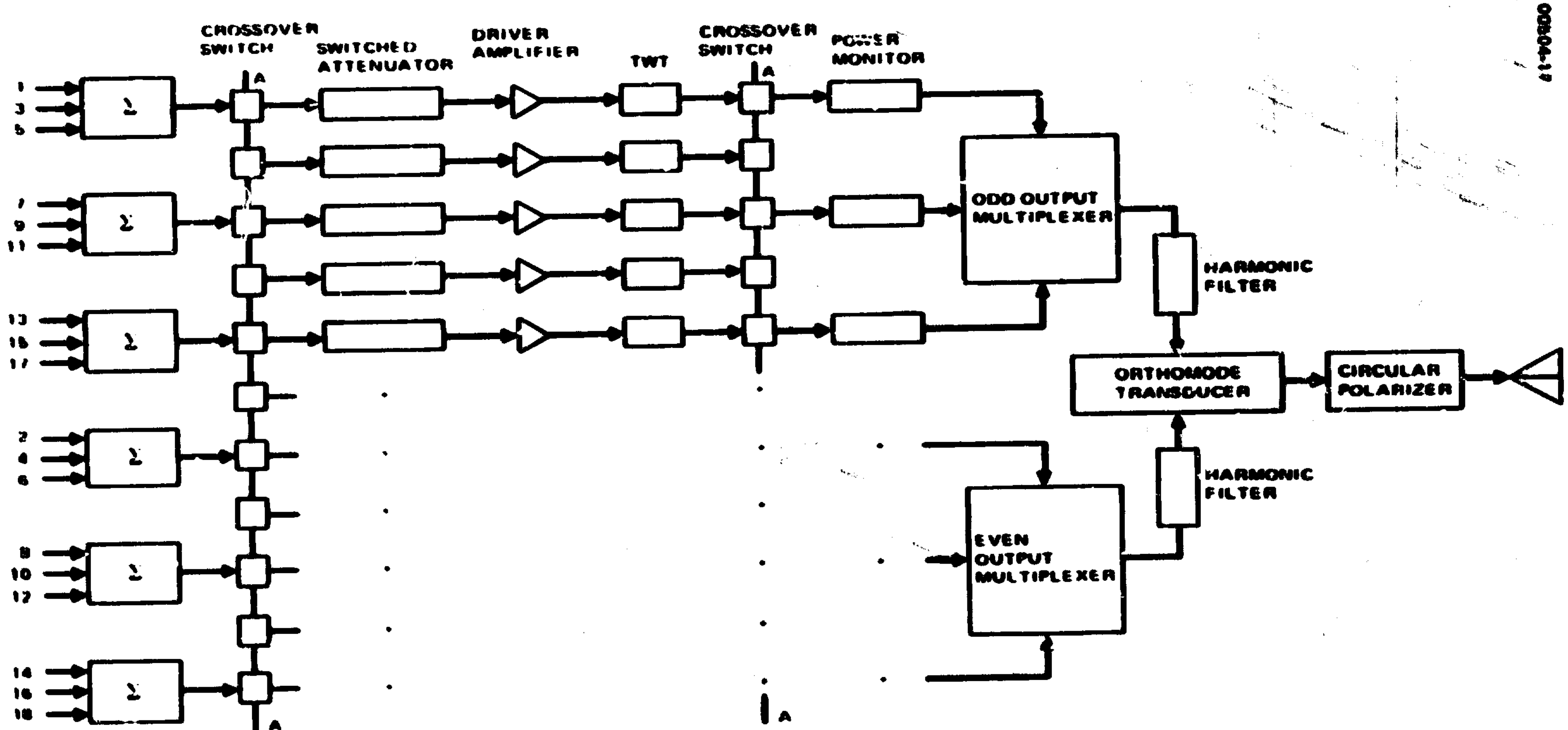


FIGURE 4-17. FDM BASELINE OUTPUT CIRCUITS

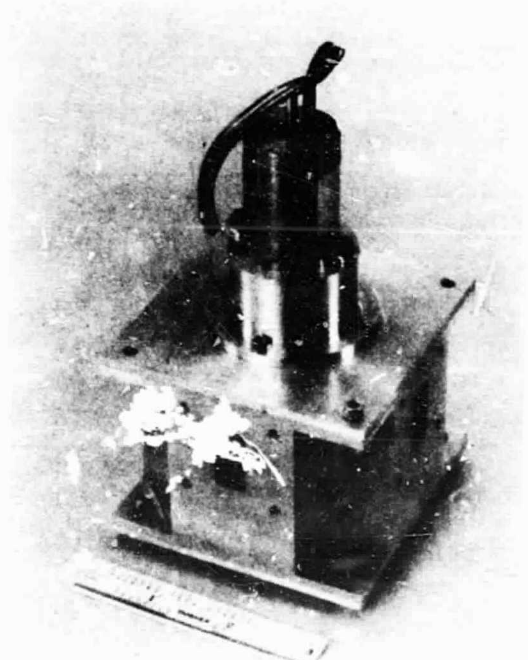
00804-17

4.2.2.1 Crossover Switch

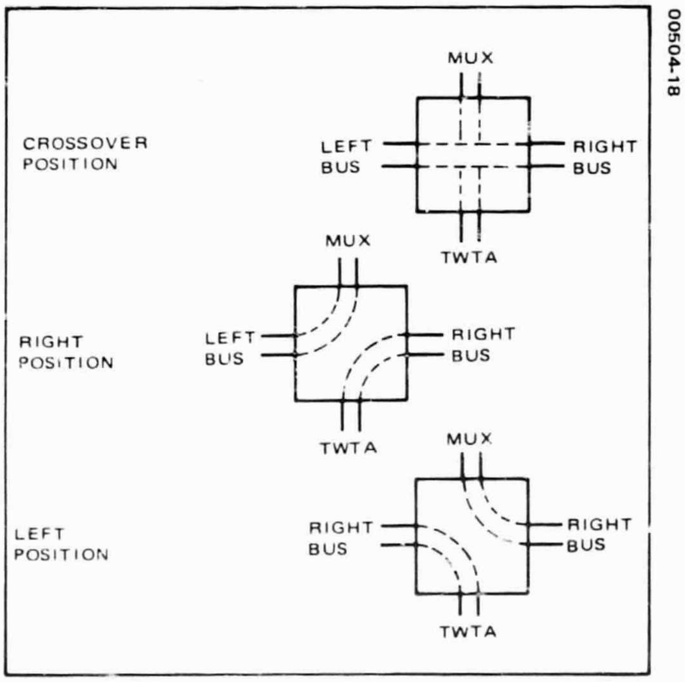
The crossover switch is a WR-51 waveguide, latching type, with a motor and reduction gear drive. A photograph of a crossover switch designed by Hughes is shown in Figure 4-18a. The operation of the crossover switch is illustrated in Figure 4-18b for the case where it is used at the TWT output. In normal operation the switch is in the crossover position, in which the TWT is connected directly to the output multiplexer. This position also connects the left bus to the right bus, which permits additional flexibility in selection of a spare TWT in the event of failure. The left bus and right bus positions connect either the spare TWT on the left or the spare TWT on the right to the output multiplexer.

The flexibility of the crossover switch is best illustrated by an example. Figure 4-19 shows such an example. Amplifier chains 2 and 4 are normally operational and chains 1, 3, and 5 are spares. Amplifier 2 has failed, and amplifier 1 is already being used as a spare and is not available. Amplifier 3 is discovered to be nonoperational when it is substituted. Switches 2, 3, and 5 are then switched to the position shown in Figure 4-18 so that amplifier chain 5 can be used in place of amplifier chain 2 without disturbing the normal operation of amplifier chain 4.

The weight of each crossover switch and driver is estimated to be 0.8 lb.



a) CROSSOVER SWITCH
(PHOTO 77-55315)



b) POSITIONS

FIGURE 4-18. CROSSOVER SWITCH

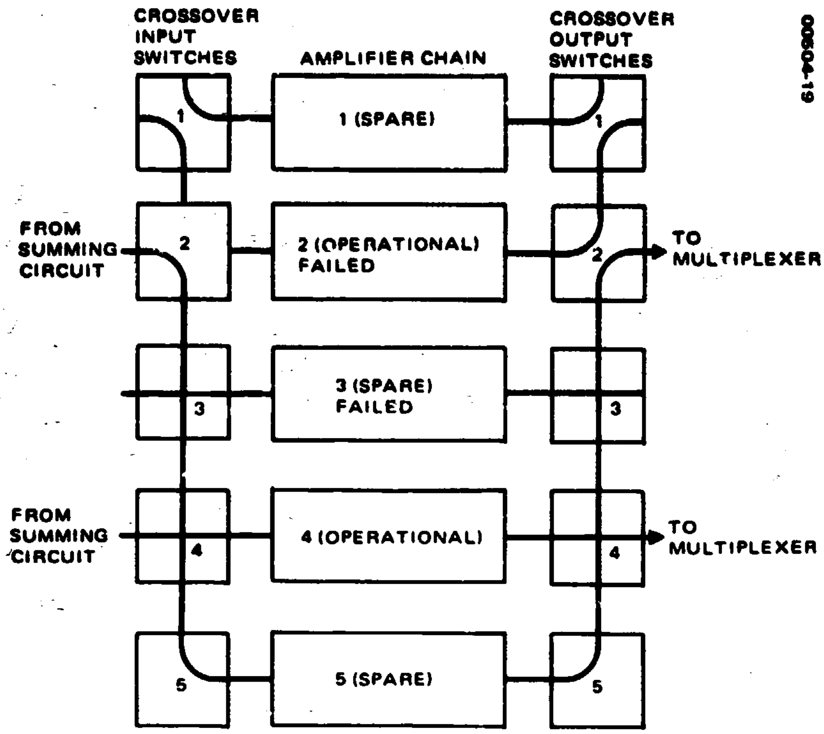


FIGURE 4-19. EXAMPLE OF CROSSOVER SWITCH OPERATIONAL USE

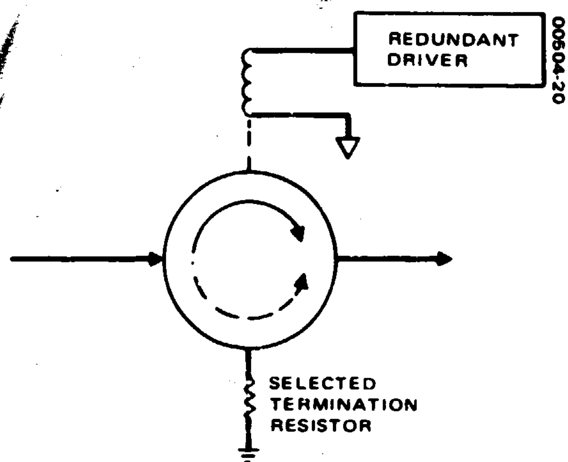


FIGURE 4-20. CIRCULATOR TYPE SWITCHED ATTENUATOR

4.2.2.2 Switched Attenuator

The switched attenuator consists of a fixed attenuator and a switched attenuator. The values of the fixed pad and the switched attenuation are selected to match the equal standardized levels at the summing circuit inputs to the dual mode high power amplifier gain characteristics in both the high power and low power modes.

The fixed attenuator consists of a short section of waveguide with a resistance card placed normal to the E field. The switched attenuator is a latching three-port waveguide circulator. Figure 4-20 is a diagram which shows the operation of the switched attenuator. When the energy flow is in the direction indicated by the solid line the attenuator is in the minimum attenuation position. When the circulator energy flow is as indicated by the dashed line the energy is partially absorbed by the selected termination resistor. The degree of mismatch of the terminating resistor determines the amount of attenuation in this mode. This design has been developed for high reliability applications. The latching circulator is essentially a passive component, and the only active components are in the driver circuit, which is redundant.

4.2.2.3 Driver Amplifier

The driver amplifier is a two-stage GaAs FET amplifier and has 13 dB power gain. The amplifier is designed as a microwave integrated circuit and is mounted on a substrate, which is in turn mounted in WR-42 waveguide. The construction is nearly identical to that used in the low noise preamplifier which is shown in Figure 3-3. A circulator is used at the amplifier input and output to ensure low input and output VSWR.

The bandpass required of the driver amplifier in this application is only 685 MHz. Three designs which differ only in the input and output tuning and impedance matching networks are used. Each amplifier has a bandpass response of approximately 1 GHz. The use of three different lower bandwidth designs as opposed to a common broadband design simplifies the task of aligning the amplifier for the desired passband response.

4.2.2.4 TWTA

The 10 watt TWT is required to operate at two power levels. The low power mode is used for normal operation and a high power mode wherein output power is increased by 10 dB. The high power mode is used to overcome atmospheric attenuation on the downlink caused by heavy rainfall. The 10 watt TWT is operated with the output backed off by 2.5 dB to reduce BER degradation due to intermodulation of the three FDM channels. Power output in the high power mode is 5.6 watts and 0.56 watts in the low power mode. Table 4-10 lists the key performance parameters of the TWT.

The helix type TWT uses dual collectors for improved efficiency. Figure 4-21 is a schematic of the TWT showing the arrangement of electrodes. Table 4-11 lists the basic parameters of the TWT for the high power and low power modes. Table 4-12 lists the electrode voltages and currents for the TWT.

TABLE 4-10. 10 WATT TWT KEY PERFORMANCE PARAMETERS (FDM)

Parameter	Value
Frequency	17.7 to 20.2 GHz
Output power	
Low mode	0.56 W
High mode	5.6 W
Saturated gain	
Low mode	40 dB
High mode	55 dB
Efficiency (two-stage collector)	
Low mode	20%
High mode	30%

TABLE 4-11. 10 WATT TWT OPERATING PARAMETERS

Parameter	Low Mode	High Mode
Beam voltage, V	-4000	-4400
Anode voltage, V	-800	0
Beam current, mA	6.0	23
DC power, W	5.0	33.0

TABLE 4-12. 10 WATT TWT ELECTRODE VOLTAGES AND CURRENTS

Voltages		Currents
E_K	≥ -4000 to -4400 V	$I_K \geq$ 6 to 23 mA
E_{A1}	≥ -800 to 0 V	$I_{A1} <$ 0.1 mA
E_{A2}	≥ +100 V fixed	$I_{A2} <$ 0.1 mA
E_F	≥ 6.0 V	
I_F	≥ 0.7 amp	
E_{B1}	≥ 0.68 E_K	$I_{B1} \geq 0.7\ I_K$
E_{B2}	≥ 0.95 E_K	$I_{B2} \geq 0.3\ I_K$
		$I_W \geq <$ 1.0 mA

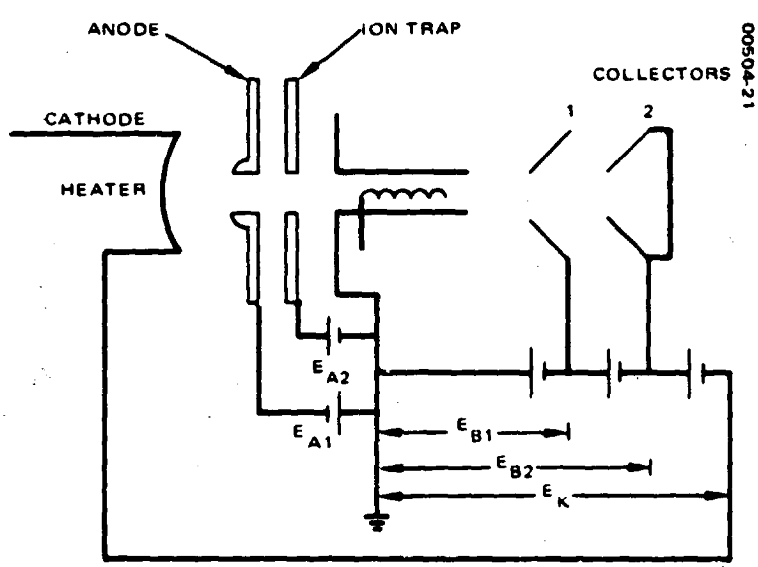


FIGURE 4-21. 25 WATT AND 10 WATT TWT SCHEMATIC

TABLE 4-13. 10 WATT TWTA THERMAL MECHANICAL DATA

Parameter	TWT	Power Supply
Weight, lb	2.2	2.3
Size (H x W x L), in.	2.5 x 2.7 x 13	3 x 3 x 12
Volume, in^3	88	108
Max heat flux density (at collector), W/in^2	5	—

Only two voltages are changed in switching between the high power and low power modes. In the high power mode the cathode voltage is at -4400 volts and the anode 1 voltage is at ground potential. In the low power mode the cathode voltage is -4000 volts and the anode 1 voltage is -800 volts.

The TWT power supply is a pulse-width modulation switching regulator type for high efficiency, and has an efficiency of 83 percent in the high power mode and 62 percent in the low power mode. The unit is packaged in a dielectric-filled enclosure.

Table 4-13 gives weight and power data for the TWT and power supply and also gives the maximum heat flux density of the TWT in the collector area.

4.2.2.5 Output Multiplexer and Harmonic Filter

A simplified version of the output multiplexer design is shown in Figure 4-22. Energy from each TWT is coupled into one of the three circular resonator filters and is coupled from the circular resonators through a directional coupler into the common waveguide manifold. Waveguide isolators are used at each input port and at the output port to ensure that the source and load VSWR presented to each filter is low. Filter design parameters are as follows:

Filter bandwidth	685 MHz
Filter spacing	822 MHz
Filter type	0.01 dB ripple Chebyshev
Number of poles	8

The eight-pole function is realized by using four circular dual mode resonators having orthogonal TE_{111} modes. Circular resonators are used because of their very high Q, which results in low insertion loss and a very flat inband response. Insertion loss of each filter is 0.4 dB. Input and output isolators add 0.2 dB each so that total multiplexer insertion loss is 0.8 dB.

The theoretical response characteristics of the even channel three-filter multiplexer are shown in Figure 4-23. Adjacent filter skirts crossover at about -11 dB, and attenuation of any filter at the passband edge of the adjacent filter is 25 to 30 dB. This attenuation is enough to ensure that adjacent channel interference is negligible.

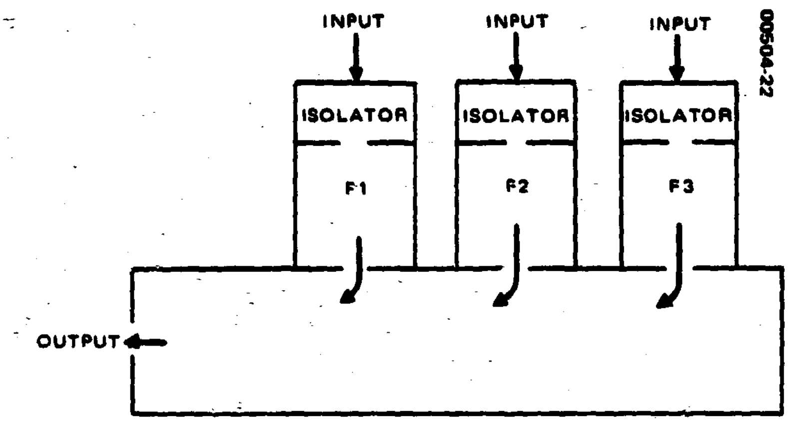


FIGURE 4-22. OUTPUT MULTIPLEXER CONFIGURATION

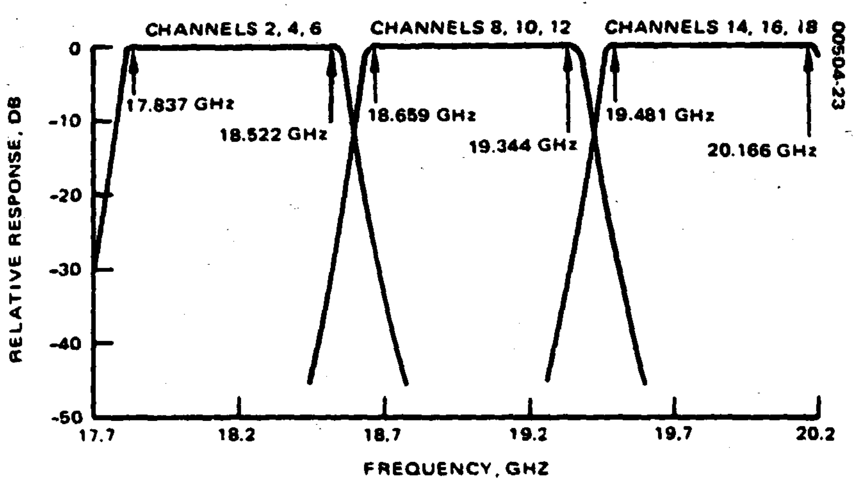


FIGURE 4-23. EVEN CHANNEL FDM REPEATER OUTPUT MULTIPLEXER RESPONSE

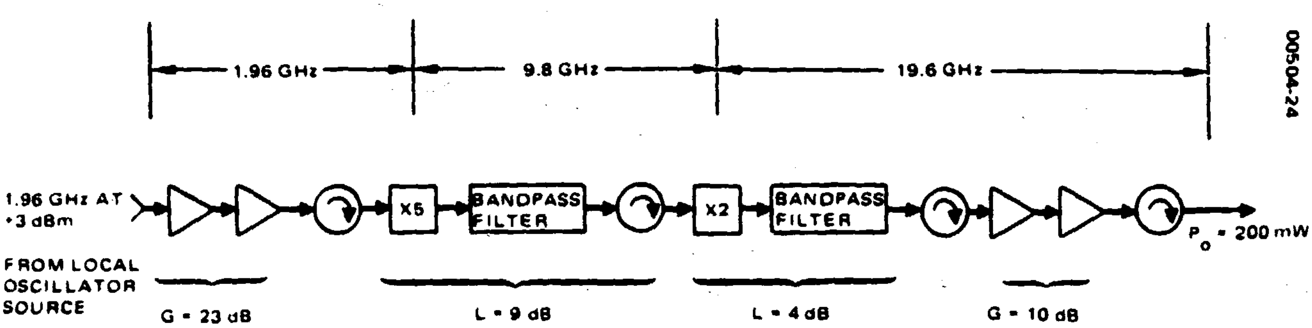


FIGURE 4-24. FDM REPEATER BEACON TRANSMITTER

This has been verified by computer simulation. The output multiplexer is made of Invar in order to achieve low temperature drift. Size and weight of the three-filter multiplexer are as follows:

Envelope size	1.1 x 2.25 x 4.5 in.
Weight	0.90 lb

The harmonic filter is used to reject harmonics of the signal bands generated in the TWT. It is also required to provide approximately 80 dB of attenuation in the repeater receive band of 27.5 to 30 GHz. This is because the TWT is a broadband helix type, which has a broadband noise output substantially above receiver thermal noise. The TWT noise power density per hertz, N_T, is given by $N_T = -174$ dBm/Hz + NF + G, where G and NF are the TWT gain and noise figures and are 55 dB and 30 dB respectively. The resulting TWT output noise density is -89 dBm/ Hz. Assuming that the feed system provides at least 20 dB isolation, 80 dB rejection in the harmonic or low pass filter results in a TWT noise density at the receive antenna feed of -189 dBm/Hz. This is 21 dB below the receiver noise density of -168 dBm/Hz. It should be noted that the harmonic filter is required to provide all of the required rejection since the multiplexer bandpass filters, like all waveguide bandpass filters, will have spurious modes in their upper stop bands.

The harmonic filter is of the waffle-iron type, which provides the very wide stopband required in this application. The cutoff frequency is slightly above 23 GHz and eight poles are required to provide the required attenuation from 27.5 to 30 GHz. Insertion loss in the passband from 17.7 to 20.2 GHz is 0.2 dB.

4.2.2.6 Beacon Transmitter

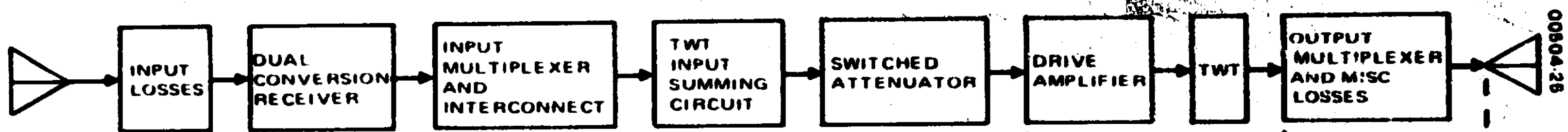
The beacon transmitter baseline design is shown in Figure 4-24. The 1.96 GHz from the local oscillator source is amplified in a two-stage MIC amplifier to a level of 26 dBm. The output is then frequency-multiplied by 5 in a silicon varactor diode multiplier and filtered in a three-pole filter to attenuate undesired harmonics. The resulting 9.8 GHz signal is then frequency-doubled in a GaAs varactor diode doubler and filtered in a three-section WR-51 rectangular waveguide filter. The 19.6 GHz signal level at this point is +13 dBm and is amplified to 200 milliwatts in a two-stage GaAs FET amplifier.

Weight of the beacon transmitter is 1.0 pound and dc power required is 4 watts.

4.2.3 Repeater Gain Distribution

The repeater gain distribution is shown in Figure 4-25 for both the high power and low power modes of operation. Where appropriate, the total spectrum signal level as well as the signal level of a single FDM signal is shown. The signal spectrum consists of up to nine FDM channels until the input multiplexer which follows the receiver. A spectrum consisting of three FDM channe is created at the TWT input summing circuit.

Signal level at the receiver output is deliberately lowered in order to avoid intermodulation of the nine-channel signal spectrum at this point. Gain of the receiver should be sufficiently large to ensure that system noise figure is not degraded by the circuit loss following the receiver, driver amplifier and TWT noise figure. It is also



GAIN, dB	1.1	46	3 0	6.4	3/-8	13	40/55	-2.0	
LOW POWER LEVEL/CH, dBm	-63 3	-64 4	18.4	21 4	-27.8	-30.8	-17.8	+22.2	+20.2
LOW POWER TOTAL LEVEL, dBm	-53 8	54.9	-8.9		-23.0	26 0	-13.0	+27.0	+29.7
HIGH POWER LEVEL/CH, dBm						35.8	22.8	+32 2	+30 2
HIGH POWER TOTAL LEVEL, dBm						31.0	-18.0	+37 0	+39.7

FIGURE 4-25. FDM REPEATER CHANNEL GAIN DISTRIBUTION

desirable to make receiver gain reasonably high in order to reduce the gain required in the TWT driver, since there are 40 TWT driver amplifiers and only 16 receivers. The gain distribution shown in Figure 4-25 represents a reasonable compromise for the conflicting requirements.

4.3 FDM REPEATER PERFORMANCE

4.3.1 FDM Channel Power Output

The amount of power delivered to the antenna feed for each FDM channel is perhaps the most important performance parameter of the FDM repeater. Table 4-14 shows the derivation of this power level. The power delivered to the feed for the individual channel is slightly greater than 1 watt, consistent with the design constraints of the statement of work for a data rate of 137 Mbps.

Table 4-15 shows the breakdown of the output circuit losses. Two cases are shown in Table 4-15; one for normal operation, and one when a spare transmitter channel must be used. Circuit loss increases in this case to 2.3 dB due to addition of a second switch in the TWT output path and added interconnection loss.

TABLE 4-14. FDM REPEATER CHANNEL POWER BUDGET

Parameter	Value
TWT saturated power output	10 dBW
TWT output backoff	-2.5 dB
Channel power sharing loss	-4.77 dB
Intermodulation power loss	-0.5 dB
Output circuit loss to feed	-2.0 dB
Total	0.23 dBW, 1.05 W

TABLE 4-15. FDM OUTPUT CIRCUIT LOSS AND POWER TO FEED

	Loss, dB	
Output Loss Element	Nominal	Spare Amplifier Used
Power monitor	0.1	
Switch	0.2	0.5
Multiplexer	0.6	
Isolator	0.2	
Harmonic filter	0.2	
Waveguide, 6 ft	0.7	
Total	2.0	2.3

Power delivered to feed ≈ 1.0 W.

TABLE 4-16. FDM REPEATER UPLINK BUDGET

	MBPS	
	137	196
Uplink frequency = 30 GHz		196
M (number of bits per symbol) = 2		
Terminal transmitter power, dBW	23.39	24.95
Antenna gain, dB (40 ft dia, 0.06° HPBW)	68.98	
Feed loss, dB	-5.00	-5.00
EIRP, dBW	87.37	88.93
Loss – terminal antenna pointing error, dB (0.01°)	-0.29	
Margin, dB	-1.00	
System aging effects, dB	-1.00	
Random variation of elements, dB	-1.50	
Rain loss, dB (0.125% outage, CCIR rainfall region 4)	-20.03	
Polarization loss, dB	-0.25	
Atmospheric loss, dB	-0.59	
Propagation loss, dB (latitude = 47.5°, relative longitude = 27.5°)	-213.73	
Beam edge loss, dB	-0.00	
Spacecraft stationkeeping loss, dB	-0.00	
Spacecraft antenna pointing error, dB (0.05°)	-0.92	
Spacecraft feed loss, dB	-1.10	
Spacecraft antenna gain, dB (14 ft dia, 0.17° HPBW)	58.80	
Spacecraft received carrier power, dBW	-94.24	-92.68
Received noise power density, dBW Hz (T(R) = 530, T(E) = 700)	-200.15	
Bandwidth, dB (Hz)(BT = 2.00)	81.37	-82.92
Spacecraft receiver noise power, dBW	-118.78	-117.22
Uplink carrier-to-noise power ratio, dB	24.54	24.54

4.3.2 FDM Repeater Communication Link Budgets

Table 4-16 shows the FDM repeater uplink budget. This budget is based upon the one furnished by NASA's Lewis Research Center. Changes from the original budget have been made to be consistent with the results of this study. Spacecraft antenna gain, feed loss, and receiver noise power density line items were changed to be consistent with the predicted performance. The second column data rate was also changed to apply to the highest single channel data rate (Dallas to New York) used in the FDM repeater, and scaling changes were made in the terminal transmitter power for this case. The received E_b/N_o ratio is slightly higher than the 23 dB in the original LERC budget.

Table 4-17 shows the FDM repeater downlink budget. Again, this budget is based upon the LERC budget, and the changes to it are based upon the FDM repeaters predicted performance. The received E_b/N_o for 137 Mbps is 15.24 dB, including the uplink noise contribution. This leaves 4.71 dB total for BER degradation at 10^{-6} BER, due to all causes. The E_b/N_o for 196 Mbps is 13.84 dB and the margin remaining for BER degradation at 10^{-6} BER is 3.31 dB.

TABLE 4-17. FDM REPEATER DOWNLINK BUDGET

	MBPS	
	137	196
Downlink frequency = 20 GHz		
M (number of bits per symbol) = 2		
Spacecraft output power, dBW	2.23	2.23
Antenna gain, dB (14 ft dia, 0.75° HPBW)	56.03	
Feed loss, dB	-2.00	-2.00
EIRP, dBW	56.26	56.26
Loss – spacecraft antenna pointing error, dB (0.05°)	-0.39	
Spacecraft stationkeeping loss, dB	-0.00	
Margin, dB	-1.00	
System aging effects, dB	-1.00	
Random variation of elements, dB	-1.50	
Rain loss, dB (0.073% outage, CCIR rainfall region 4)	-10.03	
Polarization loss, dB	-0.25	
Atmospheric loss, dB	-0.70	
Propagation loss, dB (latitude = 47.5°, relative longitude = 27.5°)	-210.21	
Beam edge loss, dB	0.00	
Terminal antenna pointing error, dB (0.01°)	-0.13	
Terminal antenna feed loss, dB	-1.50	
Terminal antenna gain, dB (40 ft dia, 0.09° HPBW)	65.46	
Terminal received carrier power, dBW	-105.19	-105.19
Received noise power density, dBW Hz (T(R) = 150°, T(E) = 423°)	-202.34	
Bandwidth, dB (Hz) (BT = 2.00)	81.37	82.92
Uplink noise contribution, dB (uplink E_B/N_O = 23.00 dB)	0.54	0.39
Terminal receiver noise power, dBW	-120.43	-119.03
Link carrier to-noise power ratio, dB	15.24	13.84
Hardware implementation loss, dB	4.71	3.31
E_B/N_O, dB (BER = 10^{-6})	10.53	10.53

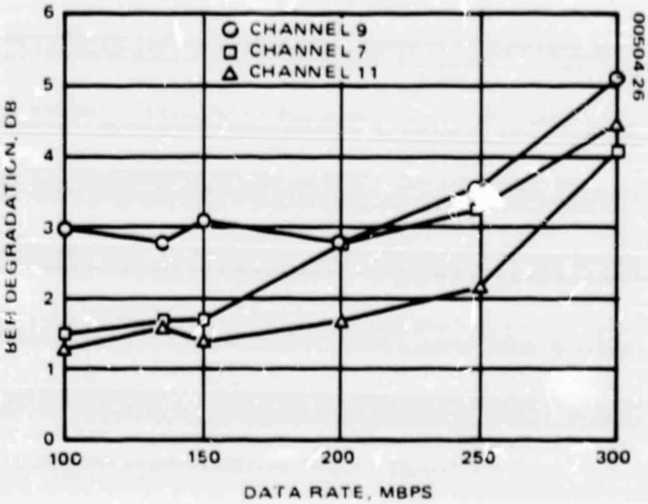


FIGURE 4-26. BER DEGRADATION VERSUS DATA RATE AT 10^{-6} BER

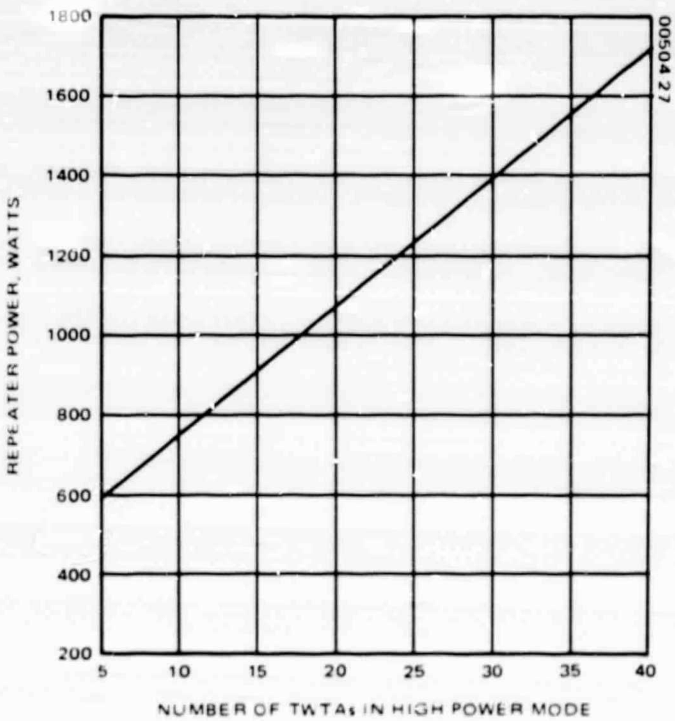
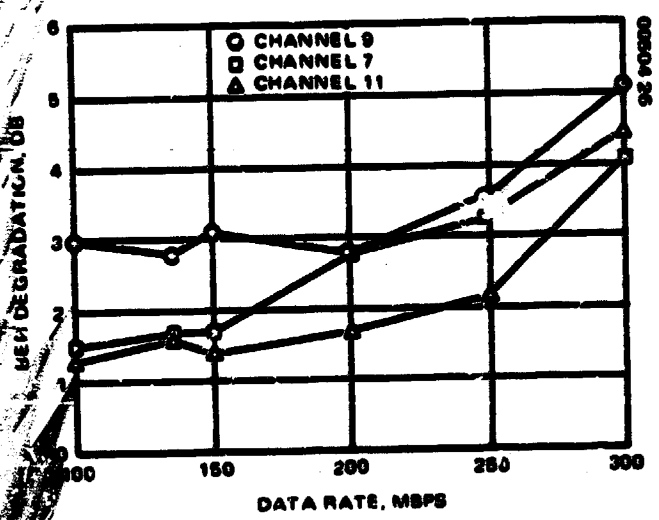


FIGURE 4-27. FDM REPEATER POWER AS FUNCTION OF NUMBER OF TWTAs IN HIGH POWER MODE



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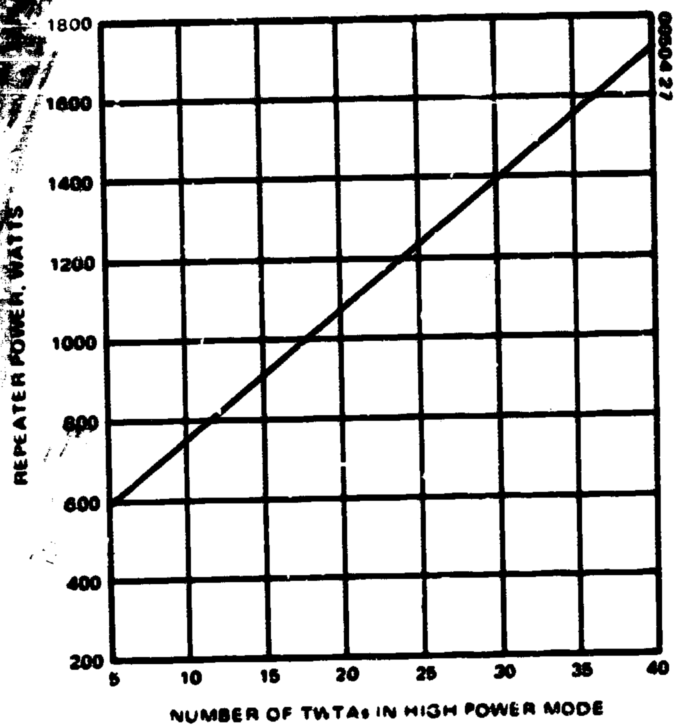

FIGURE 4-27. FDM REPEATER POWER AS FUNCTION OF NUMBER OF TWTAs IN HIGH POWER MODE

4.3.3 BER Degradation Performance

The prime contributor to BER degradation is the nonlinearity of the TWT. A secondary contributor is the effect of the input multiplexer filter used to separate the uplink spectrum. The three-channel multiplexer filters following the TWT appear to have little effect on BER degradation.

Figure 4-26 shows the results of a computer simulation in which the link data rate was varied from 100 to 300 Mbps. The conditions for this simulation are as follows:

Ground transmitter filter	Three-pole Butterworth, BT = 1.0
Ground receiver detection filter	Three-pole Butterworth, BT = 0.53
FDM repeater input multiplexer	Four-pole, 0.1 dB ripple, equal ripple bandwidth of 137 MHz
TWT output multiplexer	Eight-pole, 0.01 dB ripple, equal ripple bandwidth of 685 MHz
TWT input and output backoff	7 dB, 2.8 dB
Number of channels on	Nine odd-channels
Bit error rate	1×10^{-6}

The three channels, 7, 9, and 11 shown in Figure 4-26 are the lower, middle and upper channels which are amplified in a common TWT. The BER degradation of the middle channel appears to be dominated by the TWT nonlinearity and does not vary significantly with data rate until the data rate exceeds 200 Mbps. The lower and upper channels behave in the same general fashion, but the lower channel seems to degrade more rapidly with data rate. The reason for this is not understood at this time but it is believed to be due to the TWT intermodulation products. In any event, the BER degradation is within acceptable limits for all three channels up to at least 200 Mbps. The maximum data rate that would be used in any channel of the FDM repeater is 196 Mbps for the New York to Dallas and Dallas to New York links.

Based upon the results shown in Figure 4-26 and the link power budgets of Tables 4-16 and 4-17 it appears that the FDM repeater meets the basic requirements of the statement of work although there is very little performance margin at the maximum data rate of 196 Mbps. The remainder of the statement of work performance requirements, such as local oscillator stability, spurious responses, etc., are not difficult to meet, and the baseline FDM repeater design should meet them with little difficulty.

4.3.4 Baseline Design Power and Weight

Table 4-18 is a breakdown of the estimated power and weight of the components of the repeater. The power required by the TWTAs is based upon 24 units operating in the high power mode. Figure 4-27 shows how the transponder power varies as a function of the number of TWTAs operated in the high power mode.

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TWT input and output backoff	7 dB, 2.8 dB
Number of channels on	Nine odd-channels
Bit error rate	1×10^{-6}

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TABLE 4-18. FDM REPEATER WEIGHT AND POWER

Component	Quantity (Operating Plus Redundant)	Weight, lb Unit	Weight, lb Total	Power, W Unit	Power, W Total
Receivers	16 + 16	3.0	96.0	4.5	72.0
Input multiplexer filters	114	0.3	34.2	–	–
Receiver redundancy switches	64	0.3	19.2	–	–
Local oscillator redundancy switches	16	0.3	4.8	–	–
Interconnect	114	8.0	8.0	–	–
TWT driver and switched attenuator	40 + 20	0.2	12.0	0.3	12.0
10 W TWTA	40 + 20	4.5	270.0	40.8	1088*
Beacon transmitter	1 + 1	1.0	2.0	4.0	4.0
Output multiplexer filters	40	0.3	12.0	–	–
20 GHz redundancy switches	120	0.8	96.0	–	–
Local oscillator source	1 + 1	4.0	8.0	10.0	20.0
Power monitor	40	0.2	8.0	0.2	8.0
Harmonic filter	16	0.4	6.4	–	–
Repeater miscellaneous			10	–	–
Total			587		1204

*24 TWTAs in high power mode.

4.4 TELEMETRY AND COMMAND REQUIREMENTS

The command requirements for the FDM repeater are as follows:

Command Function	Number of Commands
Receiver select	32
Transmitter select/ power ON/OFF	240
Local oscillator source select	16
Transmitter power level	60
Beacon transmitter select	2

Each receiver select command is used to switch the pair of circulator switches at the receiver input and output and also to switch power to the selected receiver. These commands not only provide for selection of redundant receivers but can also be used to turn off the receivers at the node that is not in use.

TABLE 4-18. FDM REPEATER WEIGHT AND POWER

Component	Quantity (Operating Plus Redundant)	Weight, lb		Power, W	
		Unit	Total	Unit	Total
Receivers	16 + 16	3.0	96.0	4.5	72.0
Input multiplexer filters	114	0.3	34.2	-	-
Receiver redundancy switches	64	0.3	19.2	-	-
Local oscillator redundancy switches	16	0.3	4.8	-	-
Interconnect	114	8.0	8.0	-	-
TWT driver and switched attenuator	40 + 20	0.2	12.0	0.3	12.0
10 W TWTA	40 + 20	4.5	270.0	40.8	1088*
Beacon transmitter	1 + 1	1.0	2.0	4.0	4.0
Output multiplexer filters	40	0.3	12.0	-	-
20 GHz redundancy switches	120	0.8	96.0	-	-
Local oscillator source	1 + 1	4.0	8.0	10.0	20.0
Power monitor	40	0.2	8.0	0.2	8.0
Harmonic filter	16	0.4	6.4	-	-
Repeater miscellaneous			10	-	-
Total			587		1204

*24 TWTAs in high power mode.

4.4 TELEMETRY AND COMMAND REQUIREMENTS

The command requirements for the FDM repeater are as follows:

Command Function	Number of Commands
Receiver select	32
Transmitter select/ power ON/OFF	240
Local oscillator source select	16
Transmitter power level	60
Beacon transmitter select	2

Each receiver select command is used to switch the pair of circulator switches at the receiver input and output and also to switch power to the selected receiver. These commands not only provide for selection of redundant receivers but can also be used to turn off the receivers at the node that is not in use.

The transmitter select commands require 180 commands, since each pair of crossover switches associated with a transmitter are three-position switches. Separate commands are provided to control power on or off to each transmitter.

Sixteen commands are required to independently select the source of the local oscillator for each receiver. Sixty commands are required to select the high power or low power mode for each transmitter. These commands control the TWT power supply and the switched attenuator, which determines the TWT RF drive level. Two commands are provided to select the beacon transmitter and can also be used to turn off the beacon transmitter if desired.

The telemetry requirements of the FDM repeater are as follows:

Function	Signal Type	Number Required
Received power level monitor	Analog	16
Transmitter power level monitor	Analog	40
Transmitter switch status	Digital-2 bits	60
Transmitter power status	Discrete	60
Receiver selection and power	Discrete	32
TWTA parameters	Analog	180
Secondary voltages	Analog	20
Temperature monitors	Analog	10

5. TDM REPEATER DESIGN

This section describes the satellite switched time division multiplex (TDM) repeater trades, baseline design, and performance parameters.

5.1 TDM REPEATER DESIGN TRADES

The design tradeoffs in the TDM repeater are less complex than in the FDM repeater. The basic repeater concept is illustrated in Figure 5-1. The design trades to be considered are as follows:

1) Switch matrix frequency selection

2) Burst rate and switch matrix requirements

3) Switch matrix design including device selection and switch architecture

4) Method of carrier signal level control

5) Beacon transmitter frequency selection

5.1.1 Switch Matrix Frequency Selection

As shown previously, dual frequency conversion is necessary to avoid serious spurious response problems in the repeater. Although this problem could be avoided by the use of a demodulate/remodulate type repeater in conjunction with a baseband switch matrix, this implementation is not considered here because of the obvious substantial increase in satellite complexity.

The choice then reduces to implementation of the switch matrix at the intermediate frequency or at the ouput frequency of 17.7 to 20.2 GHz. Implementation of the switch matrix at the output frequency is simply beyond the capability of currently available technology and implementation at the IF is the only feasible choice.

5.1.2 Burst Data Rate and IF Switch Matrix Requirements

This design trade concerns the choice of the burst data rate and the degree of channelization in the TDM transponder. Factors to be considered in this tradeoff are TDM repeater complexity, repeater weight and power, and bandwidth required of the IF matrix switch. A complete system tradeoff should also consider the complexity and difficulty of earth station burst modems as a function of data rate. However, such a

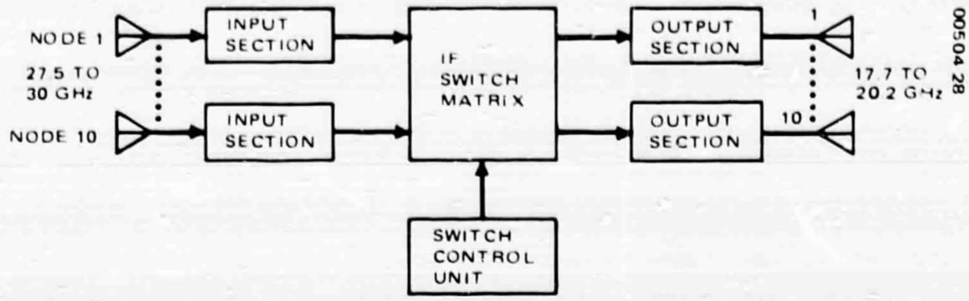


FIGURE 5-1. TDM REPEATER DESIGN

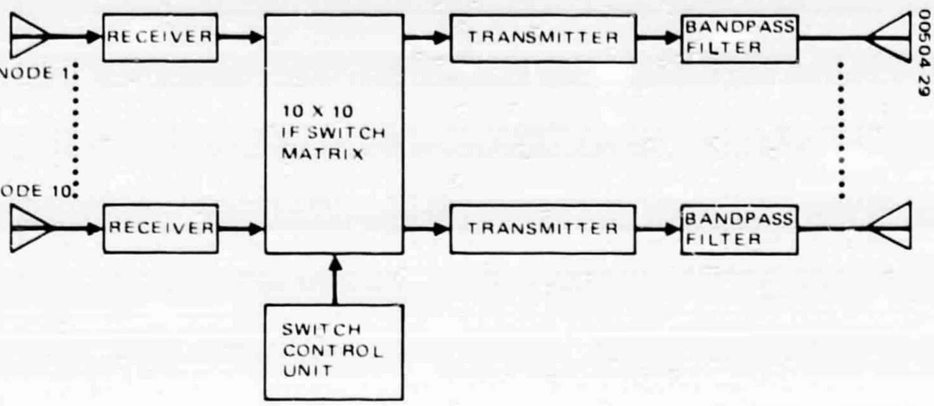


FIGURE 5-2. TDM REPEATER CONFIGURATION FOR 2.5 GBPS BURST DATA RATE

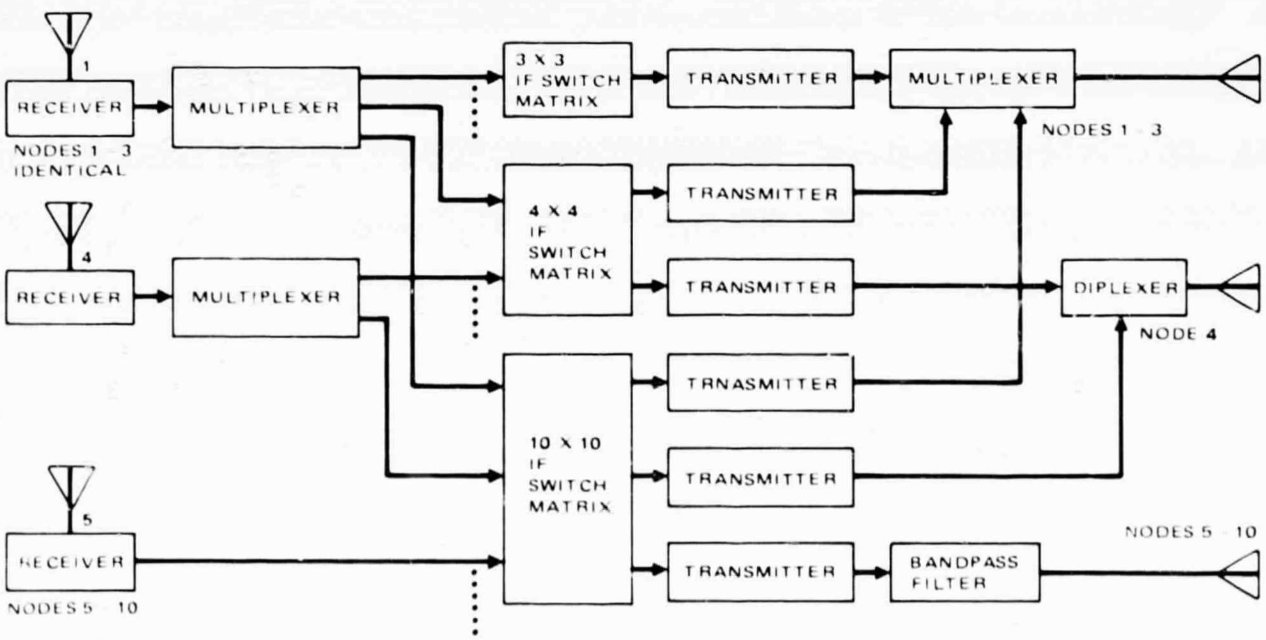


FIGURE 5-3. TDM REPEATER CONFIGURATION FOR 833 MBPS BURST DATA RATE

tradeoff is beyond the scope of this study which is restricted to the satellite repeater implementation.

The data rate for the city having the most traffic, New York, is 2.5 Gbps, while the lowest data rate is 521 Mbps for Denver. Figure 5-2 illustrates the simplest TDM repeater conceptual design wherein the burst data rate is 2.5 Gbps for all nodes. This configuration has the advantage of having ten identical receiver and transmitter designs. The disadvantages are that the matrix switch must cover the entire IF band of 3.98 to 6.48 GHz, and the lower data rate transmitters do not use transmitter power efficiently since their duty cycle in this type of burst system is simply the data rate required by the node or city divided by the 2.5 Gbps burst data rate.

An alternate conceptual design, which reduces the bandwidth requirement for the IF switch matrix and utilizes transmitter power more efficiently, is illustrated in Figure 5-3. The burst data rate in this case is 833 Mbps. Nodes 1 to 3 (New York, Chicago, Los Angeles) demultiplex the uplink spectrum into three separate frequency bands and have three transmitters and a three-channel output multiplexer. Node 4 (San Francisco) divides the uplink spectrum into two separate bands and has two transmitters and an output diplexer. Nodes 5 to 10, all of which have data rates less than or very close to 833 Mbps, do not require an input demultiplexer or an output multiplexer and have only one transmitter.

The three IF switch matrices are arranged to provide complete interconnectivity and also are arranged so that each switch matrix has the same frequency inputs and outputs to reduce the switch matrix bandwidth.

Table 5-1 compares the features of the two FDM repeater concepts. The RF circuit loss from the transmitter to the antenna feed is slightly greater for the concept of Figure 5-3 due to the presence of the output multiplexer. In cases where no output multiplexer is used the increased insertion loss is due to the more narrow bandwidth of the output bandpass filter. BER degradation performance is expected to be somewhat worse for the configuration of Figure 5-3 due to bandwidth restrictions imposed by multiplexing the input and output signals of the repeater.

The 10 watt transmitter power output in the configuration of Figure 5-3 is derived by scaling to 1/3 of the 25 watt transmitter used for the 2.5 Gbps burst data rate and increasing the scaled power of 8.33 watts to 10 watts to compensate for the higher RF output circuit loss and the slightly degraded BER performance.

The total spacecraft weight impact given in Table 5-1 for the two configurations includes the solar array and power conditioning circuit weight necessary to support the power required by each transponder. The configuration of Figure 5-2 has a net advantage of approximately 9 percent less spacecraft weight impact. The one disadvantage of the 2.5 Gbps burst data rate configuration is the IF switch matrix bandwidth requirement of 2.5 GHz; however, this magnitude of bandwidth does appear feasible for the IF switch matrix. The configuration of Figure 5-2 is chosen as the baseline design for the TDM repeater because of its lower spacecraft weight impact and simpler design. However, in the event that the IF matrix switch bandwidth cannot be achieved, the configuration of Figure 5-3 is considered an acceptable alternative.

TABLE 5-1. COMPARISON OF TDM REPEATER CONCEPTS OF FIGURES 5-2 AND 5-3

Parameter	Figure 5-2 Concept	Figure 5-3 Concept
Burst data rate, Gbps	2.5	0.833
IF switch matrix bandwidth	2.5 GHz	833 MHz
Number of transmitters	10+6(1)	17+9(1)
RF circuit loss, transmitter to feed	–	0.3 to 0.5 dB greater
Transmitter type	TWT	TWT
Transmitter power output, W	25/2.5(2)	10/1(2)
Transmitter dc power, W	90/14(2)	40/8(2)
BER degradation performance	–	Slightly worse for multiple channel nodes
Repeater weight, lb	241	318
Repeater power, W	671(3)	577(4)
Spacecraft weight impact(5), lb	509	549

NOTES:
1) Operating plus redundant
2) High power/low power mode
3) 6 of 10 transmitters in high power mode
4) 10 of 17 transmitters in high power mode
5) Power assessed at 0.4 lb/W

TABLE 5-2. IF SWITCH MATRIX REQUIREMENTS

Matrix order	10 x 10
Center frequency	5.23 GHz
Bandwidth	2.5 GHz
Switching time	10 ns
Reconfiguration time	2 μsec
Isolation	50 dB
Insertion loss	15 dB
Differential gain	1 dB
Input and output VSWR	1.25

TABLE 5-3. CIRCUIT ARCHITECTURE OPTIONS

Option	Advantages	Disadvantages
Biplanar	Smallest size	Complex assembly, high VSWR
Coupler crossbar	Medium size, good decoupling	Complex assembly, high insertion loss
Rearrangeable multistage	Simplest modules	Reliability questionable, very complex control unit
Cubic center junction	Low insertion loss	Matrix order limited, isolation reduced as order increases
Cubic power divider	Simplest construction	Insertion loss

5.1.3 IF Switch Matrix Design Tradeoffs

The two primary design trades for the IF switch matrix are the switch device and the method of interconnecting switches or circuit architecture. The basic performance requirements for the IF switch matrix are given in Table 5-2.

The circuit architecture options which were considered are compared in Table 5-3. The cubic power divider was selected on the basis of simple construction. The closest competitor to this selection was the coupler crossbar configuration. The various architecture concepts are illustrated in Figures 5-4 and 5-5.

The primary candidates considered for the switch device were the GaAs FET switch and the pin diode. Tables 5-4 and 5-5 give a summary of the advantages of the pin diode switch and the GaAs FET switch, respectively. The GaAs FET switch has been chosen in preference to the pin diode switch on the basis of its lower power requirement for a given switching speed, potential compatibility with monolithic microwave integrated circuits, and simpler driver circuit requirement. Table 5-6 summarizes the design selection rationale for the IF switch matrix.

TABLE 5-5. FET OPTION

FET Advantages	FET Disadvantages
Fast transition time	Technology needs more development to achieve monolithics
Low dc bias power	Insertion loss of nonamplifying mode high
Simple TTL interface driver circuit	
Amenable to monolithic integration	
Amplification possibility exists	

Reference:

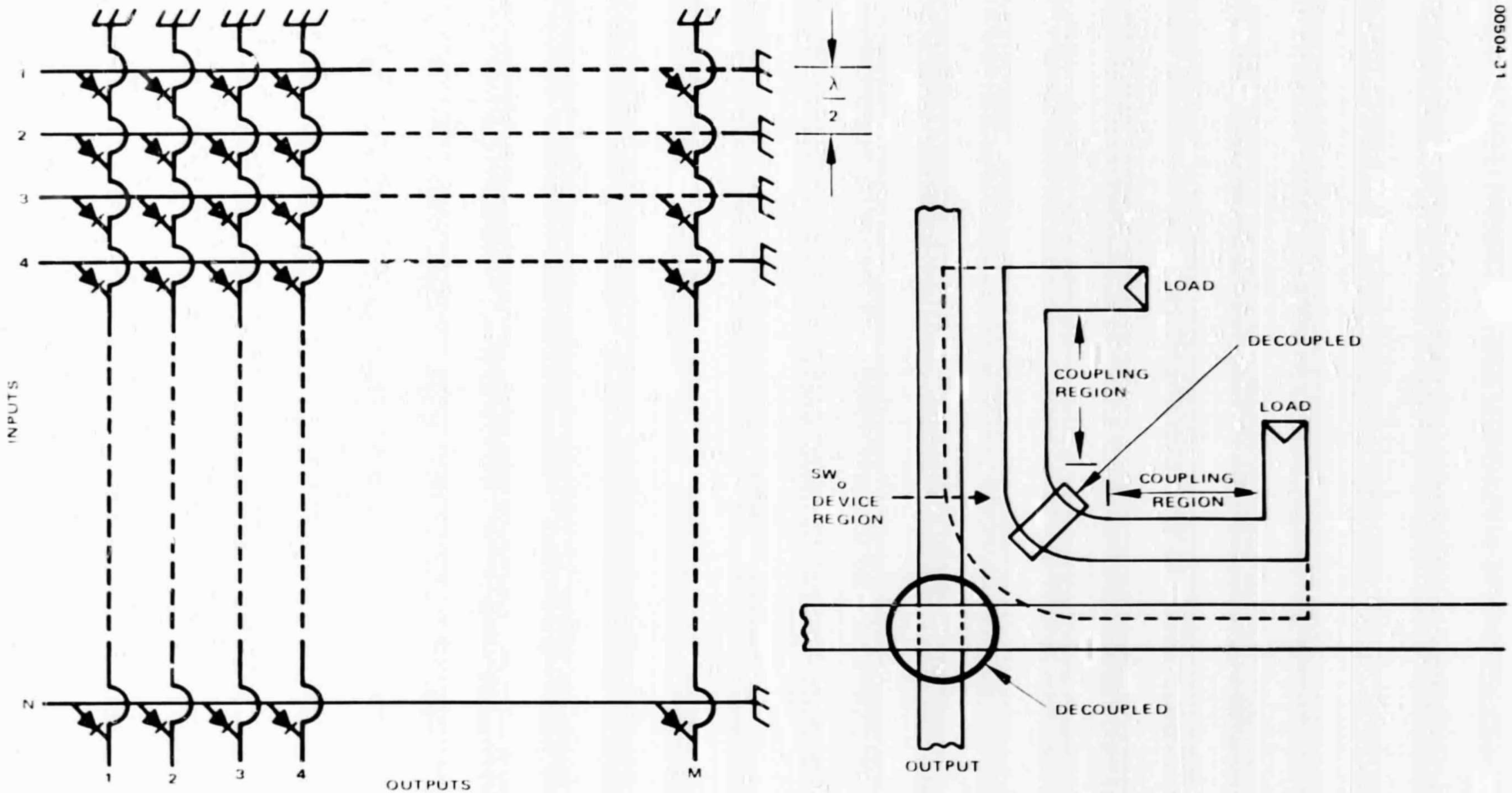
R. Gaspari and H. Yee (Hughes Aircraft Company)
"Microwave GaAs FET Switching," G-MTT (IEEE International Microwave Symposium Digest) Ottawa, 1978

TABLE 5-4. PIN DIODE OPTION

Advantages	Disadvantages
Low insertion loss	Higher bias power – slower transition time
Good isolation	More complex driver
Good bandwidth	Not amenable to monolithic integration
Common technology	

References:

1) F ASSAL (COMSAT) and X ROZEC (Thompson – CSF), "Microwave Switch Matrix for Communications Satellites," ICC-76 (IEEE International Conference on Communications Conference Digest), June 1976
2) X ROZEC, "16 x 16 Fast Switching Matrix," European Conference Proceedings (Paris), September 1978



BIPLANAR

COUPLER CROSS BAR VARIATION

FIGURE 5-4. CIRCUIT ARCHITECTURE OPTIONS

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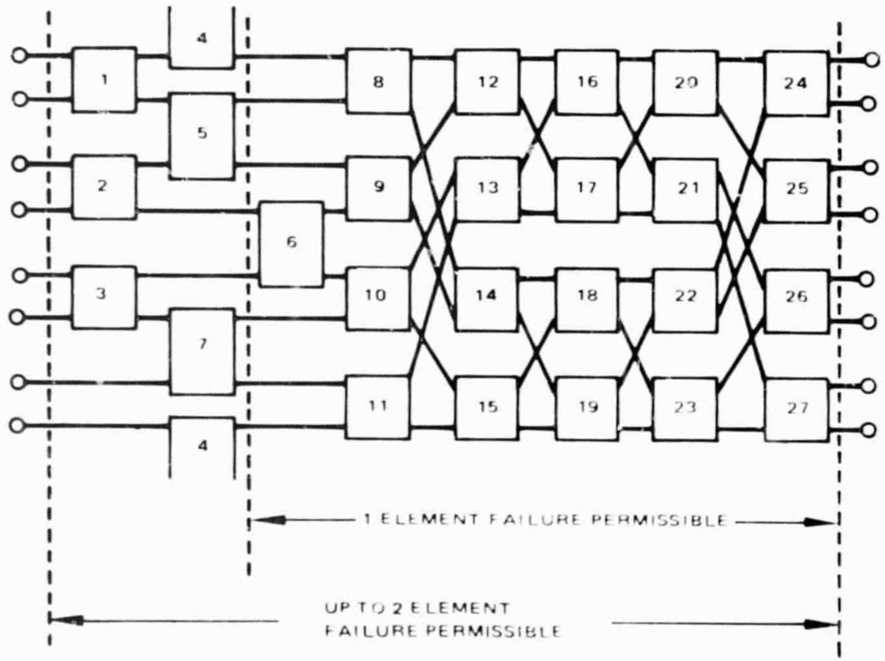
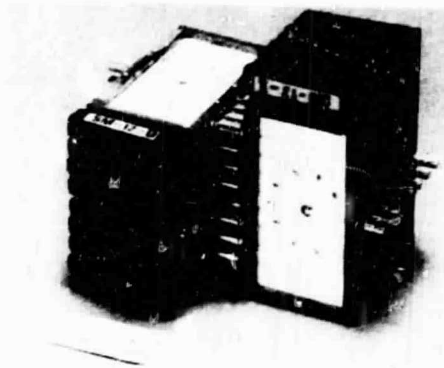
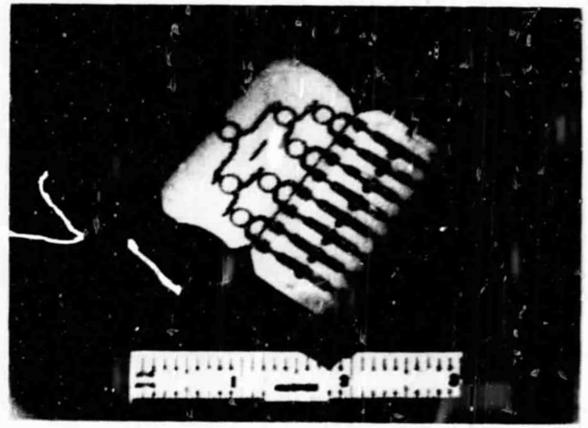


a) REARRANGEABLE MULTISTAGE REDUNDANT STRUCTURE



b) CENTER JUNCTION MODULE



c) SP8T SWITCH MODULE USING POWER DIVIDER APPROACH

FIGURE 5-5. ARCHITECTURE OPTIONS - REARRANGEABLE MULTISTAGE AND CUBIC

TABLE 5-6. IF SWITCH MATRIX DESIGN SELECTION

Choice	Design Selection	Rationale
Circuit	Cubic, power divider	Simplest construction, lowest cost, very high reliability, good isolation (also provides broadcast capability)
Device	FET	High composite reliability, monolithic integration compatibility (also provides fast speed, low dc power)
Driver	FET multiple-FET-chip	Integrated 3-FET circuit minimizes attachment bonding operations (also introduces integration concept)
IF	5.23 GHz	Multiple-FET-chip discrete breadboard prefers lower frequencies (also compatible with receiver requirement)

5.1.4 Method of Output Carrier Signal Level Control

In the TDM baseline repeater, each output TWT is driven by sequential bursts of ten different inputs. It is important to maintain the TWT input drive level within a relatively narrow range in order to maintain the TWT output power nearly constant. The duration of each burst signal can be very short in duration, which tends to preclude the use of an automatic gain control circuit since the response time of the AGC loop would have to be a small fraction of the minimum burst duration of 2 μsec.

A more attractive alternate is to use a hard limiter preceding the TWT. GaAs FET amplifiers have superior limiting characteristics and also can be designed to have a lower AM to PM conversion coefficient than a TWT. Since the limiter tends to remove most of the AM that may be present on the input signal, the AM/PM conversion coefficient of the combined limiter and TWT closely approaches that of the limiter by itself. The limiter can be designed to compress a total input level variation of $\pm$5dB to an output variation of $\pm$0.2 dB. Total output power variation for a TWT input variation of $\pm$1dB is typically about $\pm$0.2 dB, which is quite acceptable. Typical performance of limiters designed for this application at X band and Ku band is an AM/AM coefficient of 0.2 dB/dB and an AM/PM conversion coefficient of 2 deg./dB.

The preferred method of TWT drive level control for the TDM repeater is a GaAs FET limiter which provides excellent gain compression and low AM/PM conversion coefficient over a wide bandwidth and also has a typical response time to input level changes of less than 10 ns.

5.1.5 Beacon Transmitter Frequency Selection

The basis for selection of a beacon transmitter frequency is to avoid significant interference from the data. ERP for each data transmitter is higher than in the FDM (see 4.1.5) case by 10 $\log_{10}$ $(2.5 \times 10^{9}/137 \times 10^{6})$ and is 98.6 dBm. The ratio of data ERP to beacon ERP is then 60.6 dB and the discrete spectral line power relative to total carrier power, as derived in 4.1.5, should be more than 60 dB below the carrier.

The beacon carrier must be coherent with the satellite local oscillator and the frequency must lie between 17.7 GHz and 20.2 GHz. The data carrier frequencies are centered at 18.95 GHz, and the first nulls of the 2.5 Gbps QPSK modulated spectrum are at 17.7 and 20.2 GHz. The closest multiples of the satellite local oscillator fundamental frequency of 122.5 MHz that are inside the 17.7 to 20.2 GHz band are 145 times 122.5 MHz and 164 times 122.5 MHz. These frequencies are 17.7625 GHz and 20.09 GHz. Discrete spectral line power for a 2.5 Gbps QPSK modulated spectrum is -70.8 dBc for the lower frequency and -65.6 dBc for the upper frequency. The lower frequency is therefore chosen for the beacon transmitter since it will provide a beacon carrier to data interference ratio of at least 10 dB. Spectral line separation is 4.883 MHz for a 512 bit data frame, and the spectral line closest to 17.7625 GHz is approximately 980 kHz below the nominal beacon frequency. The chance of the interfering spectral line being very close to the beacon carrier is remote since a frequency tolerance of 5 x 10^{-6} on the satellite local oscillator and the ground transmitter results in a total offset frequency variation of 230 kHz in the frequency separation.

5.2 TDM REPEATER BASELINE DESIGN

Figure 5-6 is a block diagram of the TDM repeater baseline configuration. Only one node or channel of the repeater is shown in complete detail since all nodes are identical. The input circuits employ a 2 for 1 redundancy in the single conversion receiver. The output circuits use 16 for 10 redundancy for the transmitter section. The transmitter section consists of the upconverter which converts the IF to the output frequency band, a limiting amplifier which provides gain and maintains a constant TWT drive level, the switched attenuator, and the 25 watt TWT power amplifier. Waveguide CX switches identical to those described in 4.2.2.1 are used at the output of the TWT. A power monitor is provided to monitor the health of each transmitter downlink. A bandpass filter is used to limit the spectrum of the downlink signal to the 17.7 to 20.2 GHz allocated frequency range. The harmonic filter attenuates TWT harmonics and TWT noise in the repeater receive frequency band.

5.2.1 Input Circuits

The TDM repeater input circuits consist of the test coupler, the ferrite circulator switches which provide redundancy switching at the receiver input, the single conversion receiver, and the coaxial switches used to select the receiver IF output. A detailed block diagram of the single conversion receiver is shown in Figure 5-7. The receiver is very similar to the FDM receiver described in 4.2.1.1. The only differences are deletion of the AGC function and elimination of the second frequency conversion. A detailed description of the remaining circuits was given in 3.1 and 4.2.1.

The size, weight, and power of the TDM receiver are as follows:

Size	6 x 8 x 2 inches
Weight	2.5 pounds
Power	6.0 watts

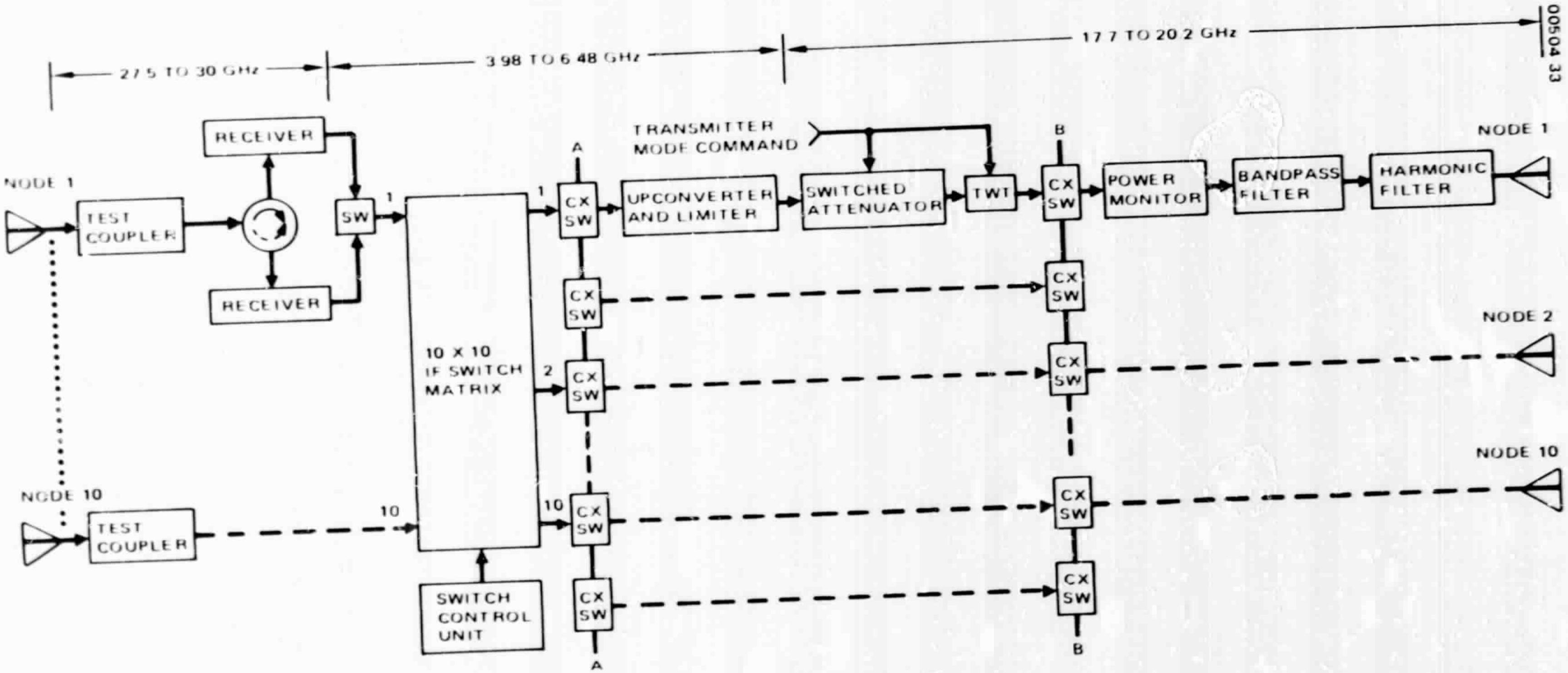


FIGURE 5-6. TDM REPEATER BASELINE DESIGN

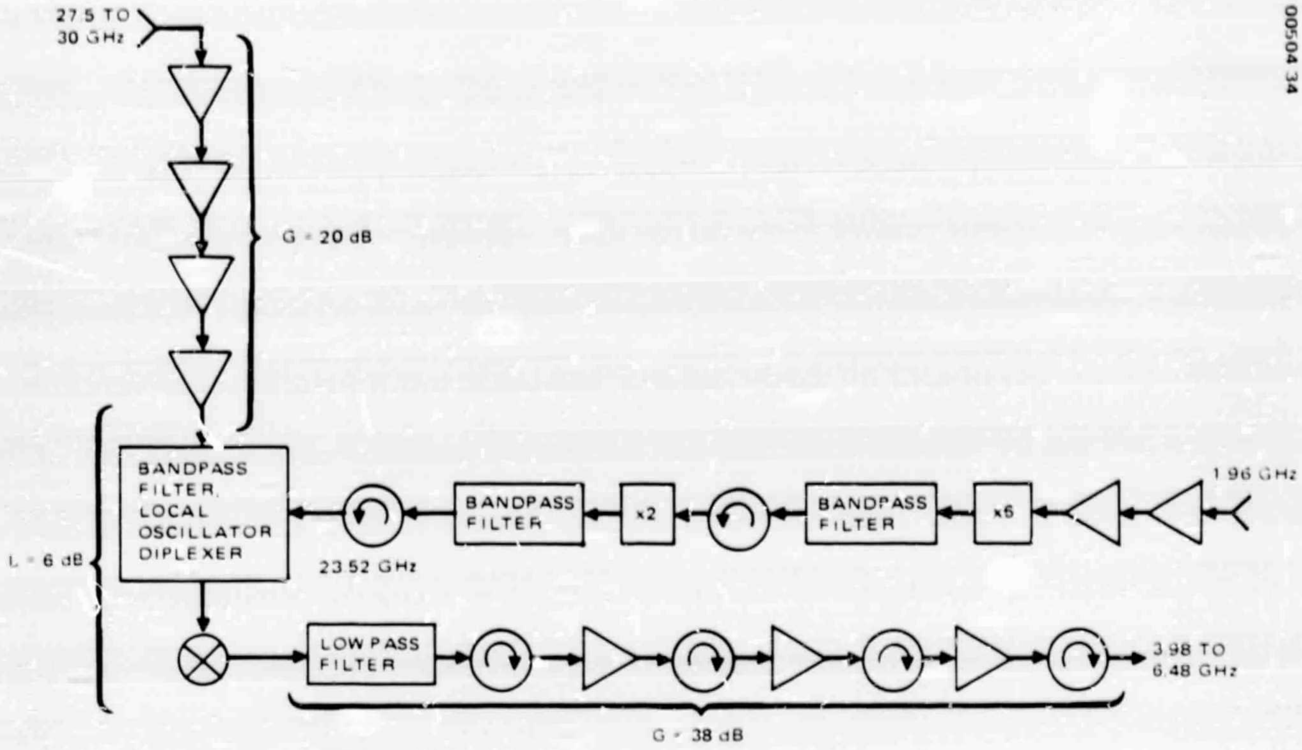


FIGURE 5-7. TDM SINGLE CONVERSION RECEIVER

TABLE 5-7. IF SWITCH MATRIX PARAMETERS

Parameter	Projected
Matrix size	12 x 12 with 2 redundant (operational 10 x 10)
Connectivity	Full, with broadcast capability built in
Switching time	10 ns
Command signals	Parallel
Control circuit	Driver part of GaAs integrated circuit
IF	5230 MHz
Bandwidth	2500 MHz
DC power	3.5 (excluding DCU) watts
Size	5.2 x 5.2 x 5.2 in.
Weight	4 lb
Reliability, 10 yr	0.99988
Switch element device	Multiple-FET-chip
Circuit configuration	cubic

5.2.2 IF Switch Matrix

The parameters of the TDM repeater IF switch matrix are given in Table 5-7. Note that the switch matrix is implemented with a 12x12 design to provide two redundant switch paths in order to increase switch reliability.

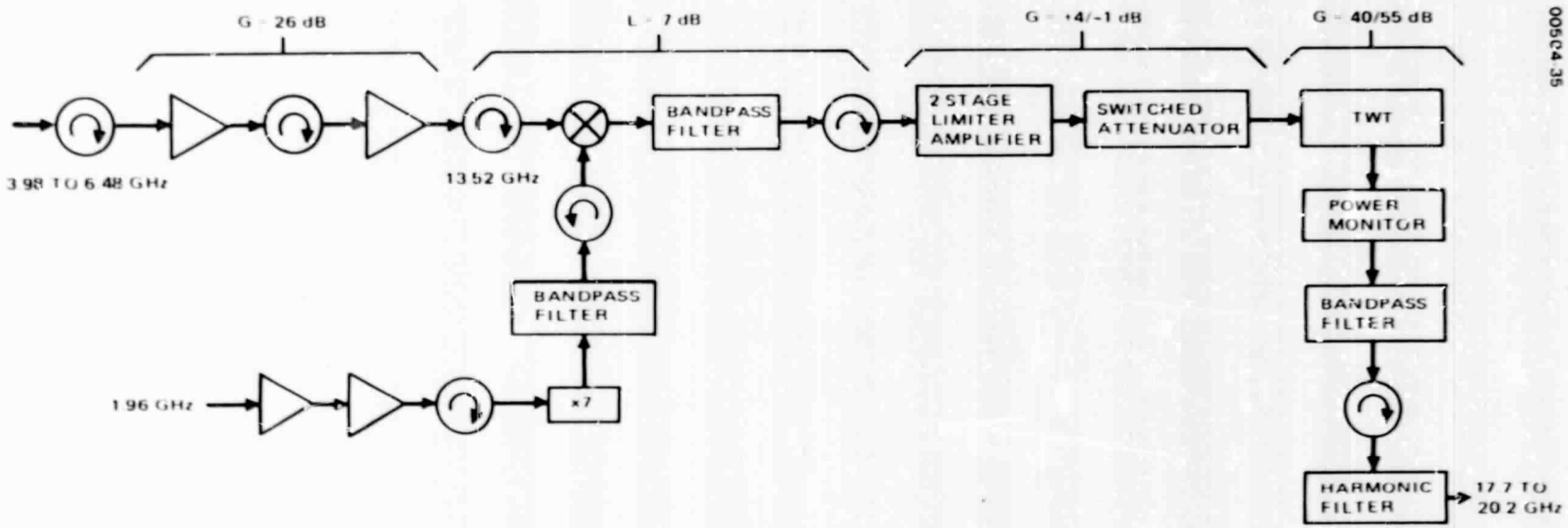


FIGURE 5-8. TDM REPEATER OUTPUT CIRCUIT

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5.2.3 Output Circuits

A detailed block diagram of a single output circuit without the redundancy switches is shown in Figure 5-8. A two-stage IF amplifier is used to amplify the signal prior to upconversion. The IF amplifier is constructed in MIC form and is identical to the IF amplifier used in the receiver as described in 4.2.1.1.

The upconverter and local oscillator frequency multiplier are identical to the circuits used in the FDM receiver. Following the upconverter, a three-section waveguide bandpass filter with a bandwidth of 2.5 GHz is used to attenuate spurious products and local oscillator leakage of the upconverter mixer.

A two-stage GaAs FET limiting amplifier follows the upconverter and bandpass filter. This amplifier has a linear gain of 12 dB and is driven into gain compression to reduce TWT input drive level variation. The switched attenuator is identical to that used in the FDM repeater as described in 4.2.2.2.

The 25 watt TWT is a helix type with a dual depressed collector to improve efficiency and is capable of operation at the full 25 watt saturated power output or at a reduced level of 2.5 watts for normal operation. The TWT is operated fully saturated in both the high power and low power modes.

Table 5-8 lists the key performance parameters of the TWT. Table 5-9 gives the beam voltages, beam currents, and total input dc power of the TWT for the high power and low power modes of operation. Table 5-10 lists the TWT electrode voltages and curents of the 25 watt TWT. As in the 10 watt FDM TWT the only voltages that require switching in the high and low power modes of operation are the beam voltages and the anode 1 voltage. Table 5-11 gives the thermal and mechanical data for the TWT and the TWT power supply. Design of the power supply is basically the same as for the 10 watt TWT power supply described in 4.2.2.4 except for the increased beam voltage and current. Dielectric encapsulation is used to prevent arcing and corona in the space environment.

TABLE 5-8. 25 WATT TWT KEY PERFORMANCE PARAMETERS (TDM)

Parameter	Value
Frequency	17.7 to 20.2 GHz
Output power	
Low mode	2.5 W
High mode	25 W
Saturated gain	
Low mode	40 dB
High mode	55 dB
Efficiency (two stage collector)	
Low mode	25%
High mode	35%

TABLE 5-9. 25 WATT TWT OPERATING PARAMETERS

Parameter	Low Mode	High Mode
Beam voltage, V	6300	7000
Beam current, mA	10	36
DC power, W	10	72

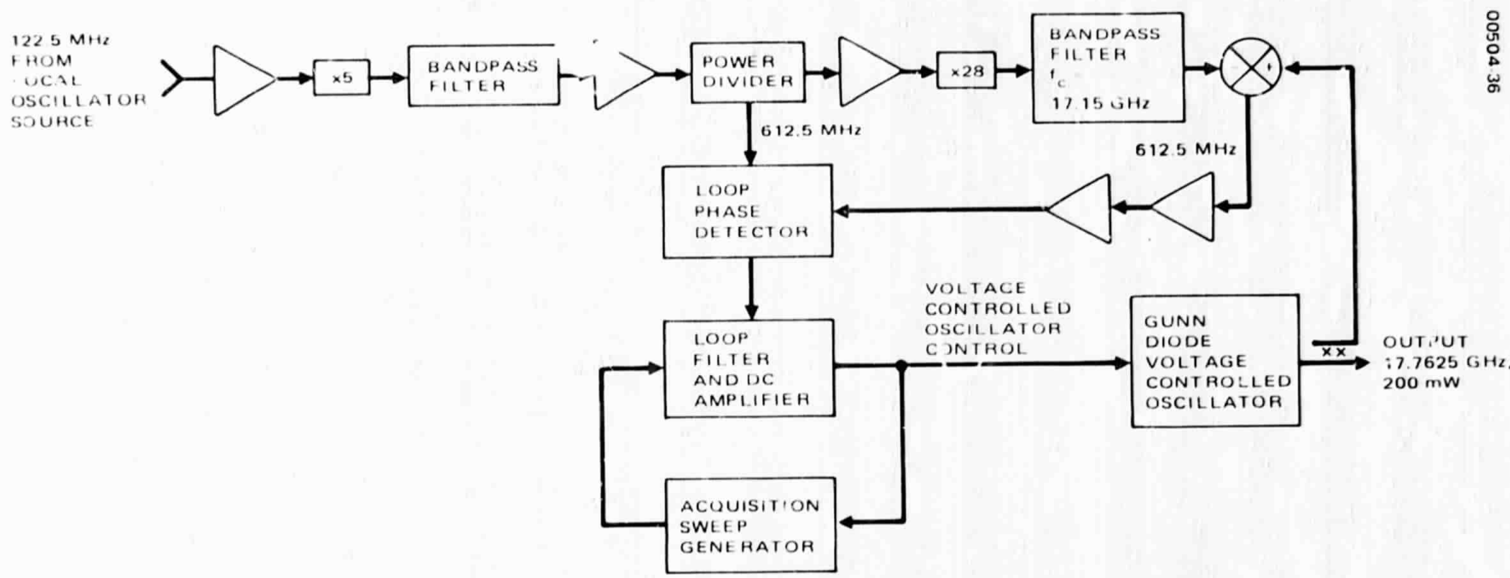


FIGURE 5-9. TDM REPEATER BEACON TRANSMITTER

TABLE 5-10. 25 WATT TWT ELECTRODE VOLTAGES AND CURRENTS

Voltages		Currents	
E_K	$\cong$ -6300 to -7000	I_K	$\cong$ 10 to 36 mA
E_{A1}	$\cong$ -1200 to 0V	I_{A1}	< 0.1 mA
E_{A2}	$\cong$ +100 V fixed	I_{A2}	< 0.1 mA
E_F	$\cong$ 6.0 V	I_F	$\cong$ 0.8 A
E_{B1}	$\cong 0.68\ E_K$	I_{B1}	$\cong 0.7\ I_K$
E_{B2}	$\cong 0.95\ E_K$	I_{B2}	$\cong 0.3\ I_K$
		I_W	< 2 mA

TABLE 5-11. 25 WATT TWTA THERMAL/MECHANICAL DATA

Parameter	TWT	Power Supply
Weight, lb	2.2	2.8
Size (H x W x L), in.	2.5 x 2.7 x 13	3 x 4 x 12
Volume, in.^{3}	88	108
Maximum heat dissipation (at collector), W/in.^{2}	13	–

The remaining output circuits consist of the power monitor, the output bandpass filter, and the harmonic filter. The power monitor is nearly identical to the design used in the FDM repeater except that the coupling ratio of the directional coupler is reduced because of the higher power level. The harmonic filter is identical to that used in the FDM repeater.

The output bandpass filter is a five-section, 0.1 dB ripple Chebyshev filter with an equal ripple bandwidth of 1.9 GHz and a 3 dB bandwidth of 2.16 GHz. Center frequency of the filter is 18.89 GHz, which is 60 MHz lower than the nominal center frequency in order to produce nearly equal attenuation at the band edge frequencies. Attenuation at 17.7 and 20.2 GHz is 11 dB. Because of the high percentage bandwidth this filter is realized in rectangular WR-51 waveguide.

5.2.4 Beacon Transmitter Design

Figure 5-9 is a block diagram of the beacon transmitter for the TDM repeater. The design is somewhat more complex than the FDM repeater beacon transmitter because the 155th harmonic of the local oscillator fundamental frequency of 122.5 MHz must be generated as the transmitter output frequency. A phaselocked approach has been chosen to generate the desired output frequency because of the high order of frequency multiplication.

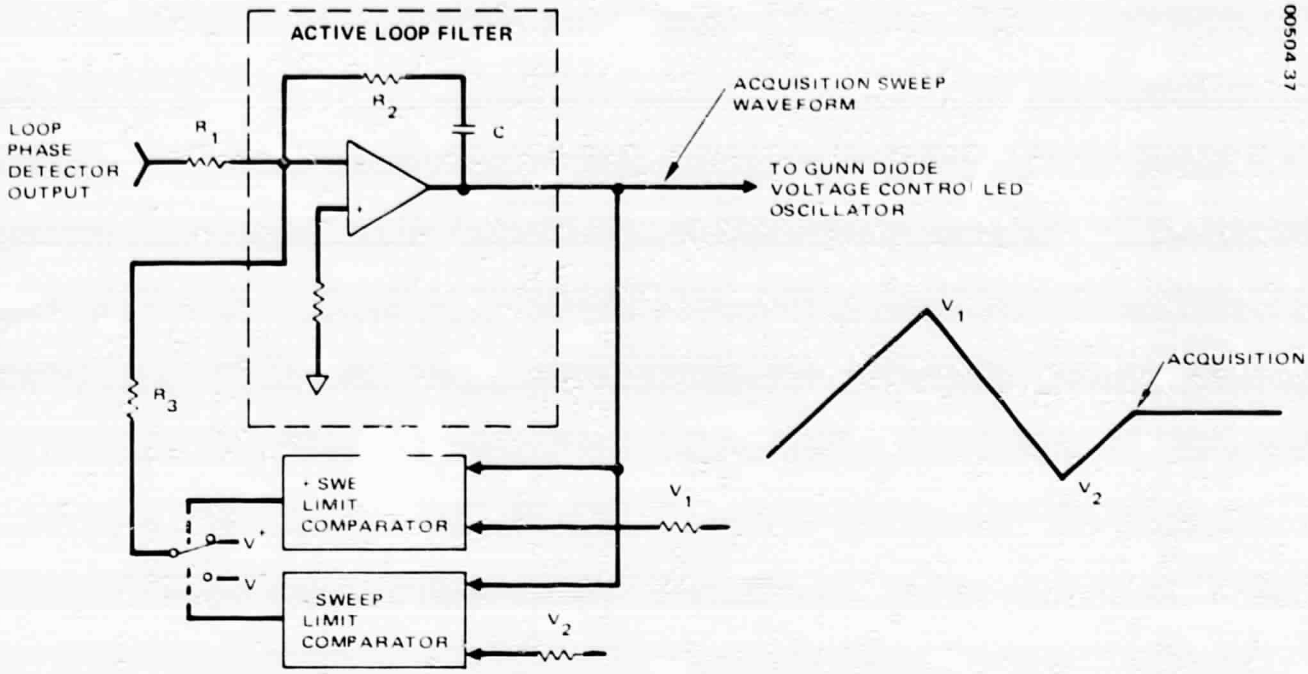


FIGURE 5-10. TDM BEACON TRANSMITTER PHASE LOCK CIRCUIT DIAGRAM

The Gunn diode voltage controlled oscillator (VCO) free running frequency is tuned to approximately the 145th harmonic of 127.5 MHz, 17.7625 GHz. A coupler at the VCO output directs a small portion of the VCO output to a mixer where it is mixed with the output of a X140 frequency multiplier chain. The difference frequency, nominally 612.5, is amplified and phase compared with 612.5 MHz generated in the first X5 multiplier of the X140 multiplier. The phase detector is amplified and filtered in an active loop filter and phase locks the Gunn diode VCO to the desired output frequency.

Since the free running frequency of the VCO can be substantially offset from the phase lock frequency it is necessary to provide an acquisition aid to establish phase lock. This is accomplished by using the active loop filter to generate a frequency sweep as shown in the schematic of Figure 5-10. During acquisition the VCO control voltage output is sampled by two voltage comparators, each of which has a fixed bias reference input. The V^+ and V^- voltages are selected by a toggle switch whose state is controlled by the comparators. A small current is always fed through resistor R3 into the integrator formed by the operational amplifier, resistor R2, and capacitor C. When the upper or lower sweep limit is reached, a comparator state change reverses the toggle switch and reverses the sweep. The sweep frequency range is deliberately made much larger than the VCO frequency uncertainty. When the VCO sweeps through the desired frequency, the error signal generated by the loop phase detector takes over and maintains the VCO in phase lock. It is not necessary to disable the sweep once phase lock has occurred since the small tracking error caused by the sweep current is constant and can be made quite small by increasing the sweep period.

The major portion of the dc power required in the beacon transmitter is used by the Gunn diode VCO. The remaining circuits are all low or moderate signal level circuits which require about 1 watt total. Weight of the beacon transmitter is 1.5 pounds and total dc power required is 4 watts.

5.2.5 TDM Repeater Gain Distribution

Figure 5-11 shows the gain distribution of the TDM repeater. Since each uplink and downlink channel is a single QPSK modulated carrier, the gain distribution is less critical than it is in the FDM repeater and more gain can be provided in the receiver. The microwave switch matrix has a rather high insertion loss since the architecture is based upon the cubic power divider which introduces a high insertion loss in both the input and output circuits of the switch. This gain is made up in a two stage IF amplifier which follows the IF switch matrix. Gain compression in a GaAs FET limiter is provided to maintain a nearly constant TWT drive level.

5.3 TDM REPEATER PERFORMANCE

5.3.1 TDM Repeater Power Output

The power delivered to the feed is an important TDM repeater performance parameter. Table 5-12 shows the breakdown of circuit loss between the TDM TWT output and the antenna feed. The total power delivered to the feed is 17.3 watts or 12.4 dBw. The statement of work constraints give a required power of 3.0 watts for 548 Mbps. If this is scaled to 2.5 Gbps, the required power would be 13.7 watts.

5.3.2 TDM Repeater Communication Link Budgets

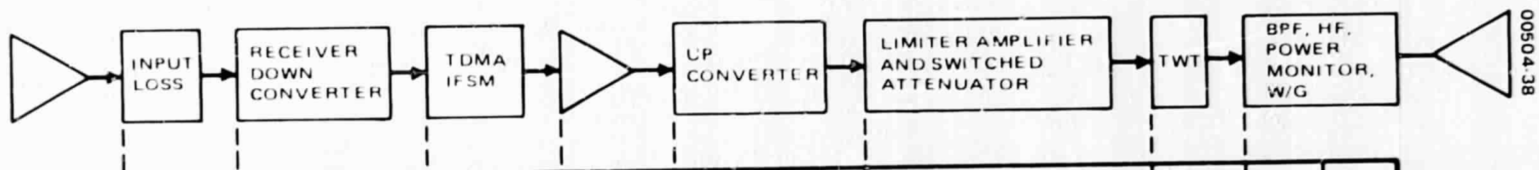
Tables 5-13 and 5-14 are the communication uplink and downlink budgets for the 2.5 Gbps burst rate TDM repeater. These budgets are based upon NASA LERC budgets and the only time items changed are those where the predicted TDM repeater values have been substituted. The uplink E_b/N_o is 25.64 dB, which is somewhat higher than the original value of 23.00 dB.

The downlink E_b/N_o including the noise contributed by the uplink is 13.78 dB, which allows a total of 3.25 dB for implementation loss.

5.3.3 BER Degradation and Other Parameters

The principal TDM repeater contributors to BER degradation are the input bandpass filter located in the receiver prior to the first mixer and the output bandpass filter following the TWT. Figure 5-12 is a summary of the effects of these two filters and shows that the total TDM repeater degradation is 1.3 dB for the filter characteristics listed in the figure when interference is neglected. If link to link isolation is only 20dB, the degradation is 1.8 dB at a BER of 10^{-6}.

Based on Figure 5-12 and the communication link budgets of Tables 5-13 and 5-14, it appears that the TDM repeater baseline design meets the general requirements of the statement of work system constraints. Other requirements of the statement of work are relatively easy to meet with the possible exception of spurious outputs which are required to be -50 dBc within any downlink channel. This requirement implies that the IF switch matrix individual channel to channel isolation should be -60 dB in order to meet a total requirement of -50 dB since there are nine channels of undesired signal coupling to the desired signal. However, a computer simulation indicated only 0.5 dB of BER degradation at a downlink carrier to interference ratio of 20 dB. This implies that a total isolation in the range of 30 to 40 dB would be adequate.



GAIN, dB	-1.1	52	-27	26	-7	+4.2/-0.8	40/55	-2	
LOW POWER LEVEL, dBm	53.1	54.2	-2.2	-29.2	-3.2	-10.2	-6	+34	+32
HIGH POWER LEVEL, dBm							-11	+44	+42

FIGURE 5-11. TDM REPEATER CHANNEL GAIN DISTRIBUTION

TABLE 5-12. TDM OUTPUT CIRCUIT LOSS AND POWER TO FEED

Output Loss Element	Loss, dB	
	Nominal	Spare Amplifier Used
Power monitor	0.1	
Switch	0.2	0.5
Bandpass filter	0.2	
Isolator	0.2	
Harmonic filter	0.2	
Waveguide, 6 ft	0.7	
Total	1.6	1.9

Power delivered to feed = 12.4 dBW, 17.3 W.

TABLE 5-13. TDM REPEATER UPLINK BUDGET

Uplink frequency = 30 GHz	
Bit rate = 2.5 GHz	
M (number of bits per symbol) = 2	
Terminal transmitter power (2196 W)	33.42 dBW
Antenna gain (40 ft dia, 0.06° HPBW)	68.98 dB
Feed loss	-2.40 dB
EIRP	100.00 dBW
Loss-terminal antenna pointing error (0.01°)	-0.29 dB
Margin	-1.00 dB
System aging effects	-1.00 dB
Random variation of elements	-1.50 dB
Rain loss (0.125% outage, CCIR rainfall region 4)	-20.03 dB
Polarization loss	-0.25 dB
Atmospheric loss	-0.59 dB
Propagation loss (latitude = 47.5°, relative longitude = 27.5°)	-213.73 dB
Beam edge loss	-0.00 dB
Spacecraft stationkeeping loss	-0.00 dB
Spacecraft antenna pointing error (0.05°)	-0.02 dB
Spacecraft antenna gain (14 ft dia, 0.17° HPBW)	58.30 dB
Spacecraft received carrier power	-80.51 dBW
Received noise power density (T (R) = 530°, T (E) = 700°)	-200.15 dBW/Hz
Bandwidth 2.5 GHz, BT = 2.00)	94.0 dB(Hz)
Spacecraft receiver noise power	-106.15 dBW
Uplink carrier-to-noise power ratio	25.64 dB

TABLE 5-14. TDM REPEATER DOWNLINK BUDGET

Uplink frequency = 20,000 GHz	
Bit rate = 2.5 Gbps	
M (number of bits per symbol) = 2	
Spacecraft output power (21.09 W)	12.4 dBW
Antenna gain (14.00 ft dia, 0.25° HPBW)	56.34 dB
Feed loss	-1.80 dB
EIRP	66.94 dBW
Loss-spacecraft antenna pointing error (0.05°)	-0.39 dB
Spacecraft stationkeeping loss	-0.00 dB
Margin	-1.00 dB
System aging effects	-1.00 dB
Random variation of elements	-1.50 dB
Rain loss (0.073% outage, CCIR rainfall region 4)	-10.03 dB
Polarization loss	-0.25 dB
Atmospheric loss	-0.70 dB
Propagation loss (latitude = 47.5°, relative longitude = 27.5°)	-210.21 dB
Beam edge loss	-0.00 dB
Terminal antenna pointing error (0.01°)	-0.13 dB
Terminal antenna feed loss	-1.50 dB
Terminal antenna gain (40.00 ft dia, 0.09° HPBW)	65.46 dB
Terminal received carrier power	-94.31 dBW
Received noise power density (T (R) = 150°, T (E) = 420°)	-202.37 dBW/Hz
Bandwidth 2.5 GHz, BT = 2.00)	94.0 dB(Hz)
Uplink noise contribution (uplink E_B/N_o = 25.64 dB)	0.28 dB
Terminal receiver noise power	-108.09 dBW
Link carrier to noise power ratio	13.78 dB
Hardware implementation loss	3.25 dB
E_B/N_o, (BER = 1×10^{-6})	10.53 dB

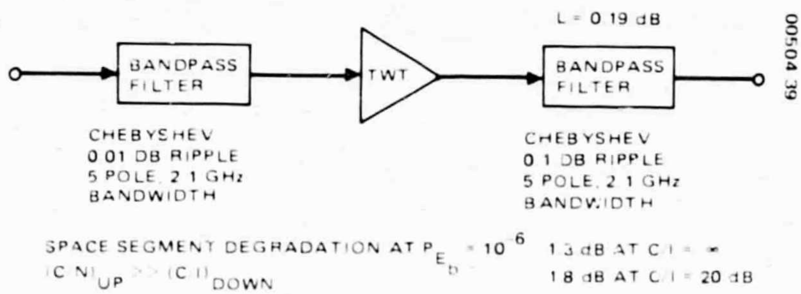


FIGURE 5-12. TDM REPEATER BASELINE FILTER SELECTION

5.3.4 TDM Repeater Weight and Power

Table 5-15 is a breakdown showing the weight and power of the TDM repeater components.

TABLE 5-15. TDM REPEATER WEIGHT AND POWER

Component	Quantity (Operating plus Redundant)	Weight, lb		Power, W	
		Unit	Total	Unit	Total
Redundancy switches - 30 GHz, circulator type	20	0.3	6.0	–	–
Output redundancy switches - waveguide	16	0.8	12.8	–	–
Coaxial redundancy switches	52	0.3	15.6		
Receivers	10 + 10	2.5	50.0	4.0	40.0
10 x 10 IF switch	1	4.0	4.0	3.5	3.5
25 W TWTA	10 + 6	5.0	80.0	84.14	560*
Upconverter, TWT driver, switcher attenuator	10 + 6	1.5	24.0	1.2	12.0
Output bandpass filter and harmonic filter	10	0.5	5.0	–	–
Local oscillator source	1 + 1	4.5	9.0	10.0	20.0
Digital switch control unit	1 + 1	6.0	12.0	6.0	6.0
Beacon transmitters	1 + 1	1.5	3.0	4.0	4.0
Repeater miscellaneous			20.0		5.0
Total			241		651

*6 TWTAs operating in high power mode.

5.4 TDM REPEATER TELEMETRY AND COMMAND REQUIREMENTS

The command requirements for the TDM repeater are as follows:

Command Function	No. of Commands
Receiver select	20
Transmitter select/power on-off	64
Local oscillator source select	26
Transmitter power level	16
Beacon transmitter select	2

The telemetry requirements of the TDM repeater are as follows:

Function	Signal Type	No. Required
Received power level monitor	Analog	10
Transmitter power level monitor	Analog	10
Transmitter switch status	Digital-2 bits	16
Transmitter power status	Discrete	16
Receiver selection and power	Discrete	20
TWTA parameters	Analog	48
Secondary voltages	Analog	15
Temperature monitors	Analog	10

6. TDM/FDM REPEATER DESIGN

This section describes the hybrid TDM/FDM repeater design trades, baseline design, and performance parameters. The TDM/FDM repeater design is based upon the designs developed for the FDM and TDM repeaters in Sections 4 and 5 of this report.

6.1 FDM/TDM REPEATER DESIGN TRADES

The basic concept of the TDM/FDM repeater is illustrated in Figure 6-1. The design concept is based upon the individual FDM and TDM repeater design selections already made. A dual conversion receiver is used for the FDM portion of the repeater and the FDM repeater interconnect is accomplished at the repeater output frequency. The first IF of the dual conversion receiver is used for the TDM portion of the repeater, the switch matrix is implemented at the IF, and the switch matrix output is upconverted to the output frequency range.

6.1.1 FDM/TDM Partitioning

Many of the design trades relevant to the FDM/TDM repeater design have already been discussed in Sections 4 and 5 of this report. The most significant design tradeoff to be considered is the method of dividing the data capacity between the FDM and the TDM portions of the receiver.

Figures 6-2 and 6-3 illustrate two alternative FDM/TDM channel configurations considered for comparison purposes in the study. The configuration shown in Figure 6-2 does not satisfy the statement of work constraints in that the TDM data rates required for New York, Dallas, Minneapolis/St. Paul and Atlanta exceed the nominal burst rate of 548 Mbps. In all cases except New York, the additional data capacity could be added as additional FDM channels. However, these channels would have to be assigned to smaller cities. All available FDM frequency slots for New York, Los Angeles, and Chicago are in use. For example, the additional 263 Mbps capacity required for Dallas would require additional FDM channels for five of the remaining six cities.

The configuration of Figure 6-3 represents a more extreme case of FDM capacity. This configuration approaches the complexity of the FDM-only repeater of Section 4. A more reasonable division of FDM and TDM capacity is illustrated in Figure 6-4. This configuration has been selected for the baseline design because it is fully compliant with the requirements of the traffic matrix, and growth capability is provided in the TDM channel. The full TDM burst rate capability is not used except in the case of New York. The maximum data rate for any 274 MHz FDM channel is 288 Mbps for the two New York to L.A. and the two New York to Chicago channels. Orthogonal

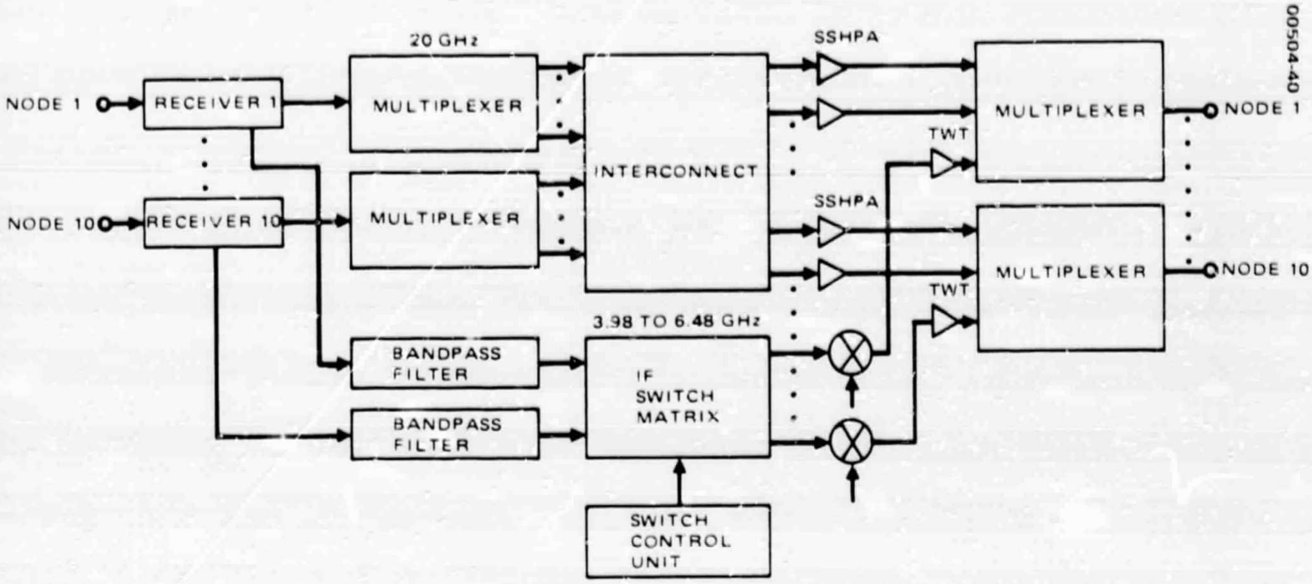


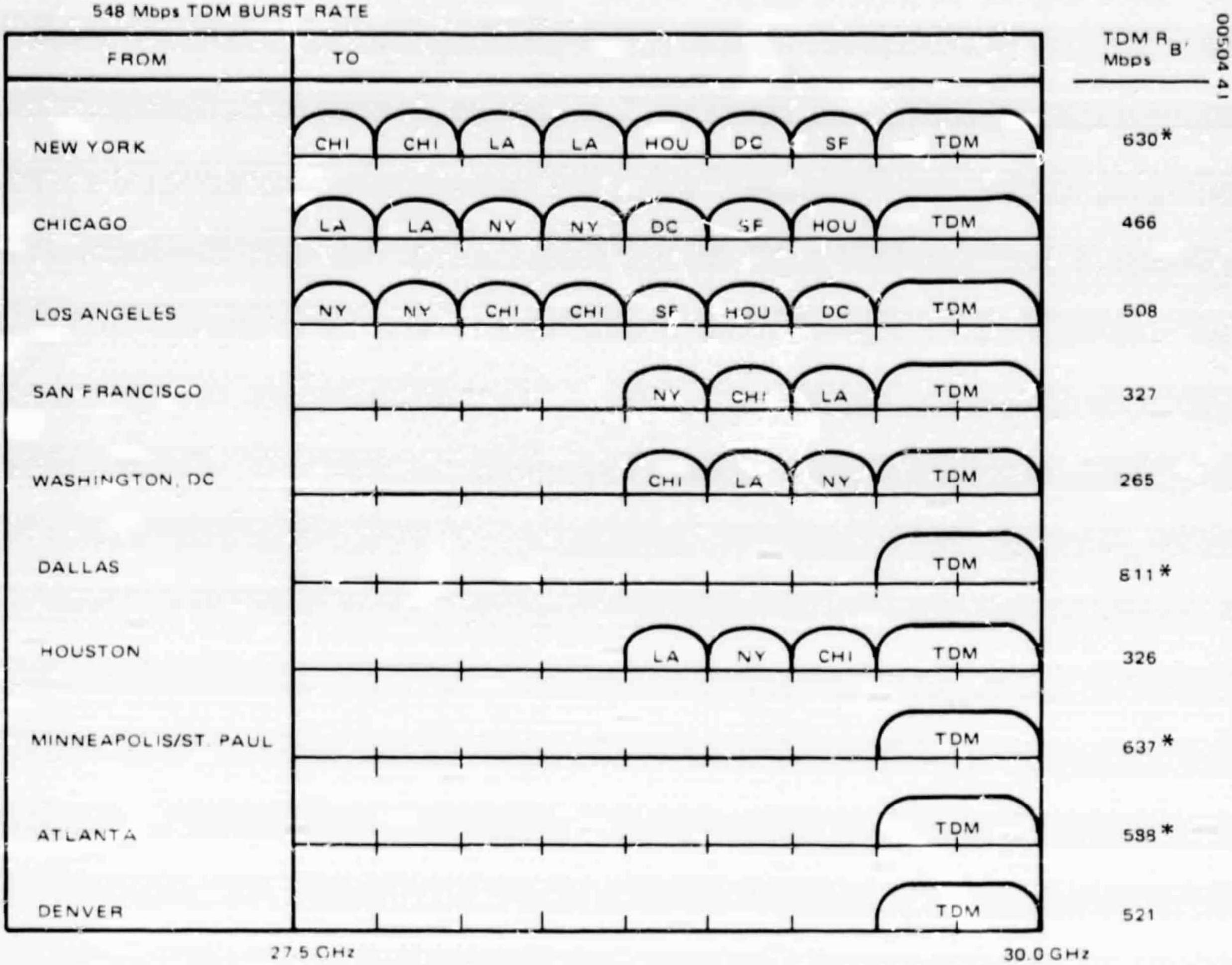
FIGURE 6-1. FDM/TDM REPEATER CONCEPT

polarizations are used for the New York, Chicago, and Los Angeles nodes to simplify the multiplexing requirements. This is almost mandatory to obtain adequate olation between the 1.096 Gbps burst rate TDM channel and the adjacent 274 MHz FDM channel without sacrificing excessive bandwidth in guard bands between channels.

6.1.2 Input Circuit Design Trades

The only significant design trade in the input circuits is to determine the method of separating the FDM and TDM portions of the uplink spectrum. Two possible configurations are illustrated in Figure 6-5. Figure 6-5a shows a configuration where the demultiplexer for both the FDM and TDM spectrum is implemented at the repeater output frequency. This configuration requires ten additional downconverters for the ten TDM channels. In Figure 6-5b the TDM channel filter is implemented at the receiver IF and does not require additional frequency conversion since the receiver IF is readily accessible. The disadvantage of this configuration is that the TDM channel filter is implemented at IF where the percentage bandwidth of the TDM channel is much higher. However, the IF filter can be designed as an interdigital stripline filter compatible with the wide bandwidth. This configuration would weigh slightly more than an RF TDM channel filter. Note that filtering at IF with only a bandpass TDM channel filter does not provide band limiting of the adjacent FDM spectrum. This is not really a disadvantage since the ground station must provide some band limiting of both the TDM and FDM spectrums at the point where they are summed.

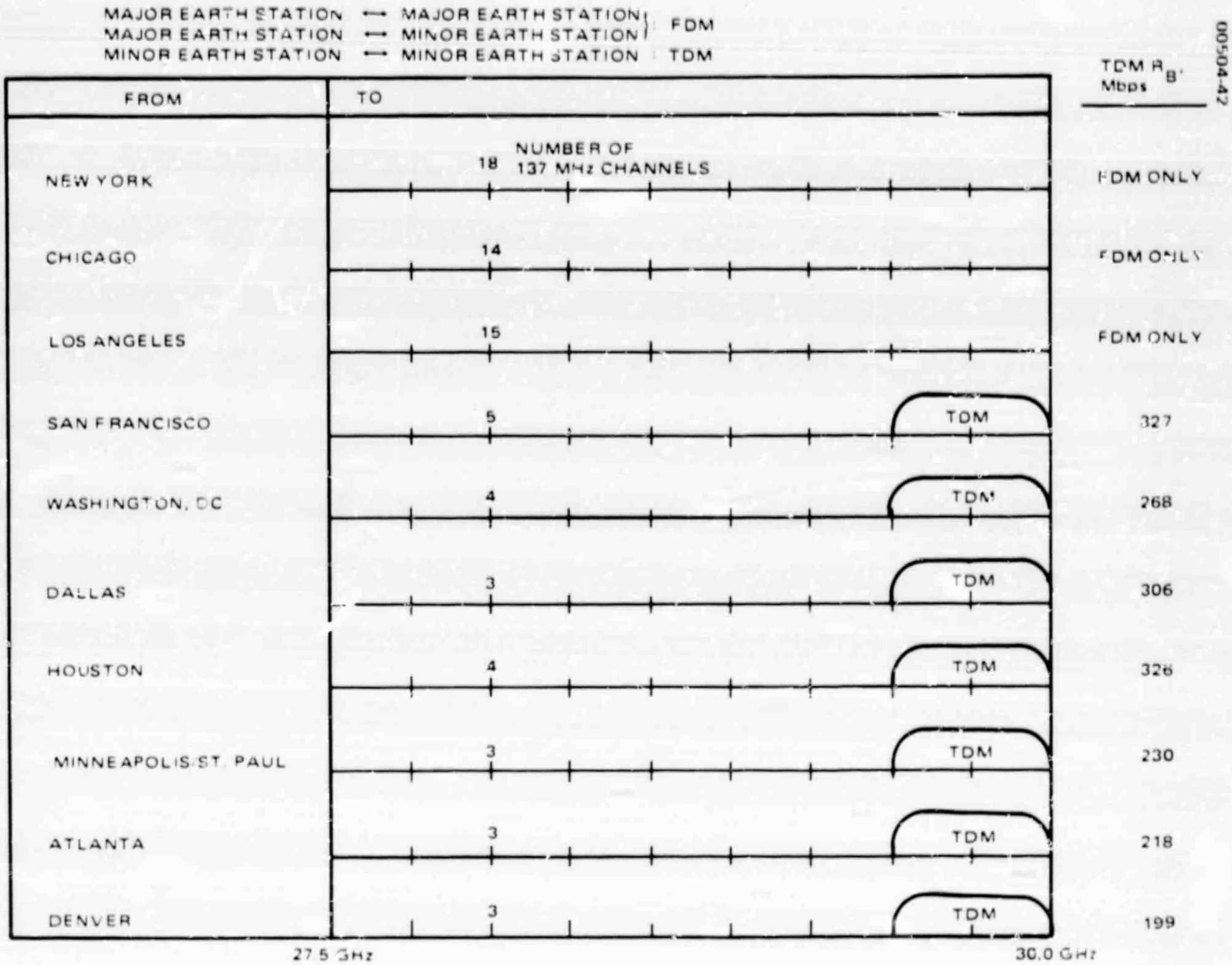
The preferred configuration for the baseline design is to implement the TDM channel filter at the IF. The RF demultiplexing filters that separate the 274 MHz wide FDM channels have the same basic requirements as the FDM input demultiplexing filters of the FDM repeater. The same filter type, a four-pole 0.1 dB ripple Chebyshev filter, will be used in this case and each FDM channel filter will have a 274 MHz equal ripple bandwidth. The IF TDM filter will be a stripline, interdigital, six-pole, 0.1 dB ripple Chebyshev filter with an equal ripple bandwidth of 1096 MHz. This filter choice provides 14 dB attenuation at the band center of the adjacent 274 MHz FDM channel. Total isolation would be 34 dB if a polarization isolation of 20 dB is assumed. Attenuation at the band edge of the nearest FDM channel of the same polarization is 28 dB. Adjacent channel interference should be negligible for both cases.



* EXCESSIVE CHANNEL REQUIREMENTS, NEGLIBLE FDM OFF-LOADING CAPABILITY

- 274 MHz UNIFORM FDM CHANNELIZATION. REQUIRED 4 W SS/HPA
- 548 Mbps TDM BURST RATE. REQUIRED 6 W TWT HPA
- MINIMUM BT_S = N/A
- ALTERNATE CHANNELS ORTHOGONALLY POLARIZED

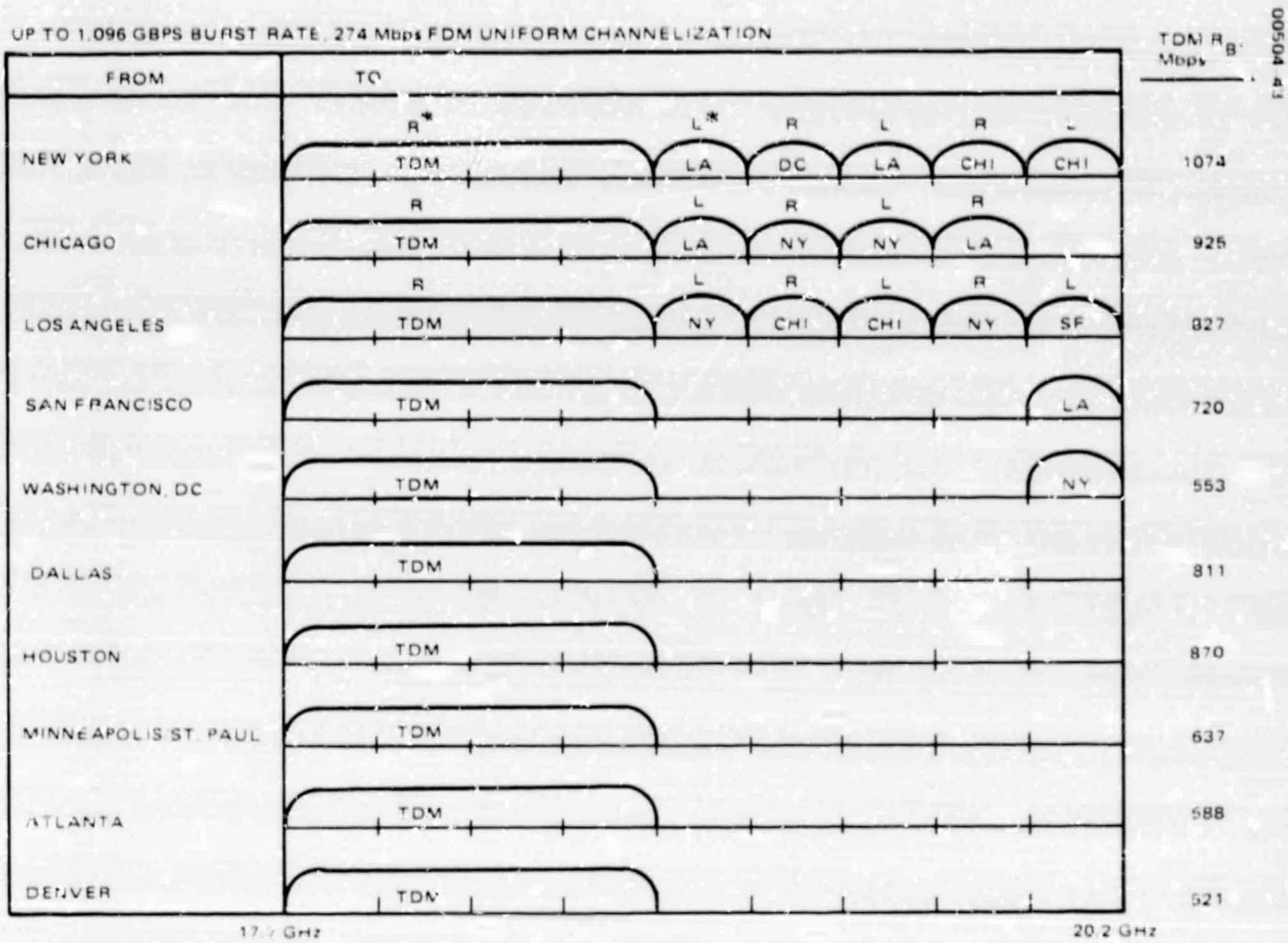
FIGURE 6.2. ALTERNATIVE FDM/TDM CHANNELIZATION, 548 MBPS TDM BURST RATE



- 137 MHz UNIFORM FDM CHANNELIZATION . REQUIRED 2.0 W SS/HPS
- 548 Mbps TDM BURST RATE . REQUIRED 6 W/TWT HPA
- MINIMUM BT_S = 1.40 (NY ⟷ DAL)
- ALTERNATE CHANNELS ORTHOGONALLY POLARIZED

FIGURE 6-3. ALTERNATIVE FDM/TDM PARTITIONING

UP TO 1.096 GBPS BURST RATE, 274 Mbps FDM UNIFORM CHANNELIZATION



*R AND L INDICATE ORTHOGONAL POLARIZATIONS (REQUIRED ON NY, CHI, AND LA)

- 274 MHz UNIFORM FDM CHANNELIZATION REQUIRED 4 W SS HPA
- 1.096 Gbps TDM BURST RATE REQUIRES 10 W TWT HPA
- MINIMUM BT_S = 1.9
- ALTERNATE CHANNELS ORTHOGONALLY POLARIZED

FIGURE 6.4. FDM/TDM REPEATER BASELINE CHANNELIZATION

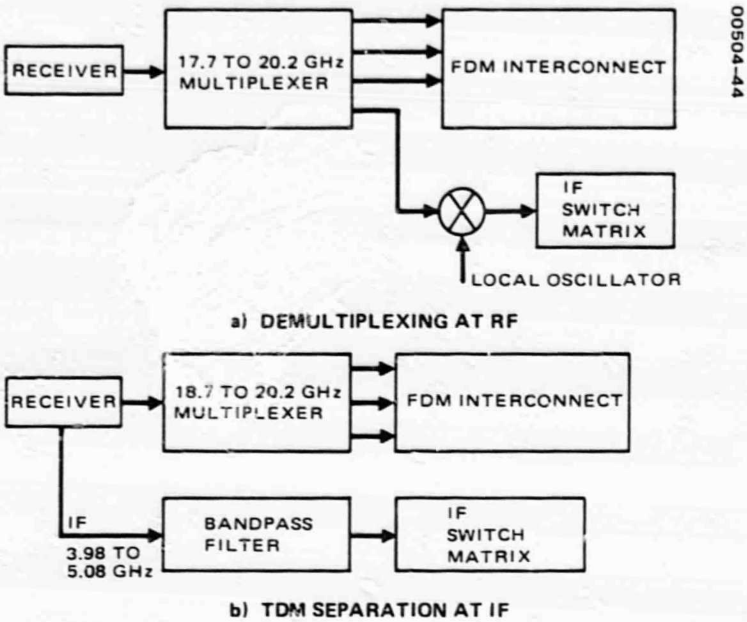


FIGURE 6-5. FDM/TDM SPECTRUM DEMULTIPLEXING ALTERNATES

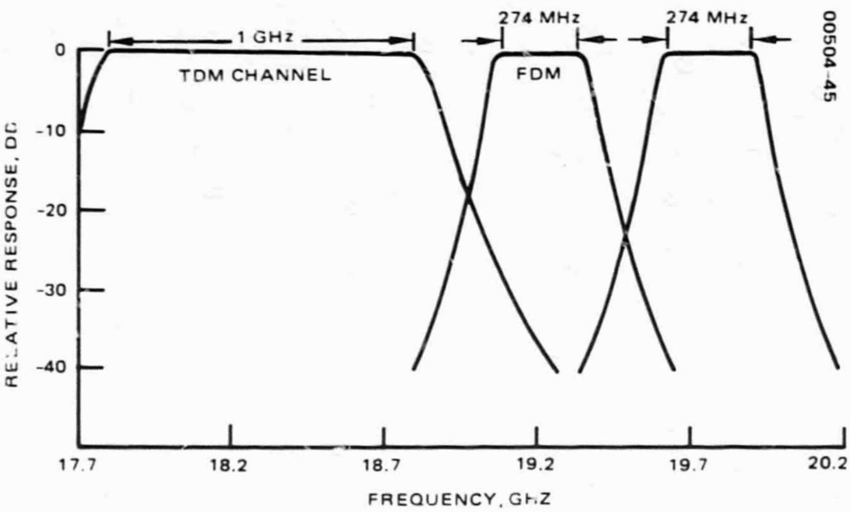


FIGURE 6-6. OUTPUT MULTIPLEXER RESPONSE CHARACTERISTICS

6.1.2 Output Circuit Design Trades

The design trades in the output consist of selection of a transmitter type for the FDM channel and determination of the output multiplexer characteristics. Because the FDM channel power amplifier requirement is a relatively low 4 watts, a solid state power amplifier is a possible selection. TWTA overall efficiency at this power level is not significantly greater than that of a solid state power amplifier. Weight of a solid state amplifier including supply regulators is about 2.0 pound as compared to about 4 pound for a TWTA. The solid state power amplifier has a potential life and reliability advantage over the TWTA. Because of the lower weight and potential reliability advantage the solid state power amplifier has been chosen for the baseline design.

The output multiplexer sums the output of the individual FDM channels and the broadband TDM channel. Bandwidth of each FDM channel is 274 MHz and the TDM channel is 1096 MHz. A six-pole 0.1 dB Chebyshev filter with an equal ripple bandwidth of 1.0 GHz and a 3 dB bandwidth of 1.1 GHz has been chosen for the TDM channel. A four-pole 0.1 dB ripple Chebyshev filter with an equal ripple bandwidth of 274 MHz has been selected for the FDM channel filters.

Figure 6-6 shows the bandpass response characteristics of the TDM channel and the two FDM channels. The filters selected provide adequate rejection of the adjacent copolarized channels. The TDM channel is attenuated 27 dB at the band edge of the adjacent 274 MHz FDM channel and each 274 MHz FDM channel is attenuated more than 35 dB at the band edge of the adjacent FDM channel.

6.1.3 Beacon Transmitter Frequency Selection

The beacon transmitter frequency selection is again based upon providing an adequate beacon to data carrier interference ratio. Since the FDM beacon transmiter which uses a frequency of 19.6 GHz, as discussed in 4.1.5, is the simplest design, it is desirable to use this configuration if possible. The 274 MHz FDM channel ERP is approximately +90 dBm and the beacon carrier ERP is +38 dBm. The 274 MHz channels nearest the beacon transmitter frequency are centered at 19.755 and 19.481 GHz. The discrete spectral line power relative to total data carrier power at 19.6 GHz is -64.2 dBc and -61.9 dBc for these two channels. Beacon carrier to data carrier interference ratio is then 12.2 and 9.9 dB, adequate for carrier recovery. Thus the FDM beacon transmitter configuration can be used for the FDM/TDM repeater.

6.2 FDM/TDM REPEATER BASELINE DESIGN

Figure 6-7 is a block diagram of the baseline TDM/FDM repeater design. Nodes 1 through 3 are identical and have orthogonally polarized uplinks and downlinks. Redundancy is not shown in Figure 6-7. However, 2 for 1 redundancy is provided for the receivers, 16 for 10 redundancy for the TWT TDM transmitters, and 25 for 16 redundancy for the solid state FDM power amplifiers. Implementation for nodes 4 and 5 is identical and each has a TDM and FDM channel. Because the frequency separation of the TDM and FDM channel is quite large, orthogonal polarization is unnecessary for these nodes. Nodes 6 though 10 have only the 1.096 Gbps burst rate TDM channel and do not require any FDM channels.

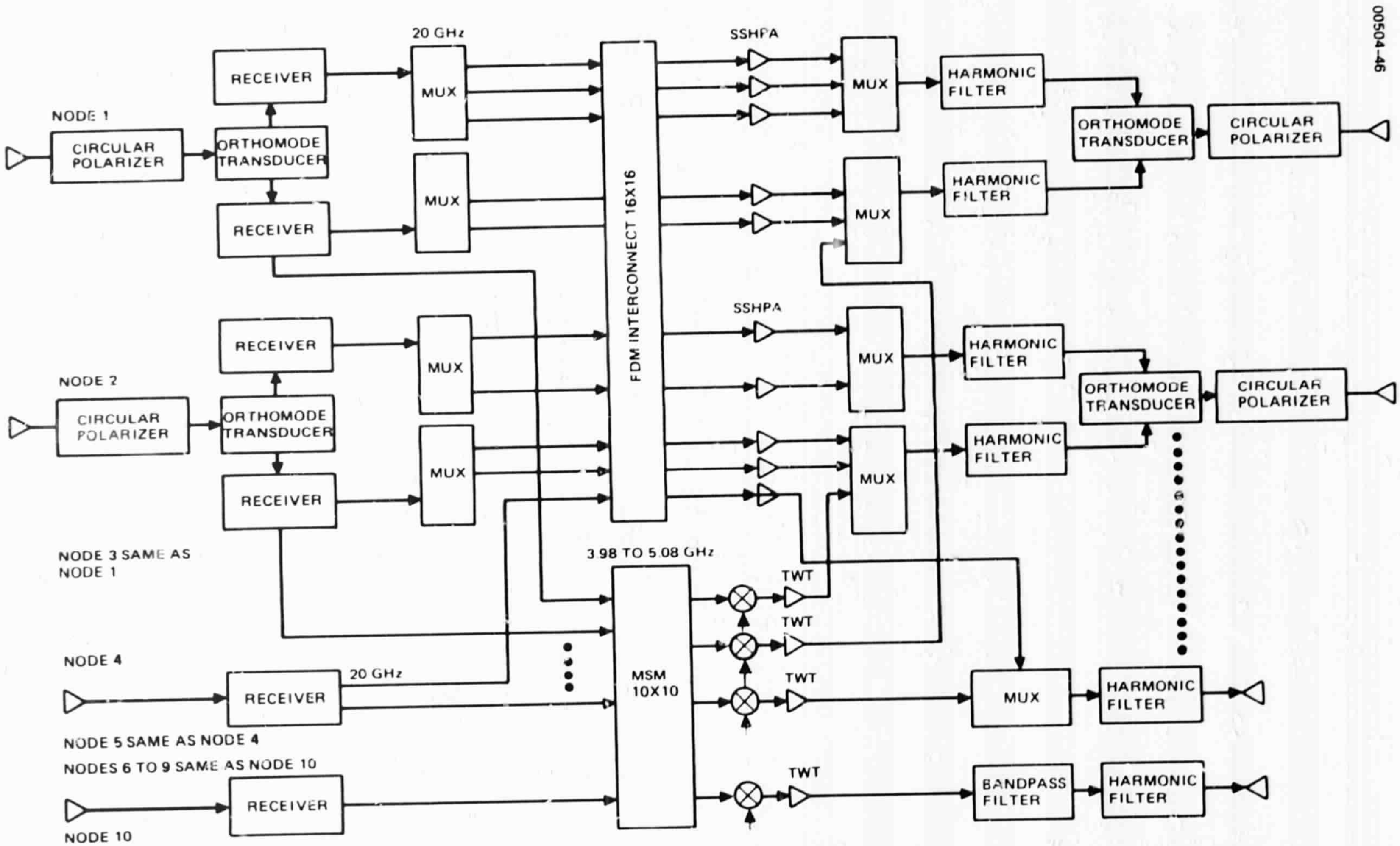


FIGURE 6-7. BASELINE FDM/TDM REPEATER

6.2.1 Input Circuits

The TDM/FDM repeater receiver block diagram is shown in Figure 6-8. The receiver is very similar to the FDM repeater receiver but provides an IF TDM output as well as the FDM output at the repeater output frequency. The TDM IF channel filter is also included in the receiver. A detailed description of the other receiver circuits is given in 3.1 and 4.2.1 of this report.

Total gain of the receiver for the FDM RF output is 49 dB and is 56 dB for the TDM IF output. The FDM bandpass filter following the receiver upconverter is a three-section rectangular waveguide, 0.01 dB ripple Chevyshev filter with an equal ripple bandpass from 18.8 to 20.2 GHz. The TDM IF filter is a stripline interdigital filter 0.1 dB ripple six-pole Chebyshev design with an equal ripple bandpass from 3.98 to 5.08 GHz.

The receiver size, weight and power are tabulated below.

Size	7 by 10 by 2 in.
Weight	3.5 lb
DC Power	5.0 W

6.2.2 IF Switch Matrix

The IF switch matrix design is identical to that described in 5.1.3 and 5.2.2 except that the bandwidth required is reduced from 2.5 GHz to 1.1 GHz.

6.2 3 FDM Output Circuits

The FDM output circuits of the FDM/TDM repeater consist of the switched attenuator, the solid state power amplifier, the power monitor, the output multiplexer, and the harmonic filter. The switched attenuator is identical to the design described previously in 4.2.2.2. The power monitor, except for coupling ratio, is identical to those used in the TDM and FDM repeaters. The harmonic filter is identical to those used in the FDM and TDM repeaters.

A block diagram of the solid state power amplifier is shown in Figure 6-9. The figure also indicates power dissipated per section, the output power level of each section, and the gain of each section of the amplifier. The input section of the amplifier consists of five cascaded GaAs FET amplifiers. The remaining amplifiers are all reflection type, IMPATT diode amplifiers. Each of these amplifiers consists of a circulator, IMPATT diode, matching network, and diode biasing network.

Power mode switching is not shown in Figure 6-9 but is accomplished by using only one of the final eight parallel amplifiers. Power to the other seven amplifiers is switched off in the low power mode and the output of the single amplifier is switched around the eight-way adder.

The amplifiers are constructed in microwave integrated circuits. The eight-way power adder is a stripline radial combiner. The two-way power divider, the two to four-way power divider, and the four to eight-way power divider are all Wilkinson type stripline designs.

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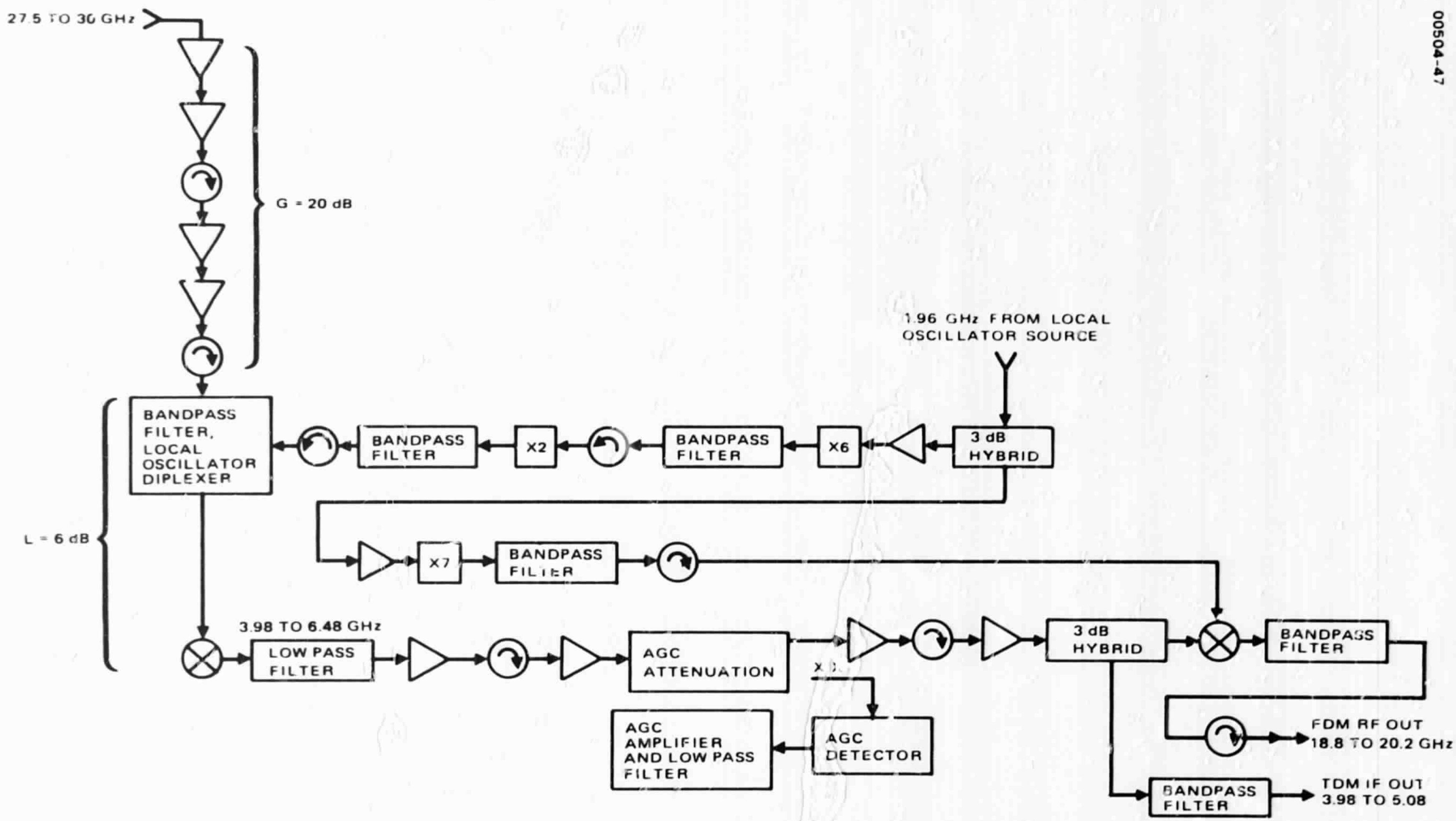

FIGURE 6-8. 30/20/5.2 GHZ FDM/TDM RECEIVER

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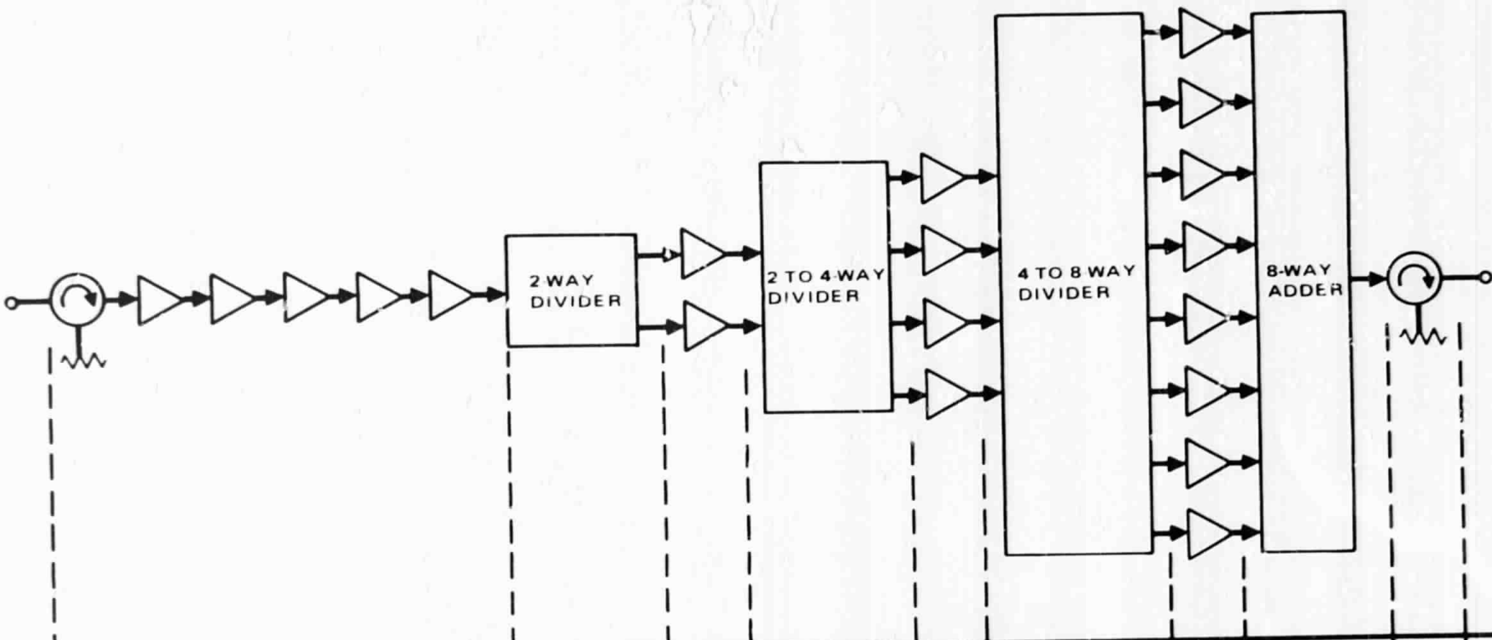

GAIN, dB	+38.25	-0.25	+7.0	-0.25	+5.0	-0.25	+4		-0.25	+5.3
POWER, dBW	-47	-8.75	-9.0	-2.0	-2.25	+2 75	+2.5	+6.5	+6.25	+6.0
* POWER DISSIPATED, W	3.3		3.5	0.04	7.0	0.12	12.8	0.28	0.27	27.3
	COMPONENT VALUES									NET

FIGURE 6-9. 20 GHZ, 4.0 WATT SOLID STATE AMPLIFIER

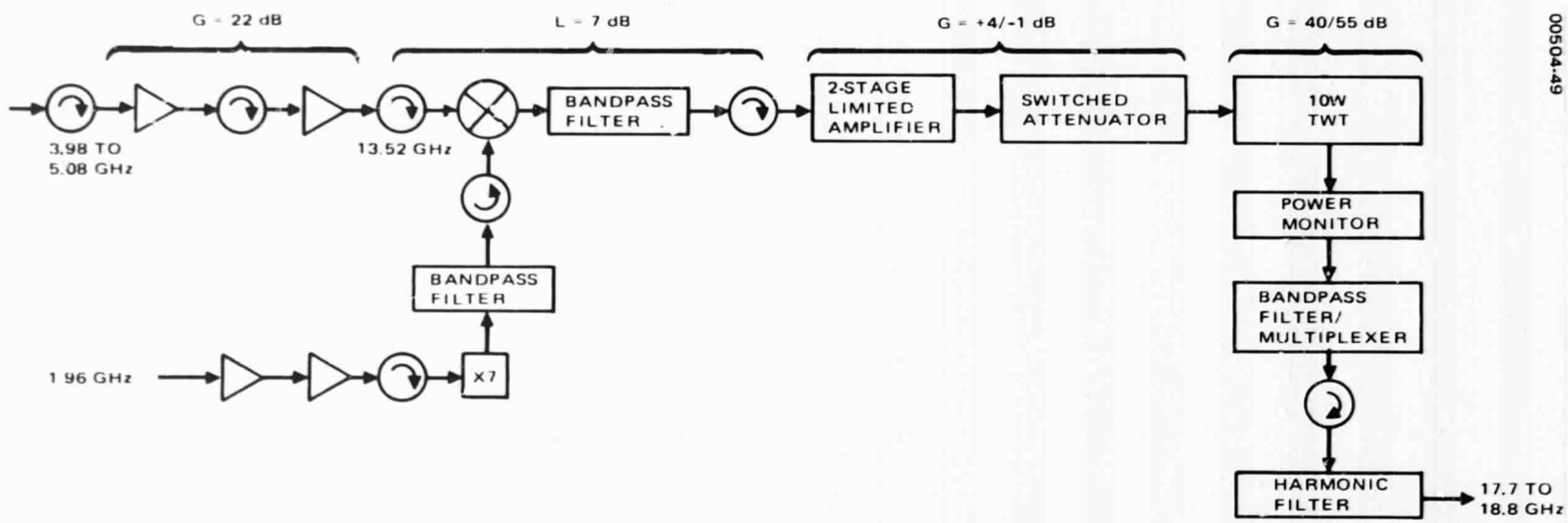

FIGURE 6-10. TDM OUTPUT CIRCUITS

TABLE 6-1. MULTIPLEXER CHARACTERISTICS

	Channel 1 Characteristics		Channel 2 Characteristics		Channel 3 Characteristics	
Multiplexer Type	Bandpass*	Filter Type	Bandpass*	Filter Type	Bandpass*	Filter Type
One TDM, two FDM channels	17.796 to 18.796 GHz	6-Pole 0.1 dB Ripple Chebyshev	19.070 to 19.344 GHz	4-Pole 0.1 dB Ripple Chebyshev	19.618 to 19.892 GHz	4-Pole 0.1 dB Ripple Chebyshev
Three FDM channels	18.796 to 19.070 GHz	4-Pole 0.1 dB Ripple Chebyshev	19.344 to 19.618 GHz	4-Pole 0.1 dB Ripple Chebyshev	19.892 to 20.166 GHz	4-Pole 0.1 dB Ripple Chevyshev
One TDM, one FDM channel	17.796 to 18.796 GHz	6-Pole 0.1 dB Ripple Chebyshev	19.892 to 20.166 GHz	4-Pole 0.1 dB Ripple Chebyshev	NA	NA

*Equal Ripple Bandpass

The output multiplexers consist of three somewhat different types. One is used in applications when two FDM channels and one TDM channel are combined. A second type is used when three FDM channels are combined. These first two multiplexers are used for nodes 1 through 3. A third type is used for nodes 4 and 5 and combines the TDM channel with one FDM channel. The multiplexers are all common waveguide types and the filter resonators are dual mode circular TE_{111} mode resonators. Characteristics of each of the three multiplexers are given in Table 6-1.

6.2.4 TDM Output Circuits

A detailed block diagram of the TWT output circuits is shown in Figure 6-10. All of the output circuits except the IF amplifier, TWT power amplifier, and the output multiplexer or BPF are identical to the TDM repeater output circuits previously described in 5.2.3. The only difference in the IF amplifier is that the gain is 22 dB in this case and is 26 dB in the TDM repeater. The TWTA is identical to the 10 watt TWT of the FDM repeater which is described in 4.2.2.4.

The multiplexer, which is used in nodes 1 through 5, is described in 6.2.3. The characteristics of the TDM channel are given in Table 6-1. The output bandpass filter, which is used only in node 6 through 10, is a six-pole, 0.1 dB ripple Chebyshev filter, and uses three dual-mode circular TE_{111} resonators. Insertion loss of this filter is 0.3 dB.

6.2.5 Beacon Transmitter

The beacon transmitter for the FDM/TDM repeater is identical to the FDM repeater beacon transmitter described in 4.2.2.6.

6.2.6 TDM/FDM Repeater Gain Distribution

The TDM channel and the FDM channel gain distributions are given in Figures 6-11 and 6-12. Both figures show high power mode and low power mode data when appropriate, and the FDM gain distribution of Figure 6-12 also shows the combined FDM channel levels when appropriate.

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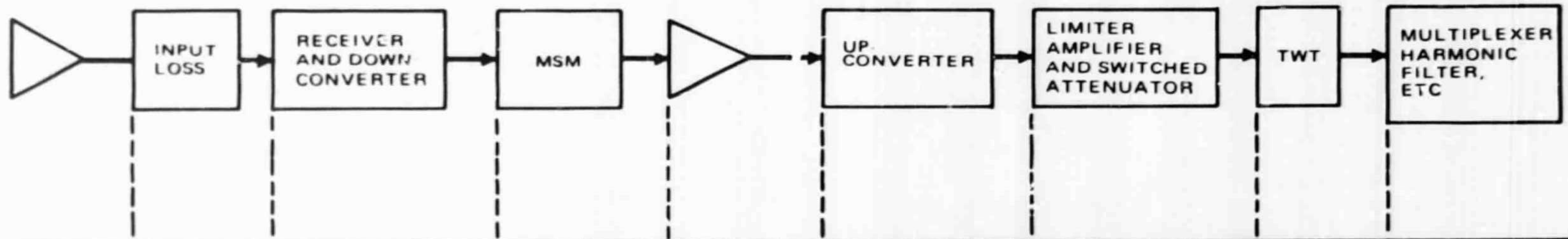

GAIN, dB	-1.3	5.6	-27	22	-7	+4/-1	40/55	-1.8	
LOWPOWER LEVEL, dBm	-56.7	-58.0	-2	-29	-7	-14	-10	+30	+28.2
HIGHPOWER LEVEL, dBm							-15	+40	+38.2

FIGURE 6-11. TDM/FDM REPEATER GAIN DISTRIBUTION (TDM CHANNEL)

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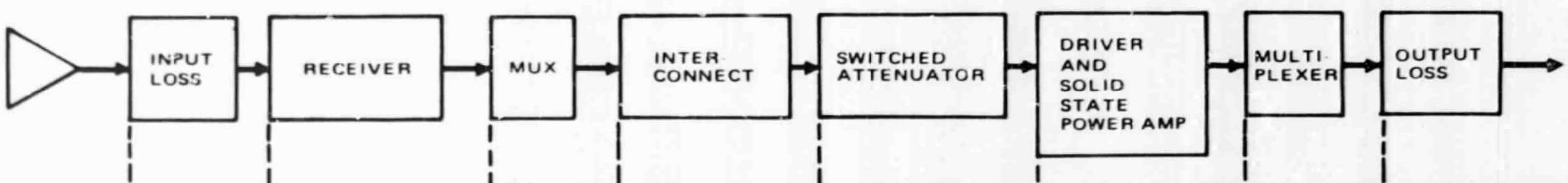

GAIN, dB	-1.3	49	-2	-1	-1.4/ -3.4	45/53	-0.6	-1.4	
LOW POWER LEVEL/CH,	-60.3	-61.6	-12.6	-14.6	-15.6	-19.0	+26	+25.4	+23
HIGH POWER LEVEL/CH,							+36	+35.4	+34
LOW POWER TOTAL LEVEL, dBm	-55.5	-56.8	-7.8					+30.2	+28.8
HIGH POWER TOTAL LEVEL, dBm								+40.2	+38.8

FIGURE 6-12. TDM/FDM REPEATER GAIN DISTRIBUTION (FDM CHANNEL)

6.3 FDM/TDM REPEATER PERFORMANCE

6.3.1 FDM/TDM Repeater Power Output

The power delivered to the feed is a prime performance parameter for the TDM/FDM repeater. Tables 6-2 and 6-3 show the circuit loss budgets and the power delivered to the feed for the 1096 Mbps burst rate TDM channel is 6.6 watts. Power required can be scaled from the statement of work requirement of 3 watts for 548 Mbps data to a requirement of 6 watts. The FDM 274 Mbps channel power output to the antenna feed is 2.5 watts and is comfortably above the statement of work requirement of 1.75 watts.

6.3.2 FDM/TDM Repeater Communication Link Budgets

Tables 6-4 and 6-5 are the uplink and downlink communication power budgets for both the TDM and FDM channels. The uplink E_b/N_o is adequate in both cases. The FDM channel downlink E_b/N_o is 14.61 dB, which allows 4.08 dB for equipment implementation loss. This margin should be adequate. The TDM channel downlink E_b/N_o is 12.84 dB, which allows only 2.31 dB for equipment implementation. This link is considered to be marginal for the maximum rain loss condition.

6.3.3 FDM/TDM Repeater BER Degradation

Simulations were not performed specifically on the TDM/FDM repeater configuration. However, it is possible to estimate the spacecraft segment BER degradation by extrapolating from the TDM repeater simulation results discussed in 5.3.3. For TDM the spacecraft BER degradation was 1.8 dB for 2.5 Gbps QPSK data using an input and output filter 3 dB bandwidth of 2.1 GHz. For the FDM/TDM repeater the TDM channel 3 dB bandwidth is equal to the 1.096 Gbps data rate. For the FDM

TABLE 6-2. FDM/TDM OUTPUT CIRCUIT LOSS AND POWER TO FEED – TDM CASE

Output Loss Budget

Output Loss Element	Loss, dB	
	Nominal	Spare Amplifier Used
Power monitor	0.1	
Switch	0.2	0.5
Multiplexer	0.4	
Isolator	0.2	
Harmonic filter	0.2	
Waveguide, 6 ft	0.7	
Total	1.8	2.1

Power delivered to feed = 8.2 dBW, 6.6 W

TABLE 6-3. FDM/TDM OUTPUT CIRCUIT LOSS AND POWER TO FEED – FDM CASE

Output Loss Budget

Output Loss Element	Loss, dB	
	Nominal	Spare Amplifier Used
Power monitor	0.1	
Switch	0.2	0.5
Multiplexer	0.6	
Isolator	0.2	
Harmonic filter	0.2	
Waveguide, 6 ft	0.7	
Total	2.0	2.3

Power delivered to feed = +4.0 dBW, 2.5 W

TABLE 6-4. FDM/TDM REPEATER UPLINK BUDGET

Uplink frequency = 30 GHz Bit rate M (No. bits/symbol) = 2	TDM 1096 Mbps	FDM 274 Mbps
Terminal transmitter power dBW	25.55	24.28
Antenna gain, dB (40 ft dia, 0.06° HPBW)	68.98	68.98
Feed loss, dB	-2.0	-3.75
EIRP, dBW	92.53	89.51
Loss-terminal antenna pointing error, dB (0.01°)	-0.29	-0.29
Margin, dB	-1.00	-1.00
System aging effects, dB	-1.00	-1.00
Random variation of elements, dB	-1.50	-1.50
Rain loss, dB (0.125% outage, CCIR rainfall region 4)	-20.03	-20.03
Polarization loss, dB	-0.25	-0.25
Atmospheric loss, dB	-0.59	-0.59
Propagation loss, dB (latitude = 47.5°, relative longitude = 27.5°)	-213.73	-213.73
Beam edge loss, dB	-0.00	-0.00
Spacecraft stationkeeping loss, dB	-0.00	-0.00
Spacecraft antenna pointing error, dB (0.05°)	-0.92	-0.92
Spacecraft antenna gain, dB (14 ft diameter, 0.17° HPBW)	58.80	58.80
Spacecraft received carrier power dBW	-87.98	-91.00
Received noise power density, dBW/Hz (T (R) = 530°, T (E) = 700°)	-200.15	-200.15
Bandwidth, dB (Hz) BT = 2.00)	90.40	84.38
Spacecraft receiver noise power, dBW	-109.75	-115.77
Uplink carrier-to-noise power ratio, dB E_b/N_o, dB	21.80	24.77

TABLE 6-5. FDM/TDM REPEATER DOWNLINK BUDGET

Downlink frequency = 20,000 GHz	TDM	FDM
Bit rate	1096 Mbps	274 Mbps
M (No. bits/symbol) = 2		
Spacecraft output power dBW	8.20	4.0
Antenna gain, dB (14 ft diameter, 0.25° HPBW)	56.34	56.34
Feed loss, dB	-1.80	-2.00
EIRP, dBW	62.74	58.34
Loss-spacecraft antenna pointing error, dB (0.05°)	-0.39	-0.39
Spacecraft stationkeeping loss, dB	-0.00	-0.00
Margin, dB	-1.00	-1.00
System aging effects, dB	-1.00	-1.00
Random variation of elements, dB	-1.50	-1.50
Rain loss, dB (0.073% outage, CCIR Rainfall Region 4)	-10.03	-10.03
Polarization loss, dB	-0.25	-0.25
Atmospheric loss, dB	-0.70	-0.70
Propagation loss, dB (latitude = 47.5, relative longitude = 27.5)	-210.21	-210.21
Beam edge loss, dB	-0.00	-0.00
Terminal antenna pointing error, dB (0.01°)	-0.13	-0.13
Terminal antenna feed loss, dB	-1.50	-1.50
Terminal antenna gain, dB (40,000 ft diameter, 0.09° HPBW)	65.46	65.46
Terminal received carrier power, dBW	-98.51	-102.91
Received noise power density, dBW/Hz (T (R) = 150°, T (E) = 423°)	-202.34	-202.34
Bandwidth, dB (Hz) BT = 2.00)	90.40	84.38
Uplink noise contribution, dB	0.59	0.44
Terminal receiver noise power, dBW	-111.35	-117.52
Link carrier-to-noise power ratio, dB	12.84	14.61
Hardware implementation loss, dB	2.31	4.08
E_b/N_o, (dB) (BER = 1×10^{-6})	10.53	10.53

TABLE 6-6. FDM/TDM REPEATER WEIGHT AND POWER

Component		Quantity (operating plus redundant)	Weight, lb		Power, Watts	
			Unit	Total	Unit	Total
Redundancy switches (30 GHz)		2[illegible]	0.7	18.2	–	–
Receivers		13 + 13	3.5	91	5	65
Upconverters		10 + 6	0.8	12.8	0.15	1.5
Local oscillator redundancy switches		23	0.3	6.9	–	–
Input multiplexer		6	0.51	3.1	–	–
10 x 10 switch		1	4.0	4.0	3.5	3.5
Switch control unit		1 + 1	6.0	12.0	6.0	6.0
IF interconnect		(16 x 16)	–	2.0	–	–
High power amplifier	TDM: TWT 10 W 25% effective	10 + 6	4.5	72	40/8	272*
	FDM: SS 4 W 12% effective	16 + 9	2.0	50	33.3	335*
Redundancy switches, 20 GHz		51	0.8	40.8	–	–
Output multiplexer		8	0.51	4.1	–	–
Local oscillator source		1 + 1	4.0	8	10.0	20.0
Beacon transmitter		1 + 1	1.0	2.0	4.0	4.0
Miscellaneous repeater		–	–	30	–	5
Total				357		712

*Six TWTAs in high power mode
**Nine solid-state power amplifiers in high power mode

channel the 3 dB bandwidth is substantially greater than the QPSK data rate. Based on this comparison, the TDM channel spacecraft segment BER degradation and the FDM channel degradation should both be less than 1.8 dB.

6.3.4 FDM/TDM Repeater Power and Weight

Table 6-6 gives a breakdown of the estimated FDM/TDM repeater power and weight by component. The power required is based upon 6 of 10 TWTs operating in the high power mode and 9 of 16 solid state power amplifiers operating in the high power mode.

6.4 TDM/FDM REPEATER COMMAND AND TELEMETRY REQUIREMENTS

The command requirements for the TDM/FDM repeater are as follows:

Function	No. of Commands
Receiver select	20
Transmitter select/power ON/OFF	164
Local oscillator source/select	29
Transmitter power level	41
Beacon transmitter select	2

The telemetry requirements of the FDM/TDM repeater are as follows:

Function	Signal Type	No. Required
Received power level monitor	Analog	13
Transmitter power level monitor	Analog	26
Transmitter switch status	Digital-2 bits	41
Transmitter power status	Discrete	41
Receiver selection and power	Discrete	26
TWTA parameters	Analog	48
Secondary voltages	Analog	30
Temperature monitors	Analog	10

7. REPEATER COMPARISON

This section compares the baseline designs of the three repeater types in terms of performance, weight, power, cost, and critical technology required.

7.1 PERFORMANCE COMPARISON

The primary performance parameters to be compared are the power delivered to the feed scaled to data rate and the BER degradation of the spacecraft segment at a BER of 10^{-6}.

Table 7-1 compares these parameters for the three repeater types. The normalized power output to data rate ratio column is derived by dividing the power delivered to the feed by the ratio of the applicable data rate divided by 137 Mbps. An exactly equal comparison of the three repeater types would require that each repeater have the same normalized power output to data rate ratio. The only signficant variation from unity is in the FDM/TDM case where the normalized ratios suggest that the FDM channel solid state power amplifier should be reduced to 4.0/1.25 or 3.2 watts and the TWT TDM transmitter output increased to 10/0.83 or 12 watts. If these changes are made in the TDM/FDM repeater, the change in the repeater power required is a decrease of less than 2 percent. The variations in the normalized power out to data rate ratio therefore do not significantly change the results of the study since a total variation of ± 5 percent probably is less than the uncertainty of the estimates. The TDM and the FDM/TDM repeaters both have somewhat better BER degradation performance than the FDM repeater. This difference is due to the intermodulation products generated in the common three-channel amplifier of the FDM repeater. The FDM repeater performance could be improved by increasing the backoff of the TWT, but this would cause a considerable increase in the power requirement of the FDM repeater.

One other significant performance parameter is flexibility of use of a repeater. The TDM repeater and the FDM/TDM repeater both can have the capability of reconfiguring terminal to terminal data rates by altering the IF switch matrix control program stored in the switch control unit memory. The FDM repeater can provide this function with fewer choices only by the addition of switches in the interconnect and by subdivision of channels.

7.2 WEIGHT, POWER, AND COST COMPARISON

Table 7-2 shows the estimated weight, power, and cost for the three repeaters. It should be emphasized that all of the data shown in Table 7-2 is subject to considerable

TABLE 7-1. REPEATER PERFORMANCE COMPARISON

Repeater Configuration	BER Degradation, dB (10^{-6} BER)	Normalized Power Output to Data Rate Ratio (1.0 = 1 W/137 Mbps)
FDM	2 to 3	1.05
TDM	1.8	0.95
FDM/TDM		
FDM channel	<1.8	1.25
TDM channel	<1.8	0.83

TABLE 7-2. REPEATER WEIGHT, POWER, AND COST COMPARISON

Repeater Type	Weight, lb	Power, W	Cost, $M		
			Nonrecurring	Recurring, 1 Unit	Total
FDM	587	1204	8.1	9.8	17.9
TDM	241	651	7.7	6.2	13.9
FDM/TDM	357	712	10.0	7.8	17.8

uncertainty in an absolute sense. However, since the same methods were used to generate this data for each repeater, the data is valid in a relative sense.

The data of Table 7-2 indicates a clear superiority of the TDM repeater. However, practical realization of a system having a TDM repeater using a very high burst data rate depends upon the availability of very high burst rate modems which currently do not exist and may present formidable development problems. If the burst data rate of the TDM repeater is reduced to a value that is a more reasonable extrapolation of present technology such as the range of 250 to 500 Mbps, then the complexity, weight, power, and cost of the TDM repeater would be very comparable to that of the hybrid FDM/TDM repeater.

The FDM repeater concept does not appear very competitive in this study principally because of the use of TWT backoff which nearly doubles the power requirement of this repeater, and also because of the use of uniform channelization which was adopted to avoid the use of more than one power amplifier type. The development of more efficient, lightweight solid state transmitters having power outputs of say 2, 4, and 8 watts would make a substantial difference in the FDM amplifier design. This development would permit the use of nonuniform channelization having data rates of 137, 274, and 548 Mbps and would permit the use of a power amplifier for each FDM channel. These two design changes would reduce the FDM repeater weight, power, complexity, and cost by a substantial amount.

7.3 CRITICAL TECHNOLOGY

Improvement of solid state power amplifiers and development of very high data rate burst modems appear to be the two most critical technologies. The development of high efficiency lightweight solid state power amplifiers having power outputs in the range of 2 to 25 watts would be of great benefit to satellite repeater designs. The development of burst modems capable of operating at burst data rates of 500 Mbps and higher would make it possible to realize the potential advantages of TDM techniques in satellite repeater design.