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A Multi-Beam SHF Satellite Communications System for Canada (1977-1985)

October 1972

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Satellite Communications System
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Final Report

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ABSTRACT

This study examines satellite systems employing spot beam antennas and frequency reuse in the 12 to 14 GHz bands, that will serve Canada's major traffic and remote communications needs in the time frame 1977 through 1985. Systems with on-board time division switching and without on-board switching are investigated.

For the major traffic service, the study concludes that a satellite with four medium size beams (0.7° × 1.7°) illuminating the dense traffic regions, without on-board switching, and using time division multiple access, would most effectively and economically serve Canada's requirements. For the remote service a system with five regional coverage antenna beams (1.5° × 2.0°), using speech compressed single-channel-per-carrier frequency modulation and frequency division multiple access, would be the preferred choice.

A satellite providing the above two services would weigh about 750 pounds, which is within the launch capability of the Thor-Delta vehicle.

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1. INTRODUCTION

The background and objectives of the study are stated.

In 1971 a study was conducted by Bell-Northern Research, Ottawa, under a contract from the Department of Communications, Government of Canada, to evaluate the significance to Canada of satellite communications in the frequency range of 12 through 18 GHz in the time frame 1977 through 1985. The services examined included commercial and educational television, major route telephone and data transmission, and remote communications. This study was submitted to the Department of Communications in a report entitled "A Super-High-Frequency Satellite Communications System for Canada (1977-1985)".

Specifically, in the areas of major route and remote communications, the forecast traffic demands could be met within the bandwidth available without resort to frequency reuse or other techniques of spectrum utilization. Hence, the use of spot beam antennas on board the satellite was not considered a necessity. Nevertheless, under "Conclusions and Recommendations" of Volume I of the report, it was pointed out that the feasibility and desirability of spot beam application to the Canadian environment should form part of further investigations.

In April 1972, the Department of Communications requisitioned consulting services from Bell-Northern Research, Ottawa, to carry out a study involving satellite spot beam technology. The objectives of this study are detailed in 1.1.

* * *

1. Introduction

1.1 SPECIFIC OBJECTIVES OF STUDY

The objectives of the study are to evaluate the technical feasibility and overall costs of satellite systems using spot beam technology.

In the Statement of Work which forms part of the present contract, the objectives of the study are stated as follows:-

> "For the purpose of obtaining further data for planning purposes, it is desired to investigate the costs and broad system concepts of 12 and 14 GHz satellite communications systems applied to the Canadian environment which exploit multiple spot beam antenna technology. The tasks to be performed are:
>
> i) synthesize and model two satellite systems using spot beams, one without and one with on-board signal processing. Both systems are to be capable of providing a communications connection between any two beams. The first system must be transparent to the modulation technique used so that this may be either analog or digital, with the technique providing an optimum system to be chosen for the model. The second system may be restricted to digital modulation and should consider time division multiple access and on-board switching with time division multiplexing among beams. Critical system concepts should be examined in broad terms to establish feasibility.
>
> ii) perform cost analyses for the systems in terms of capital investment and annual charges.
>
> iii) prepare a report.
>
> The approach taken to the above tasks in the previous study should, in general, be continued."

The depth of the study was confined to a resource allocation of 160 man-days spread over approximately a four month period. This resource limitation was considered reasonable and is consistent with the depth of the investigation desired by the Department of Communications as expressed in the following extract from the Statement of Work:

> "To preclude also the necessity to examine in detail at this time all the techniques which are required to realize the system, only those which are critical to the system concept should be examined and then only to sufficient depth to establish feasibility."

For the purpose of the study, the following baseline assumptions are defined:

> "1) The satellite will employ antennas producing spot beam coverage of the major population areas of Canada.
>
> 2) Both major and remote communications traffic, as defined in the (previous) study, will be carried. For major traffic a required space system capacity equivalent to 10 000 voice channels should be assumed. For remote traffic the previously defined model can be assumed.
>
> 3) The anticipated remainder of the satellite capacity (assuming a Thor-Delta launch vehicle) will be devoted to other services which can also take advantage of the narrow beams. These will not be considered in this analysis. However, the portion of the satellite used for the communications traffic, in terms of weight and prime power, should be estimated."

The Design Authority to whom regular progress reports of the study were made, is the Director-General, Technological and Systems Planning, Department of Communications.

This report is the property of the Department of Communications, Government of Canada.

2. SYSTEMS CONSTRAINTS

This chapter will present the traffic and technical systems constraints and the performance criteria.

In any communications satellite system evaluation, the necessary prerequisites are clear definitions of systems constraints and performance criteria. First and foremost are the forecast traffic needs, both in terms of channel capacity and in the number and locations of earth terminals. Then, in order for a system to be technically feasible and economically viable, there is a need to define the anticipated satellite technology, the preferred type of launch vehicle and its projected payload capability. Finally, it is essential to establish performance criteria for each service so that system technical trade-off can be carried out at a later stage.

In the sections that follow, the above-mentioned aspects are given consideration to the depth appropriate to this study. Much of the data will be drawn from the report [14] on the previous study.

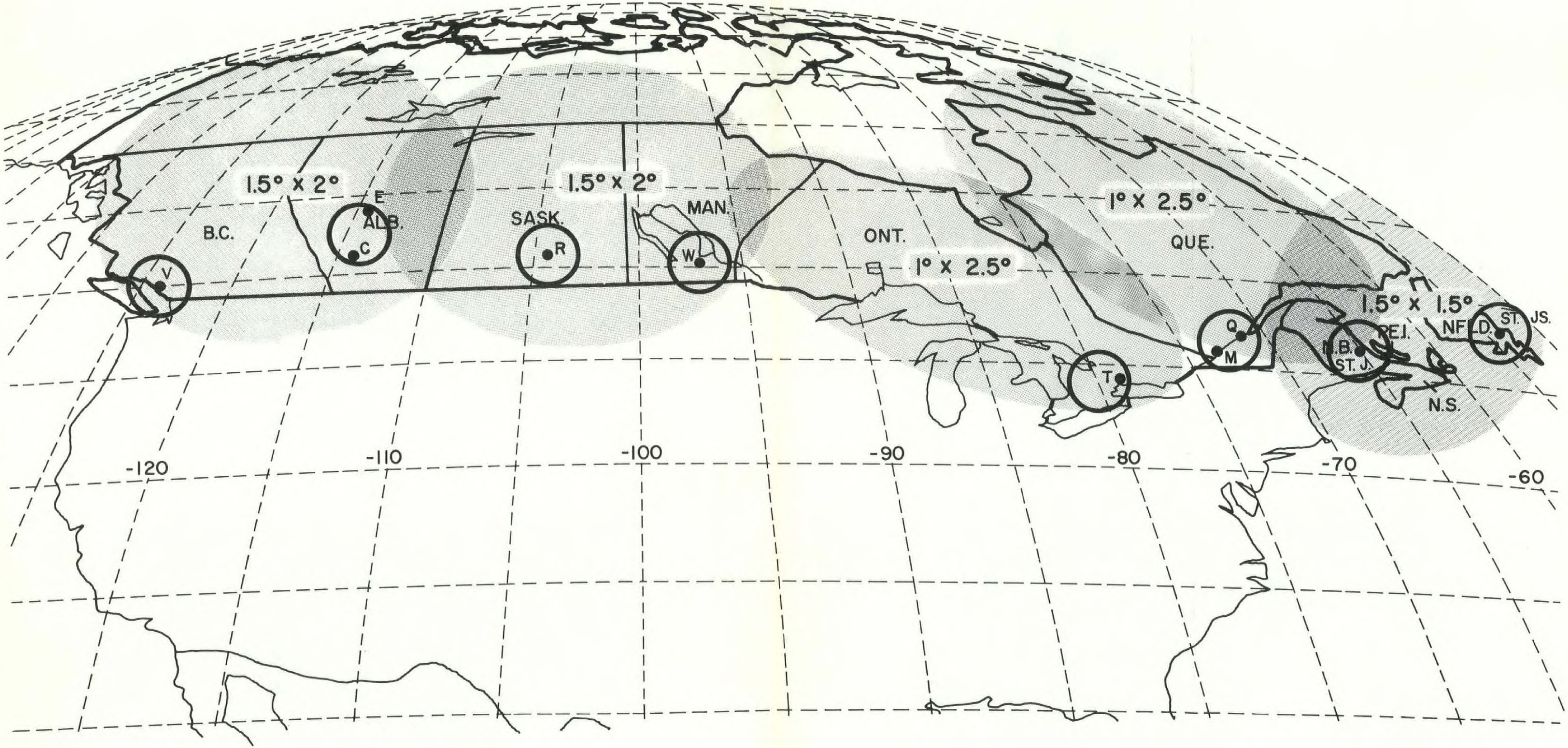

Figure 3-2
Antenna Beam Plan (A)
[8 Spot Beams, 5 Medium Beams]

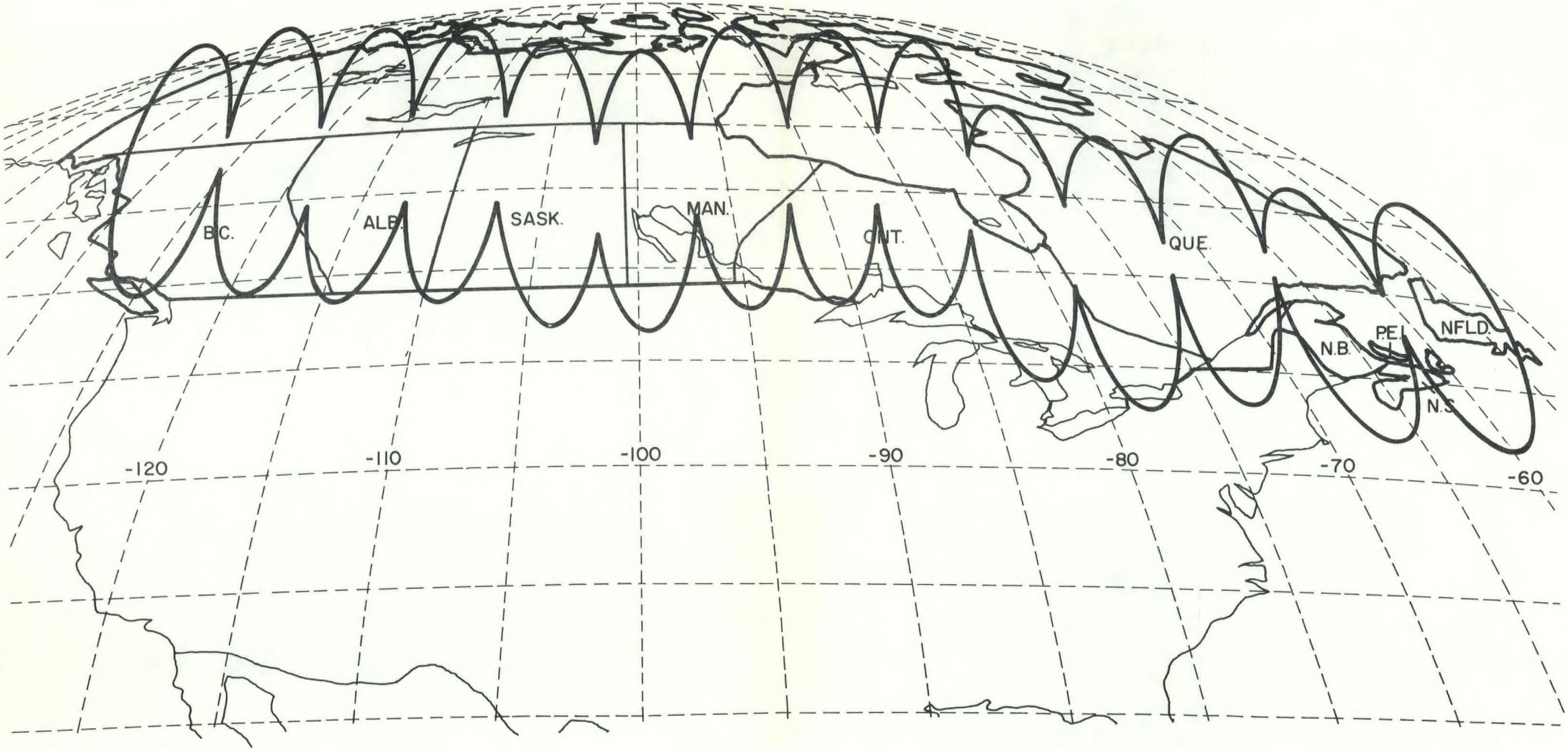

Figure 3-3
Antenna Beam Plan (B)
[14 Uniform Elliptical Beams]

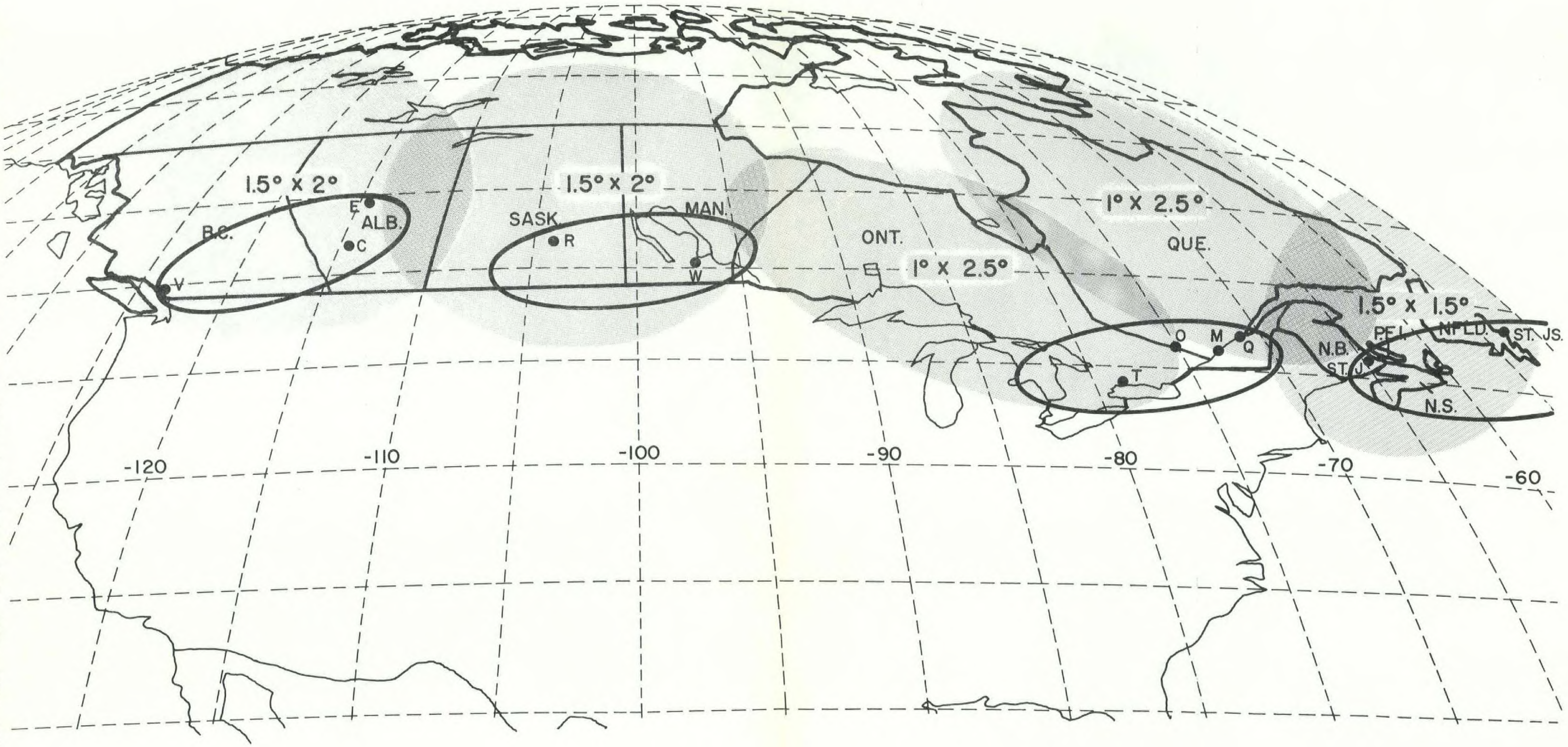

Figure 3-4
Antenna Beam Plan (C)
[4 Beams, Elliptical Shape]

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2. Systems Constraints

2.1 TRAFFIC CONSTRAINTS

The forecast traffic requirements for major route and for remote communications, in terms of channel capacities and geographical distribution are presented.

In the previous study [14] a survey was carried out on the forecast inter-Provincial telephone traffic needs of Canada in the time frame 1977 through 1985. This total requirement for channels more than 1000 miles long for the year 1985 corresponds to about 44 000 one-way voice channels. In that study it was reasoned, with justification, that not all traffic would be routed via satellite. It was considered that 15 percent of the traffic via satellite was appropriate. Consequently, the satellite system would need to provide about 6000 channels. In the satellite models considered in that study, the actual capacity used was 3000 channels, it being assumed that the Anik I system or its replacement in the 4 to 6 GHz band would take approximately an equal share of the traffic.

In the present study, the Design Authority has indicated that all the major or inter-Provincial traffic via satellite would be assumed to be routed only via the 12 to 14 GHz system under current evaluation. Further it is anticipated that channels of less than 1000 miles in length might be economically carried via satellite. In round numbers, the total capacity of the satellite system under this study is given by the Design Authority as 10 000 channels. This is considered to be reasonable.

In modeling any realistic satellite system to serve Canada, it is essential to know not only the total capacity needs, but also the geographical distribution of those needs. This is because of the fact that the coverage zone of the satellite spot beam and the total bandwidth assigned to each spot beam is governed by the capacity requirements from regions with the densest traffic demands. The system design naturally must satisfy this peak demand. Again, based on data presented in the previous report [14], it is possible to construct a traffic table showing the distribution of the 10 000 channels among the 8 major traffic centers in Canada. This traffic distribution is given in Table 2-1. It should be noted that the above centers are centers of inter-Provincial major traffic, and are not necessarily coincident with centers of dense local or intra-Provincial traffic.

Although eight centers are identified, in the system models considered in the subsequent chapters of this report a total of 12 earth terminals will be used for the major route traffic system models. This is in line with the approach taken previously [14]. The additional four centers might be Edmonton, Ottawa, Quebec City and Halifax.

Table 2-1 Distribution of 10 000 Channels of Satellite Traffic

TRAFFIC CENTERS	ONE-WAY CHANNELS
Vancouver	1 750
Calgary	500
Regina	1 100
Winnipeg	1 360
Toronto	2 450
Montreal	1 500
Moncton	630
St. Johns, Nfld.	710
TOTAL	10 000

In the case of remote communications the previous study [14] indicated a total requirement of about 1200 channels. The number of earth terminals is given as 400 with an average of three channels per terminal. The actual locations of these earth stations are not known and for the purpose of system modelling it will be assumed that most of the terminals will be spread out in the northern belt of Canada. From the point of view of community of interest it is reasonable to expect that most of the channels from any particular remote station would be either directly to a southern center of population or to nearby northern remote communities. This assumed pattern of traffic distribution has bearing on the spot beam and transponder configurations in the models considered in later chapters of this report.

2.2 TECHNICAL CONSTRAINTS

The technical constraints are defined in terms of launch vehicle type and capability, and satellite and earth segment technology as anticipated for the time frame 1977 through 1985.

In chapter 1 it was noted that the Design Authority has specified the use of the Thor-Delta as the preferred launch vehicle. This is the same vehicle considered in the previous study. This implies a total launch weight capability of about 1000 pounds (in orbit). This launch vehicle constraint is reasonable and enables direct cost comparison with the models evaluated in the previous study. In the present system evaluation, if the total satellite weight attributed to the major traffic and remote communications subsystems turns out to be under 1000 pounds, then it will be assumed that other services such as commercial or educational television or both, would be added to the satellite system to the extent that the balance of the weight permits. No system analysis of these television services will be carried out in this report. However, in the consideration of beam coverage and frequency plans and the effects of these on mutual interference, these television facilities are assumed to co-exist in the same satellite and their need for interference protection is taken into account.

The satellite technology anticipated in the 1977 through 1985 time frame has been extensively covered in the previous study [14]. Briefly, the technology is expected to be an extension of the present state of the art and would have the following additional features not found in present day operational or planned communications satellites:

a) Three-axis stabilization

b) Unfurlable reflector antennas for multiple spot beams

c) On-board signal time division switching

Solar arrays with improved efficiencies (in watts per square foot) will provide the prime power. Travelling wave tubes of medium powers (up to 20 watts) will be the main method of power amplification for the down links. The rest of the communications subsystems in the satellite will be solid state.

The satellite life is expected to be about seven years, with position maintained to within ± 0.1° in longitudinal drift and ± 0.1° in orbital inclination. Further, the satellite attitude control is expected to be maintained to within ± 0.1° in all three axes. This attitude stability, which affects satellite spot beam pointing accuracy and beam service (coverage) zone, is a critical requirement for the very narrow beams.

In line with the practice adopted by Telesat and some of the U.S. domestic satellite systems now in the process of being approved by the U.S. Federal Communications Commission, a total of three satellites would be assumed to be procured for the system under study. One would be operational, one would be an in-orbit standby and one would be a spare on the ground. The cost of launching the spare satellite will however be included in the cost analysis. This approach is considered realistic and in line with the system reliability expected of a major communications network.

The technical constraints that apply to the earth segment have also been covered extensively in the previous study. These same constraints apply in the systems under study here. The main constraints are the minimum size of antenna (10 feet) due to considerations of interference to and from neighboring satellite systems, maximum size of antenna (60 feet) from considerations of pointing accuracy and manufacture tolerance of reflector surface, and maximum power of the klystron or TWT amplifiers (8 kW). These constraints limit the range of earth segment parameters that are practical. In general these constraints act as broad system boundaries and the optimum systems that emerge, as shown in later chapters of this report, do fall within these boundaries.

2.3 INTER-RELATIONSHIPS OF PERFORMANCE CRITERIA, POWER, BANDWIDTH AND INTERFERENCE

The performance criteria and the effect of interference on power and bandwidth requirements are presented.

In this study,which is concerned with major route telephone communications and remote communications, the following types of modulation techniques will be considered:

a) Pulse Code Modulation with Phase Shift Keying (PCM-PSK).

b) Frequency Division Multiplex with Frequency Modulation with and without speech compression (lincompex).

c) Single-Channel-per-Carrier FM with analog speech compression (lincompex).

The justification of the particular modulation technique for the satellite models concerned will be given in later chapters. For each type of modulation, it is necessary to define the performance criteria and the effects of interference on the requirements of bandwidth and power.

Although television services will not be evaluated in this study, on account of the application of spot beam technology and frequency reuse, the interference criteria associated with the various television services need to be considered. FM is assumed to be used for television.

The paragraphs that follow will consider each of these modulation schemes.

PCM-PSK Telephony

In a PCM-PSK telephone transmission system, the well accepted performance criterion is the threshold bit error rate of 1×10^{-4}. This, according to reference 20, would give a noise of about 50 000 pwp in a telephone channel at a point of zero dBm reference level. Under normal operating conditions with appropriate fade margins, the bit error rate would be better than 1×10^{-7}, and this would introduce negligible noise.

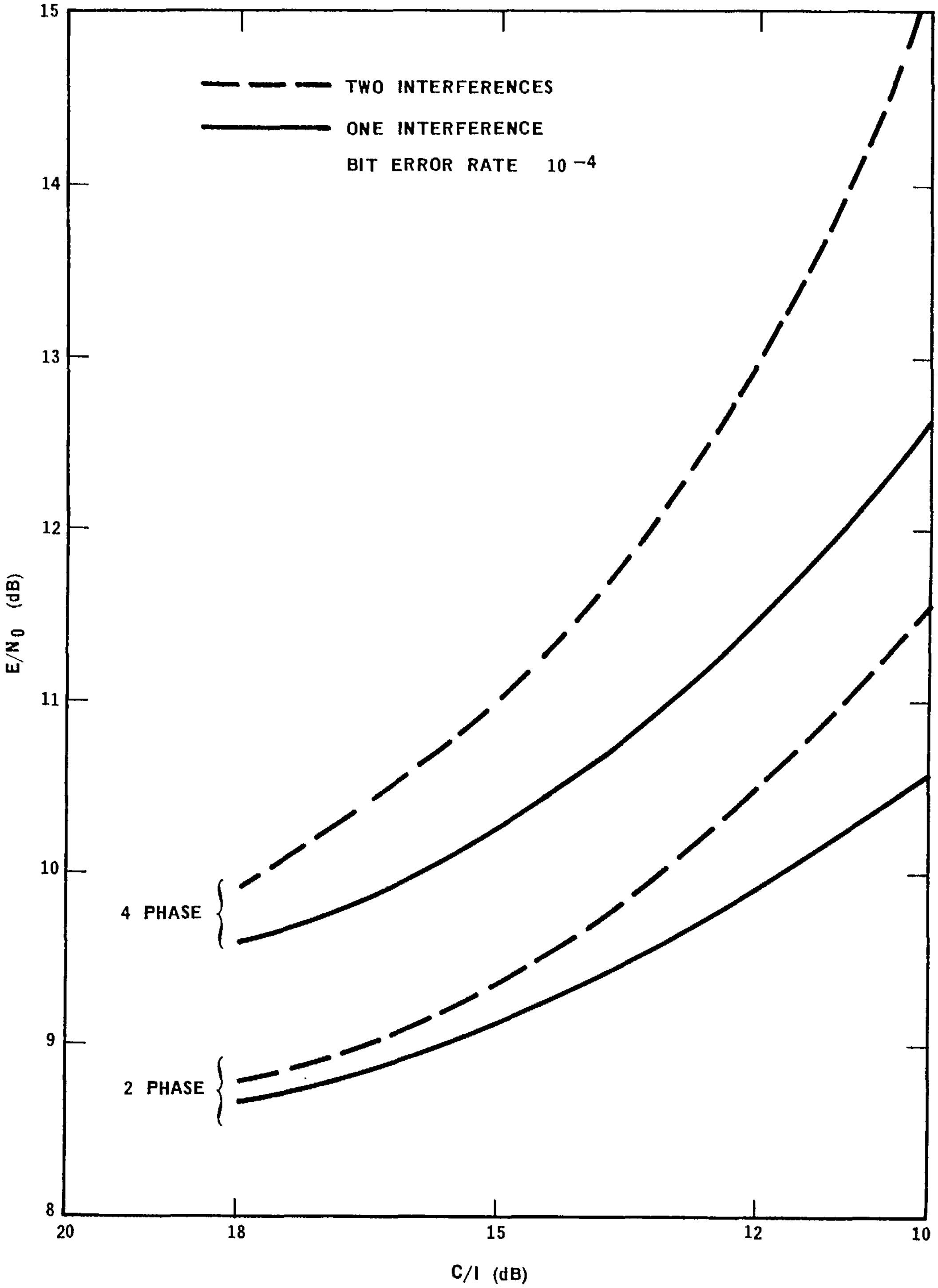

Figure 2-1 Increase in $\frac{E}{N_0}$ to Compensate for Interference

In a system disturbed purely by white gaussian noise, the theoretical ratios of E/No (energy per bit to noise density) for the threshold bit error rate for various levels of PSK are well known[23]. For both the 2-phase and the 4-phase coherent PSK systems, the required E/No for a bit error rate of 1×10^{-4} is given[23] as 8.4 dB. In the presence of interference it is established that additional carrier power is needed to keep the system at the threshold bit error rate. Work carried out at Bell-Northern Research[27] has shown considerable insight into this problem. Figure 2-1 is derived from this work and indicates the carrier power needed in the presence of various degrees of interference. Both the 2-phase and the 4-phase systems are given.

FDM-FM Telephony

For the major traffic communications system, under normal propagation conditions, the system noise would be better than 38 dBrnCo (6300 pW weighted) which is also the current long-haul terrestrial transmission objective. In the case of the speech compressed (lincompex) FDM-FM system, the performance would be at least equivalent to this objective, as explained in Appendix C. In an FDM-FM satellite system, traditionally, approximately 1000 pW (weighted) noise is allocated to interference, that is about 16 percent of the above noise budget. This noise allocation has been considered reasonable and is mainly contributed by interference between terrestrial and satellite systems. However, in a multiple spot beam satellite system with frequency reuse, additional interference within the satellite system now arises. Unlike the terrestrial system interference over which the satellite system designer has little control, the interference created by the frequency reuse situation is a function of satellite system design and is therefore completely under the control of the designer. Under such circumstances it is reasonable to allow a higher allocation of the noise budget to interference, in order to arrive at an economically viable and practical system. In the current system design the noise budget allocated to interference will be 50 percent. All of this allocation may be considered from the satellite system itself as interference from terrestrial sources will not exist due to the exclusive use of the particular frequency bands for satellite communications.

The effect of interference on an FDM-FM system could be compensated for by a higher ratio of carrier-to-thermal noise or by wider frequency deviation, or both. The exact requirements are developed in the system equations in later chapters.

Single-Channel-Per-Carrier FM Telephony

The single-channel-per-carrier FM system will be applied to the major traffic system as well as to the remote communications. In both applications speech compression with lincompex will be used. The reasons for considering this modulation technique for these two services will be given in later chapters. For the major traffic service, the noise performance objective is the same as in the FDM-FM case. In the remote communications, a noise objective of 44 dBrnc0 has been considered to be acceptable and was used in the previous study[14]. As shown in Appendix C, the speech compression technique used would meet this objective with adequate margin.

As for interference considerations, the same remarks for the FDM-FM case apply.

Television (FM)

In the previous study[14], the standards set for the television services were as follows:

a) Commercial Television (CTV): S/N = 57 dB

b) Educational Television (ETV): S/N = 42 dB

$$\text{where } \frac{S}{N} = 20 \log \left[\frac{\text{Sync. tip to white voltage}}{\text{Weighted rms noise voltage}}\right]$$

In an interference-free environment, as shown in Figure 2-2 which is extracted from the report of the previous study, these noise objectives can easily be met with practical bandwidths ranging from say 12 MHz to the threshold limit of 19 MHz for ETV and from say 20 MHz to the threshold limit of 46 MHz for CTV. The optimum bandwidths were found to be about 16 MHz for ETV and 30 MHz for CTV.

If interference is present as might be the case under certain frequency reuse configurations, as for example in the case of satellite antenna Beam Plan (B), (chapter 3), then, in order to maintain approximately the same bandwidth requirement, the noise performance objectives have to be slightly lowered. From Figure 2-3 it can be seen that with interference corresponding to C/I = 18 dB, then for CTV with a bandwidth of 31.25 MHz, the S/N would be about 52 dB. This is based on an operational carrier-to-noise of 15 dB. In the case of ETV using a bandwidth of 15.625 MHz, the S/N would be 40 dB, if the same interference as above were present.

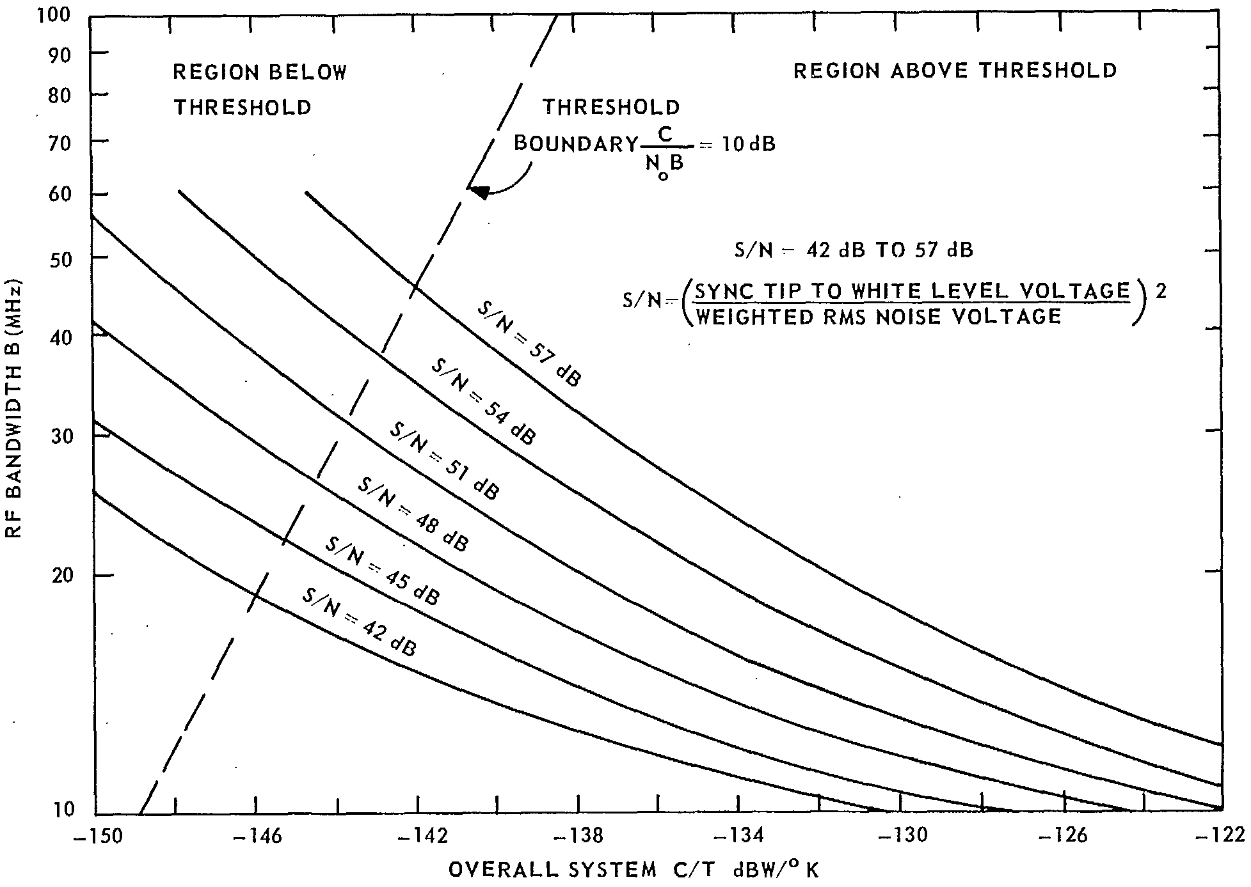

Figure 2-2 Television System Using FM Plots of RF Bandwidth vs C/T

The choice of the specific CTV and ETV bandwidths is explained in chapter 3. The performance degradation applies to co-channel interference only. In the case of ETV, as can be seen in chapter 3, there will be negligible co-channel interference due to the nature of the frequency plans. Hence the ETV service may not degrade to any significant extent. In the case of CTV, if Beam Plan (B) is not optimum and this is the case as it so turns out to be as will be seen in later chapters, then co-channel interference does not arise. Within the context of this study, therefore the interference constraints on the television services do not appear to be a major problem.

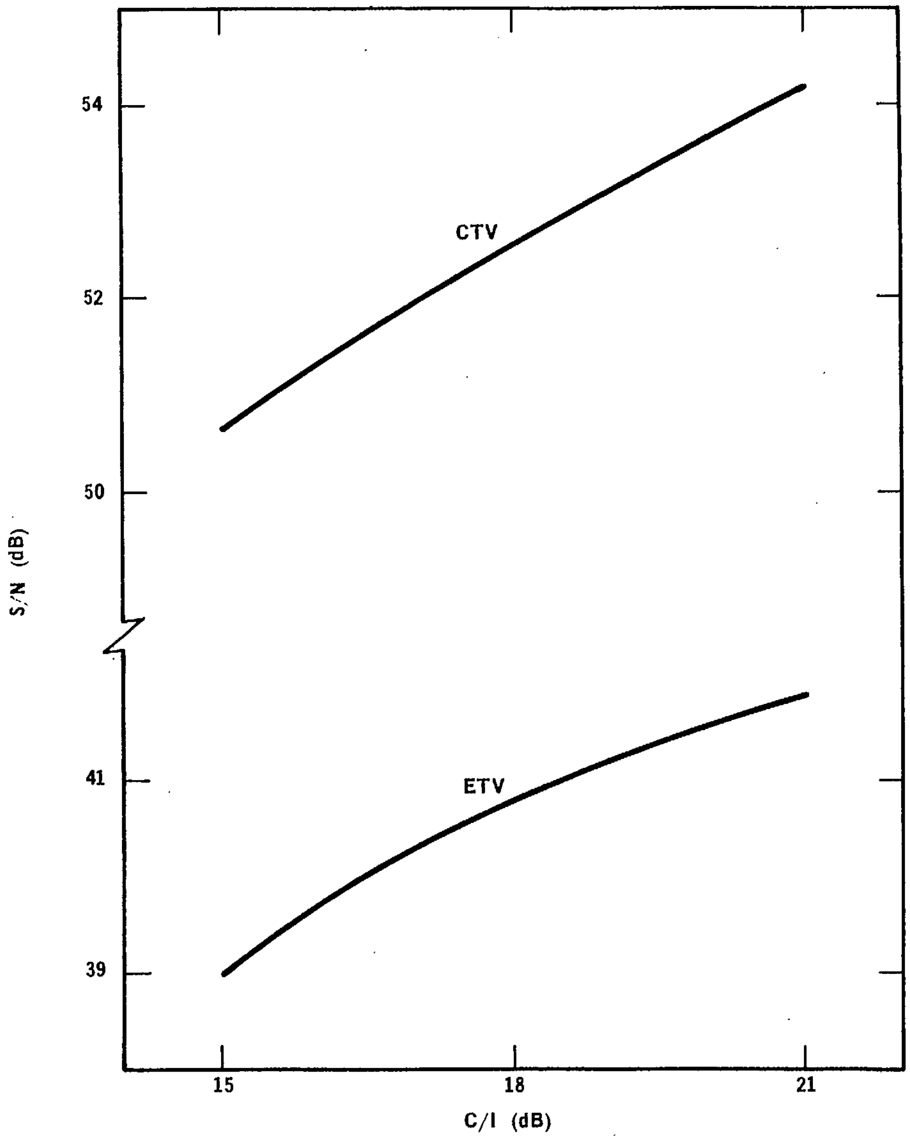

Figure 2-3 Effects of Interference on S/N for TV (Co-channel) for 31.2 MHz Band (CTV) and 15.6 MHz Band (ETV).

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3. SPOT BEAM TECHNOLOGY

The reasons for a spot beam satellite system and the many aspects of spot beam technology are discussed.

There are two basic reasons to use spot beam satellite antennas. One is to enable frequency reuse so that the spectrum available is increased many times. The other is to obtain passive antenna gain.

The major problem areas associated with spot beam technology are antenna designs, interference isolations, and frequency plans for the degree of frequency reuse desired.

This chapter will discuss in detail the above aspects of spot beam technology.

* * *

3. Spot Beam Technology

3.1 REASONS FOR SPOT BEAMS

Spot beams enable frequency reuse among beams, thus increasing by many times the total spectrum available for communications. A spot beam antenna has high passive gain but also has size and weight penalty. There is a trade-off with respect to transponder rf power.

One of the major attributes of satellite communications is area coverage, that is to say, regions that are visible from and to the satellite will be able to communicate with each other. In the simplest configuration, the satellite may have one common transmit-receive beam covering all the regions for which communication is provided. This is also the most flexible system as all the transponders on board the satellite are equally accessible to all earth terminals.

One basic limitation of the single beam system is the total system capacity. By virtue of the fact that the frequency bands allocated to satellite communications are of the order of 500 MHz wide each for up and down links in the 12 to 14 GHz bands concerned, the total capacity of a single beam satellite system will be limited to that provided by this bandwidth. Bearing in mind that the commercial and educational television services must be allocated their share of the bandwidth, the bandwidth available for telephone and data communications may become inadequate in the future. Limited expansion in system capacity can be achieved by employing frequency reuse with orthogonal polarizations. For large scale capacity expansion, frequency reuse employing spot beam isolation is a practical solution, with or without orthogonal polarizations.

The use of spot beams potentially brings another important advantage. The gain of the spot beam antenna is high, thus improving the satellite receive G/T and reducing the transponder output power. However, there is a trade-off consideration here. On the one hand, the narrower the beam, the higher is the antenna gain and so the lower is the transponder power. This in turn tends to reduce transponder weight and cost. Balanced against this on the other hand is the fact that the larger antenna size needed tends to increase satellite weight and cost. There is obviously an optimum point at some stage. Further, with spot beams, the coverage area within each beam and the added complexity of the electronic equipment on board are additional factors that need to be considered. The trade-off is not easy as all these factors interact with each other. The outcome may not necessarily favor the multiple spot beam configuration for any given service needs.

From the foregoing, it can be seen that getting additional available spectrum through frequency reuse is perhaps the major benefit of spot beam technology, with the satellite power and cost benefit as a possibility but by no means a certainty.

3.2 CHOICE OF ANTENNA TYPES

Of the five possible classes of antennas, namely, reflector, lens, horn, traveling wave and array, only the reflector antenna is considered suitable. No single reflector antenna type is significantly superior to the others in all respects and the final choice would depend on relative importance of system constraints.

There are five classes of antennas which may be used for a communication satellite system; reflector, lens, horn, traveling wave and array. The horn antenna is heavy and does not easily lend itself to multibeam operation. Therefore the horn is not considered suitable. The traveling wave antenna is generally frequency sensitive and is not practical as a high gain antenna in the 12 to 14 GHz SHF frequency bands. This type of antenna will therefore not be further considered. In the case of the array antenna, in theory, if a perfectly linear or circular polarized element antenna is used, the resultant pattern will be purely polarized. However, there are problems of designing a practical linearly or circularly polarized element and polarization degradation due to mutual coupling effects is also severe. The insertion loss of the array feed network is high, the bandwidth requirement is difficult to meet and the antenna is likely to be restricted to a single beam. Considering these factors, the array antenna is not a suitable choice for communications satellites at the present. Next is the lens antenna. Both the lens antenna and reflector antenna are designed based on geometrical optics. There are no fundamental differences between these two types of antenna from the view point of electrical performance. The lens suffers no blockage effects and theoretically it is feasible that the side lobe levels near the main beam region be kept low. However, the feed horn forward spillover may wipe out this advantage. The lens antenna in general is heavy, the insertion loss is high and deployability in space is poor. In general, it is considered that the lens antenna is not suitable for communications satellite application.

The class of antenna that remains and warrants serious consideration is the reflector type. In Appendix A, the theoretical considerations of the various types of reflector antennas are presented in depth. These considerations narrow the choice down to four types of antennas as promising candidates:

1) Paraboloidal reflector with on-axis focal feeds

2) Paraboloidal reflector with off-axis feeds

3) Two reflector system (similar to the Cassegrain type but the reflecting surface curvatures are based on Schwarzschild's geometric optics principles).

4) Double parabolic cylinder. (The two cylindrical reflectors are crossed at right angles and each collimates the beam in one plane).

The relative merits of these four types of reflector antenna are summarized in Table 3-1.

From Table 3-1 it is clear that there is no particular antenna that is outstandingly superior to the others. Each type has advantages and disadvantages in different areas. If minimum beam separation is an important consideration, then the double parabolic cylinder appears the best choice. On the other hand, if deployment in space and antenna weight are major considerations, then the paraboloid with on-axis feeds is the best current solution, with the off-axis fed paraboloid running close as a second choice.

In the satellite system under study in which the launch vehicle is constrained to the Thor-Delta, antenna weight and deployment are indeed the major considerations. Hence only the paraboloids with either on-axis or off-axis feeds will be used in the system analysis.

Table 3-1 Comparative Ratings of Four Types of Reflector Antenna

ANTENNA TYPE	SIDE LOBES	CROSS-POLARIZED COMPONENT	MINIMUM BEAM SEPARATION	SIZE AND WEIGHT	EASE OF DEPLOYMENT IN SPACE	FAR FIELD PATTERN
1. Paraboloid with on-axis feeds.	Moderate	Moderate	Moderate	Good	Good	Symmetrical
2. Paraboloid with off-axis feeds.	Good	Moderate	Moderate	Moderate	Moderate	Non-symmetrical
3. Two reflector system	Moderate	Moderate	Poor	Poor	Poor	Symmetrical
4. Double cylinder system	Moderate	Good	Good	Poor	Poor	Non-symmetrical

* * *

3.3 ANTENNA INTERFERENCE CONSIDERATIONS

The main considerations are antenna side lobe and polarization discrimination.

In any system that employs frequency reuse, a major consideration is mutual interference. In a multiple spot beam satellite system the main interference isolation mechanisms are:

a) Antenna Side Lobe discrimination

b) Polarization discrimination (if applicable).

These are discussed in detail in 3.3.1 and 3.3.2.

*　*　*

3.3.1 SIDE LOBES

The antenna side lobe interference isolation can be represented by an expression of the form

$$\frac{1}{[1 + (\frac{\Delta\theta}{\theta o})^4]} - 3 \text{ dB}$$

Antennas of the current Intelsat satellites and of the Anik I have not been designed with frequency reuse application in mind. Consequently side lobe suppression was not an important design objective. Information on satellite antennas with good side lobe suppression is therefore not available. On the other hand a great deal of information on side lobe characteristics of terrestrial earth station antennas have been gathered, especially through CCIR sources. The current CCIR agreed expression for the envelope of the side lobe peaks is given as follows:

$$G_{S.L} = 32 - 25 \log \theta \qquad dB_i \qquad (3.1)$$

where $G_{S.L}$ = Isotropic gain of antenna in direction θ degrees from the antenna axis

This formula was derived empirically from data based on existing large earth stations. The formula is not applicable to the satellite antenna side-lobe characteristic because of the following considerations:

a) Existing large earth station antennas have been designed with amphasis on high on-axis gain and low noise temperature. This calls for near constant amplitude and phase distribution of the antenna aperture illumination. This type of illumination is well known to have poor side lobe characteristics.

b) The prime interest of CCIR is the mutual interference between earth stations and terrestrial microwave systems. Wide-angle side lobes (up to ±90°) are of major concern and the formula has been derived as a best fit to these side lobe levels. In a multiple spot beam antenna satellite system, the main area of concern are side lobes close to the main beam, say not more than ±4° away in the case of Canada coverage.

A more accurate expression than the CCIR formula is warranted.

In order to assess the side lobe interference isolation it is therefore necessary to turn elsewhere for guide lines. In Figure 2, Appendix A, it is seen that a side lobe peak envelope function of the form:

$$F(\theta) = \frac{1}{[1 + (\frac{\Delta\theta}{\theta o})^4]} \quad (3.2)$$

where $\Delta\theta$ = Angle off the main beam axis, in degrees
θo = Half the 3 dB beamwidth, in degrees.

describes adequately an antenna having an aperture amplitude distribution of the form:

$$A = \cos (\frac{\pi}{2}x) \quad (3.3)$$

A more tapered illumination would have side lobe characteristics better than that given by equation (3.2). Practical radar antenna designs carried out by Northern Electric Company showed that indeed side lobe response much lower than that given by equation (3.2) was achievable.

In a practical antenna design for multiple spot beam satellite application, the following are factors that would tend to degrade side lobe performance:

a) Defocussing and other aberrations resulting from the physical problem of locating many feeds

b) Antenna surface errors due to manufacture, launch, deployment, and thermal distortion in space

c) Imperfect feed horn phase pattern

d) Feed horn and supporting structure blockage

e) Possibly, mutual coupling effects between feeds

From these considerations there is a need to be conservative and it appears reasonable to adopt equation (3.2) as the expression describing the peak side lobe envelope of a practical satellite multiple spot beam antenna system. This expression gives the ratio of on-axis gain over the off-axis gain. In a homogeneous system in which all carriers have the same effective isotropic radiated power (EIRP), then the net isolation between any two beams is given by:

$$\frac{C}{I} = F(\theta) - 3\ \text{dB} \tag{3.4}$$

where C/I is the carrier-to-interference ratio and where the reduction of 3 dB is due to the fact that in the worst case situation the wanted carrier would be from or to an earth station located at the -3 dB contour of the coverage zone. A plot of equation (3.4) is given in Figure 3-1.

The earth stations have the main lobes always pointed at the satellite. Earth station side lobe responses are therefore not relevant to the interference problem under discussion.

In the subsequent sections, equation (3.4) and equation (3.2) will be used to estimate the side lobe interference isolations for co-polarized channels.

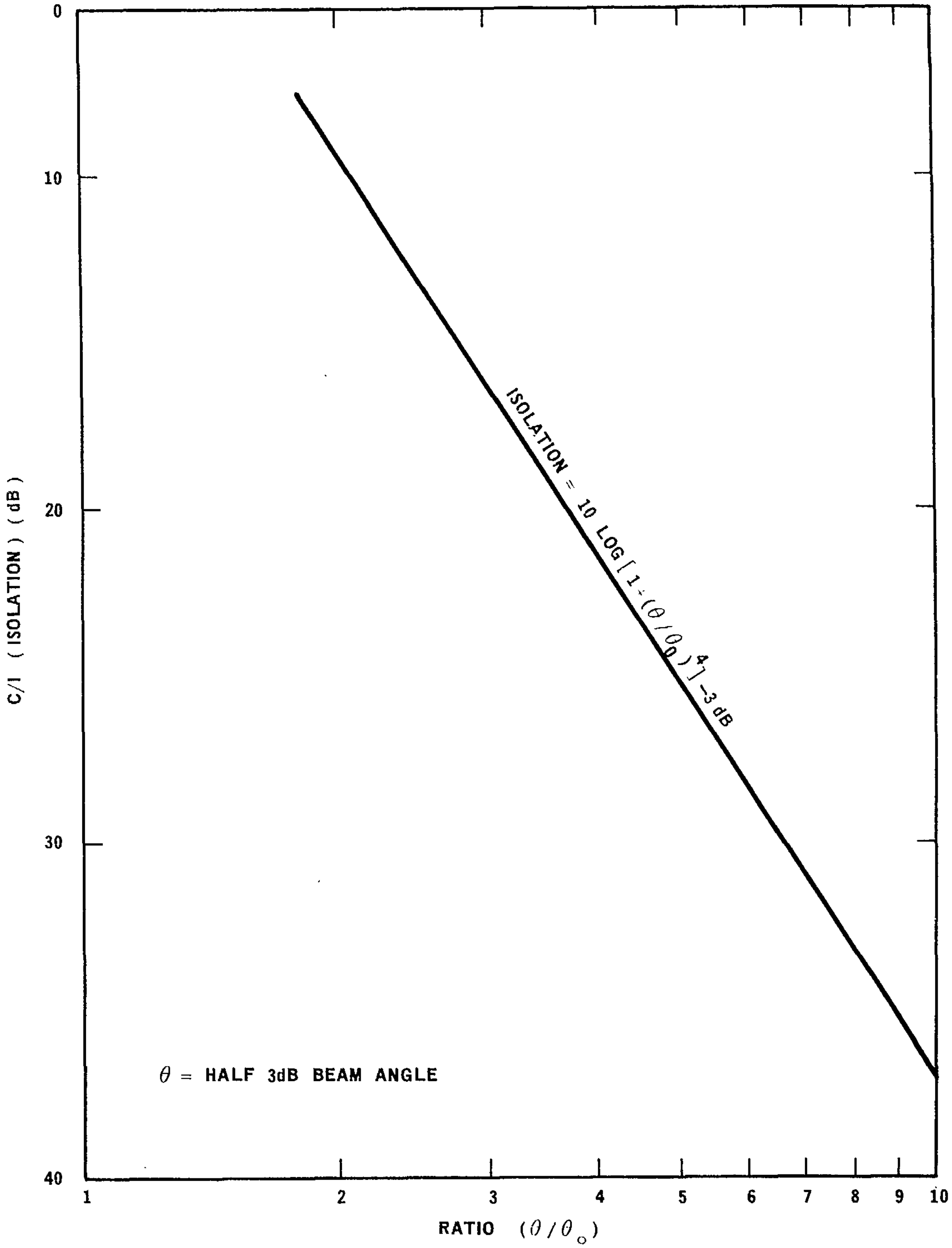

Figure 3-1 General Isolation Curve

3.3.2 POLARIZATION

Three separate cross-polarization couplings are the earth station antenna, the satellite antenna and the propagation path.

Just as in the case of the side lobe suppression, present day communication satellite and earth station antennas have not been designed with good polarization isolation as major objectives. Even so, the polarization discrimination achieved has been reasonably good[7,8]. In discussing the overall satellite link isolation it is necessary to consider the three separate contributions individually:

a) Due to earth station antenna

b) Due to satellite antenna

  (i) Within main beam
  (ii) In side lobe region

c) Due to propagation path

Polarization Discrimination in Earth Station Antenna

The earth station antenna is generally symmetric in shape, usually circular, and its axis is always lined up with the satellite. The cross-polarized component due to the antenna reflector itself is therefore negligible. If careful attention is given to the feed support design, cross-polarized component due to this source is also negligible. The main contribution then is due to the feed subsystem. For reasons to be presented later, linear polarization would be used. Feeds with linear polarization can be designed with cross components greater than 35 dB down, over the 500 MHz band. As a margin of safety, the earth station polarization isolation will be taken as 30 dB.

Polarization Discrimination in Satellite Antenna

There are two separate regions of cross-polarization coupling that need to be considered in a multiple beam satellite antenna system, viz:

a) Within the main beam
b) In the side lobe region.

Cross-polarization coupling within the main beam is of importance if dual orthogonal polarization frequency reuse is employed within the same beam. In this region, as can be seen in Figure A-1 of Appendix A, and also in CCIR documents[24] the polarization discrimination of present day satellite antennas in the 4 to 6 GHz band can achieve a value of between 30 and 35 dB. It is expected that the spot beam antennas in the 12 to 14 GHz bands, if designed with adequate regard to the cross-polarization coupling can also achieve this objective. For the purpose of the current system design, however, a figure of 27 dB can be used with reasonable confidence.

The cross-polarization coupling in the side lobe region would be important in the case where frequency reuse is applied in different beams and it is desired to obtain additional isolation by alternating the polarization vectors in the various beams. Again in CCIR documents[24] measured data indicates that the cross components in the side lobes are generally more than 10 dB below the co-polarized components. A figure of 10 dB will be used in this study.

Polarization Coupling in Propagation Path

The polarization coupling that may be attributed to the propagation path is due to:

a) Errors in satellite antenna polarization alignment due to spacecraft attitude control instabilities.

b) Errors in earth station antenna polarization alignment.

c) Polarization rotation due to Faraday effects.

d) Scattering effects from rain and other atmospheric disturbances.

Items a), b), and c) are significant only in the case of linear polarization. The coupling due to alignment errors in the satellite and earth station antennas can be made negligibly small, as it is feasible to control the satellite attitude to within ±0.1° in three axes. Faraday rotation is a major problem in the lower SHF band such as the 4 to 6 GHz and polarization tracking is necessary. However, in the 12 to 14 GHz band the Faraday rotation is not expected to exceed about one degree giving rise to a cross-polarization coupling 27 dB down. Polarization tracking may therefore not be required. There is at present inadequate information about the polarization coupling due to rain and other atmospheric disturbances along a satellite - earth transmission path, especially in the 12 to 14 GHz band. From the limited terrestrial measurements available[25], it is expected that the polarization coupling may be of the order of 25 dB over a satellite path for earth stations with elevation angles 20° through 30°, that is, in most parts of southern Canada. The total polarization coupling due to all propagation causes would not therefore be expected to exceed about 22 dB. This figure will be used in this study.

From the point of view of propagation, there is little to choose between linear and circular polarization at the 12 to 14 GHz frequency bands. However, from the point of view of hardware design, both in the satellite and in the earth stations, it is much easier to use linear polarization, especially if high power and broadband characteristics are important. Hence in the present system study, linear polarization will be postulated.

* * *

3.3.3 OVERALL INTERFERENCE DISCRIMINATION

Different sources of interference will be treated on the power additive basis.

When considering the total effects of so many interference sources, strictly speaking, one should separate the effects of coherent and non-coherent interference, as in the one case it is voltage addition and in the other case it is power addition. However, in the broad systems design which is the present objective, and in the light of the imprecise knowledge of the various interference isolations, it does not warrant in the present approach, a more rigorous analysis than treating the various interference contributions on a power-additive basis. This broad approach would apply to all types of modulation technique and any additional interference improvement factor that may arise as a result of channel frequency spacing or frequency staggering will be dealt with separately.

*   *   *

3.4 CHOICE OF BEAM SIZES AND NUMBER OF BEAMS

Three spot beam plans are selected for systems evaluation.

As discussed in 3.1, it does not necessarily follow that the narrower the spot beam and hence the higher the passive gain, the lighter will be the overall satellite weight. There is a trade-off between antenna weight and weight due to RF power amplifiers and associated solar panels. It is, however, unrealistic to approach the trade-off problem on a hypothetical basis by varying only the beam size. Additional factors that need consideration are the effect of beam size on the total number of beams required, which in turn affect the number and complexity of on-board equipment and their interconnections. Further, the beam size and spacing have an effect on the mutual interference problem and choice of frequency plans. Thus to solve the problem realistically, it is necessary to consider all these interrelated effects simultaneously. The only practical way is to postulate several plausible models based on the actual geography and traffic needs of the country, optimize each model on a cost trade-off basis and compare the end results. The models would cover a reasonable range of spot beam size and number of beams. In this way a general trend should emerge as to the overall optimum solution.

Based on the above considerations, three spot beam plans are proposed in this study, designated Beam Plans (A), (B) and (C). These plans are briefly discussed below.

Beam Plan (A)

As identified in 2.1, the major toll telephone traffic in Canada is concentrated in about eight small regions. The first model therefore is to have one spot beam for each region. The size of the beam chosen is 0.5° at the -3 dB contour. Allowing for a satellite antenna pointing instability of ±0.1°, the actual coverage area would be a circular area corresponding to a beam size of 0.3°. The choice of this beam size is governed by the following considerations:

a) Certain beams must cover more than one earth station, such as the beam for Edmonton/Calgary or Montreal/Quebec City. A coverage area of 0.3° would be about the minimum size.

b) If the antenna is to be a single reflector type such as a paraboloid, the minimum separation between adjacent beams needs to be about 1.5 times the beamwidth. From the geographical locations of the eight major traffic centres, a physical beam size of no larger than about 0.5° would just meet this constraint.

In Beam Plan (A), the 8 spot beams would cater for the major route traffic system. For remote communications, it seems logical to use the 5 medium size beams proposed for the educational television (ETV) system in the previous DOC study as these beams seem best to serve the intra-provincial or intra-regional needs[14]. Hence the 8 spot beams and 5 medium beams would constitute Beam Plan (A). This is illustrated in Figure 3-2.

Beam Plan (B)

Beam Plan (B) is a radical departure from Beam Plan (A), in the sense that the whole country is now covered by a series of uniform size, overlapping beams. After making the allowance for satellite antenna pointing instability of ±0.1°, the coverage area per beam is chosen to be about 1.5° x 0.5° (elliptical shape) the actual beam size being 1.7° x 0.7°. The reasons are as follows:

a) Canada is spread out length-wise in the East-West direction. A beam with a coverage axis of 1.5° in a general North-South direction would cover from North to South of Canada.

b) Each beam axis in the East-West direction should, on the one hand, be reasonably narrow in order to give adequate antenna gain. On the other hand too narrow a beam axis would require many beams in order to cover the whole country. Further, the beam instability of ±0.1° would extract a disproportionately large penalty in coverage area if the beam width is too small. In Appendix B a mathematical relationship is established between beam pointing stability, interference objectives and optimum antenna size and it is shown that a beam size of about 0.7° is optimum for the geographical shape of Canada.

c) A beam shape aspect ratio of about 3 to 1 (ratio of major to minor axis) is reasonably easy to design. Larger aspect ratios are feasible but would be difficult to achieve.

For Beam Plan (B), a total of 14 uniform size beams would be adequate to cover all Canada. Both major traffic and remote communications would use the same beams. Thus compared to Beam Plan (A), the major traffic system would have lower gain antennas whereas the remote communications system would have higher gain antennas. One advantage of a common beam coverage for both major traffic and remote communications is the possibility of designing an integrated network for these two services. The footprints for Beam Plan (B) are given in Figure 3-3.

Beam Plan (C)

From the point of view of major traffic communications, only 8 beams are necessary in both Beam Plans (A) and (B). The other beams in these plans are attributed to other services, such as remote communications. Since the spot beam sizes for the major traffic service vary between these two plans, it will be possible to see the effect of beam size on the overall cost for this service. For major traffic communications it is also possible to re-orient the beam axis of the elliptical shape beams in such a manner that only 4 beams are required. This will yield information on the cost benefit resulting from such reduction of number of beams. This plan is known as Beam Plan (C) and the footprints are given in Figure 3-4.

It is considered that the above three beam plans adequately cover the range of likely configuration applicable to the Canadian traffic demands and their geographical distribution.

The practical implementation of the various antenna beam plans in terms of hardware is quite difficult but is considered feasible. For the beam sizes considered, the antenna dimensions would be larger than the shroud size of the Thor-Delta launcher and deployment in space after launch is necessary. Two reflectors are needed in every case:

Beam Plan (A)	Spot Beams	12.5 ft × 12.5 ft	
	Medium Beams	5 ft × 5 ft	(See *Note*)
Beam Plan (B)	Elliptical Beams		
	Odd Beams	3.3 ft × 9.7 ft	
	Even Beams	3.3 ft × 9.7 ft	
Beam Plan (C)	Elliptical Beams	9.7 ft × 3.3 ft	
	Medium Beams	5 ft × 5 ft	(See *Note*)

Paraboloids with off-set feeds appear to be a practical solution.

Note: The design of this antenna is given in the previous study[14].

* * *

3.5 RF FREQUENCY PLANS

From considerations of capacity needs of the beams with densest traffic and interference isolations by side lobes and cross-polarizations, three frequency plans are developed for the on-board switched satellite models and three for the non-switched models.

The prime frequency bands available for the satellite system will be taken to be:

a) Up-link: 14.0 through 14.5 GHz

b) Down-link: 11.7 through 12.2 GHz

This is in line with the previous study[14].

The 500 MHz bandwidth available has to be shared among the various services, including commercial and educational television. Whilst it is not within the scope and terms of reference of the present study to include system evaluation for the television services, it is nevertheless essential in developing any practical frequency and antenna beam plans, to take into account the effect of mutual interference on the quality of the television services.

In a multiple spot beam satellite system, the receivers and transmitters in the various beams have to be interconnected in some fashion. This interconnection can be in the time, space or frequency domains. Whatever is the chosen mode of interconnection it is axiomatic that the interconnected subsystems must be compatible with each other. It follows therefore that, for any practical design, the transponders must all be of equal bandwidth for any particular system. The next question then is "what is the best unit of transponder bandwidth?" It is now necessary to look at the television services for a logical answer.

In a system carrying telephone messages or data signals, there is no inherent preferred rf bandwidth as such. The bandwidth required is a function of capacity. However practical considerations might constrain the bandwidth to some upper limit from considerations of cost of high speed memory or logic devices in a digital system. In an analog system, a corresponding constraint may be the highest forecast capacity needs from the densest traffic centre. In any case this upper limit is ill-defined.

In the television services, however, the rf bandwidth requirements are well defined, once the quality of the service is specified. It therefore seems logical to use the bandwidth of a television transponder as a basic unit. Once this is done, then the message transponder bandwidths can be multiples or submultiples of this basic bandwidth.

In 2.2, it is indicated that the bandwidth required for commercial television is about 30 MHz. If the 500 MHz band is divided into 16 units then each unit would give 31.25 MHz. This is close enough to the television requirement and will therefore be the basic transponder bandwidth unit adopted in this study. This bandwidth unit is also compatible to the educational TV service, which as indicated in the previous study[14], requires a bandwidth of about 16 MHz, or approximately half of 31.25 MHz.

In evolving any frequency plan to fit any particular antenna beam plan, it is necessary to take into account the total amount of interference isolation achievable. All interference contributions must be considered, both up-link and down-link. The relative beam sizes for up- and down-links therefore need to be considered. Since the up-link frequency is slightly higher, one would normally expect the up-link beam to be narrower. However, in the multiple feed horn antenna system design where physical space for the feed horns is a major problem, it is most likely that a common transmit and receive horn would be used for each beam. At the higher frequency, the horn illumination would be narrower which in effect results in a smaller illuminated antenna aperture. This has the effect of broadening the main beam. The overall result would be that the up- and down-link beamwidths would be approximately the same, so are the antenna gains. For the current satellite system design this would be taken to be the case.

Applying the side lobe discrimination results of Figure 3-1 and the polarization discriminations discussed earlier, it is possible to derive the total interference in each beam corresponding to various patterns of frequency reuse (i.e. full reuse, half reuse, etc). The results are presented in Tables 3-2, 3-3 and 3-4.

NOTE: The term full frequency reuse means that the 500 MHz band is repeatedly reused in every beam. The term half frequency reuse means that odd rf channels are used in odd numbered beams and even rf channels are used in even numbered beams. The same idea is extended to one-third, one-fourth, etc, frequency reuse.

Table 3-2 Interference Isolations for Beam Plan (A)

a) Between 8 Spot Beams

SPOT BEAMS	FULL FREQUENCY REUSE SINGLE CO-POLARIZATION (dB)	HALF FREQUENCY REUSE SINGLE CO-POLARIZATION (dB)	HALF FREQUENCY REUSE DUAL POLARIZATION (dB)
Vancouver	19.6	30	21.4
Calgary/Edmonton	16.5	27.5	21.0
Regina	14.5	30	21.4
Winnipeg	16.0	27.5	21.0
Toronto	15.0	24.0	19.9
Montreal	13.5	25.0	20.2
Halifax	12.0	24.0	19.9
St. Johns, Nfld.	15.0	25.0	20.2

b) Between 5 Medium Beams

MODE OF FREQUENCY REUSE	INTERFERENCE ISOLATION (dB)
Full frequency reuse, single co-polarization	Not feasible
Half-frequency reuse, single co-polarization (Worst case interference angle = 2 half beamwidths)	7
Half-frequency reuse, with alternative orthogonal polarizations in adjacent beams.	17

Table 3-3 Interference Isolations for Beam Plan (B)

MODE OF FREQUENCY REUSE	INTERFERENCE ISOLATION (dB)
Full frequency reuse	Not feasible
Half-frequency reuse	Not feasible
One-third frequency reuse	16
One-fourth frequency reuse	22
One-fourth frequency reuse with dual polarization reuse. (Cross-component isolation $\approx$22 dB)	19

Table 3-4 Interference Isolation for Beam Plan (C)

MODE OF FREQUENCY REUSE	INTERFERENCE ISOLATION (dB)
Full frequency reuse	Not feasible
Half-frequency reuse	20
Half-frequency reuse, with dual polarization reuse. (Cross-polarization isolation $\approx$22 dB)	18

Using as a desirable objective a minimum interference isolation of about 18 dB, it will be noted from Tables 3-2, 3-3 and 3-4 that the following frequency reuse schemes for the three beam plans are feasible:

Beam Plan (A)

8 Spot Beams: Half-frequency reuse with co-polarization. Although the 18 dB isolation criterion would permit dual polarization frequency reuse, this is not necessary in Beam Plan (A) because of adequate bandwidth availability with co-polarization alone.

5 Medium Beams: Half-frequency reuse, with alternate orthogonal polarizations in different beams.

Beam Plan (B)

14 Uniform Beams: One-fourth frequency reuse with dual polarization frequency reuse.

Beam Plan (C)

Half-frequency reuse with dual polarization frequency reuse.

In developing a frequency plan it is necessary to consider the total frequency spectrum needed by the beams serving the regions with the densest traffic, such as Toronto and Montreal. The spectrum in the beams serving the other regions will be excessive, but this is anavoidable. Further it is necessary to divide the frequency plans into two groups, one for the satellite system with on-board switching and the other for the non-switched system. This is because of the fact that the nature of the system configurations is quite different and a common frequency plan cannot be optimum for both. In the switched model, it is advantageous both from the point of view of cost and system simplicity to have as wide a transponder bandwidth as practicable. In the non-switched system, however, there is a need for a minimum number of interconnecting links between the various spot beams. The rf channel bandwidth is therefore governed by this number. The frequency plans for the switched and non-switched models are given in Tables 3-5a and 3-5b respectively.

Table 3-5a Switched Model Frequency Plans

FREQUENCY PLAN	BEAM PLAN	MESSAGE TRANSPONDER BANDWIDTH (MHz)	CTV TRANSPONDER BANDWIDTH (MHz)	ETV AND R.C. TRANSPONDER BANDWIDTH (MHz)
Plan (S1)	Plan (A)	125	31.25	31.25
Plan (S2)	Plan (B)	62.5	31.25	31.25
Plan (S3)	Plan (C)	125	31.25	31.25

Table 3-5b Non-Switched Model Frequency Plans

FREQUENCY PLAN	BEAM PLAN	MESSAGE TRANSPONDER BANDWIDTH (MHz)	CTV TRANSPONDER BANDWIDTH (MHz)	ETV AND R.C. TRANSPONDER BANDWIDTH (MHz)
Plan (N1)	Plan (A)	15.625	31.25	31.25
Plan (N2)	Plan (B)	15.625	31.25	31.25
Plan (N3)	Plan (C)	62.5	31.25	31.25

Table 3-6a. Frequency Plan (S1) for Beam Plan (A), Switched Model

RF Chan	VAN	CAL	REG	WIN	TOR	MONT	HALIF	St. J.	Medium (B.C.)	Medium (SASK)	Medium (ONT)	Medium (QUE)	Medium (MARIT)
1	M				M								
2													
3													
4			M				M						
5		M				M							
6													
7													
8				M				M					
9	◄— CTV (half-Canada beam) —►												
10									2:E		RC		2:E
11	◄— CTV (half-Canada beam) —►												
12										2:E		RC	
13	◄— CTV (half-Canada beam) —►												
14									RC		2:E		RC
15	◄— CTV (half-Canada beam) —►												
16										RC		2:E	

Table 3-6b. Frequency Plan (S2) for Beam Plan (B), Switched Model

RF CHAN	BEAM 1	BEAM 2	BEAM 3	BEAM 4	BEAM 5	BEAM 6	BEAM 7	BEAM 8	BEAM 9	BEAM 10	BEAM 11	BEAM 12	BEAM 13	BEAM 14
1	M				M				M				M	
2	M				M				M				M	
3		M				M				M				M
4		M				M				M				M
5			M				M				M			
6			M				M				M			
7				M				M				M		
8				M				M				M		
9	C C				2:E RC				C C				2:E RC	
10		C C				2:E RC				C C				2:E RC
11			C C				2:E RC				C C			
12				C C				2:E RC				C C		
13	2:E RC				C C				2:E RC				C C	
14		2:E RC				C C				2:E RC				C C
15			2:E RC				C C				2:E RC			
16				2:E RC				C C				2:E RC		

Table 3-6c. Frequency Plan (S3) for Beam Plan (C), Switched Model

RF CHAN	BC/ALB	SASK/MAN	ONT/QUE	MARIT	Medium (BC)	Medium (SASK)	Medium (ONT)	Medium (QUE)	Medium (MARIT)
1	M		M						
2									
3									
4	M		M						
5		M		M					
6									
7									
8		M		M					
9	← CTV (half-Canada beam) →								
10					2:E		RC		2:E
11	← CTV (half-Canada beam) →								
12						2:E		RC	
13	← CTV (half-Canada beam) →								
14					RC		2:E		RC
15	← CTV (half-Canada beam) →								
16						RC		2:E	

Table 3-6d. Frequency Plan (N1) for Beam Plan (A), Non-Switched Model

RF CHAN	Spot beams												
	VAN	CAL	REG	WIN	TOR	MONT	HALIF	St. J.	Medium (BC)	Medium (SASK)	Medium (ONT)	Medium QUE)	Medium (MARIT)
1A	M		M		M		M						
1B		M		M		M		M					
2	← CTV (half-Canada beam) →												
3A	M		M		M		M						
3B		M		M		M		M					
4	← CTV (half-Canada beam) →												
5A	M		M		M		M						
5B		M		M		M		M					
6	← CTV (half-Canada beam) →												
7A	M		M		M		M						
7B		M		M		M		M					
8	← CTV (half-Canada beam) →												
9A	M		M		M		M						
9B		M		M		M		M					
10									2:E		RC		2:E
11A	M		M		M		M						
11B		M		M		M		M					
12										2:E		RC	
13A	M		M		M		M						
13B		M		M		M		M					
14									RC		2:E		RC
15A	M		M		M		M						
15B		M		M		M		M					
16										RC		2:E	

Table 3-6e. Frequency Plan (N2) for Beam Plan (B), Non-Switched Model

RF CHAN	1	2	3	4	5	6	7	8	9	10	11	12	13	14
1A	M M				M M				M M				M M	
1B			M M				M M				M M			
2A		M M				M M				M M				M M
2B				M M				M M				M M		
3A	M M				M M				M M				M M	
3B			M M				M M				M M			
4A		M M				M M				M M				M M
4B				M M				M M				M M		
5A	M M				M M				M M				M M	
5B			M M				M M				M M			
6A		M M				M M				M M				M M
6B				M M				M M				M M		
7A	M M				M M				M M				M M	
7B			M M				M M				M M			
8A		M M				M M				M M				M M
8B				M M				M M				M M		
9	C C				2:E RC				C C				2:E RC	
10		C C				2:E RC				C C				2:E RC
11			C C				2:E RC				C C			
12				C C				2:E RC				C C		
13	2:E RC				C C				2:E RC				C C	
14		2:E RC				C C				2:E RC				C C
15			2:E RC				C C				2:E RC			
16				2:E RC				C C				2:E RC		

Table 3-6f. Frequency Plan (N3) for Beam Plan (C), Non-Switched Model

	Spot beams								
RF CHAN	BC/ALB	SASK/MAN	ONT/QUE	MARIT	Medium (BC)	Medium (Sask)	Medium (ONT)	Medium (QUE)	Medium (MARIT)
1	M		M						
2	M		M						
3		M		M					
4		M		M					
5	M		M						
6	M		M						
7		M		M					
8		M		M					
9	← CTV (half-Canada beam) →								
10					2:E		RC		2:E
11	← CTV (half-Canada beam) →								
12						2:E		RC	
13	← CTV (half-Canada beam) →								
14					RC		2:E		RC
15	← CTV (half-Canada beam) →								
16						RC		2:E	

Full details of the rf channel and polarization allocations are given in Tables 3-6a through 3-6f. It will be noted from Table 3-7 that for the major route telephone traffic service all the frequency plans yield the same total system capacity, that is, the capacity corresponding to a total of 1000 MHz.

Table 3-7 Total Bandwidth for Major Traffic System

FREQUENCY PLAN	NUMBER OF BEAMS	NUMBER OF CHANNELS/BEAM	R.F. CHANNEL BANDWIDTH (MHz)	TOTAL BANDWIDTH (MHz)
Plan S1	8	1	125.0	1000
Plan S2	8*	2	62.5	1000
Plan S3	4	2	125.0	1000
Plan N1	8	8	15.625	1000
Plan N2	8*	8	15.625	1000
Plan N3	4	4	62.5	1000

* 8 beams out of 14 are adequate for the major route communications system.

This common system capacity would enable a valid comparison of the costs and technical merits between the various system models evaluated in this study. Although the 1000 MHz would give a capacity far exceeding the 10 000 voice channels specified, as for example, a 4-phase PSK system would give a total system capacity of 24 700 channels, there is a need to meet the demands of the densest traffic regions such as Toronto and Montreal. With the models investigated, which is the only practical approach, there is bound to be an excess capacity in the other regions of Canada.

In all the frequency plans postulated, in the case of the analog fm commercial and educational television services, wherever feasible, an rf channel would be used for commercial television in one beam and for educational television in another beam so that additional advantage may be taken of non-co-channel interference effects.

One important advantage of the multiple spot beam satellite system is that the isolation afforded by the beam discrimination, together with interleaved frequency plans as adopted in this study, makes it unnecessary to have guard bands in the channellized satellite transponders. This improves spectrum utilization. The up-link bandwidth still needs to be controlled, but the stringent filtering is now done in the ground equipment and is therefore easily achieved.

* * *

3.6 SATELLITE RECEIVE/TRANSMIT ANTENNA GAINS AND G/T_{SAT}

Satellite antenna gains and satellite receive system G/T are given.

The gain of a satellite antenna within its 3 dB contour can be estimated from the following expression[16].

$$G = 10 \log \left(\frac{6000\pi}{\theta_E \times \theta_H}\right) - 2 \text{ dB} \tag{3.5}$$

where θ_E and θ_H are 3 dB beamwidths in degrees in two orthogonal planes.

As discussed in 3.5 because of the peculiar nature of the antenna illumination, it is expected that the up- and down-link antenna gains would be about the same, even though the up-link frequency is slightly higher. Allowing for a transmit and receive diplexer loss of 0.8 dB each, the effective antenna gains for down-links and up-links for the various beam plans are therefore given in Table 3-8.

Table 3-8 Effective Antenna Gains

BEAM TYPE	BEAMWIDTH	GAIN (dB)
Spot Beam (Plan A)	0.5° x 0.5°	45.9
Medium Beam (Plan A)	1.5° x 2.0°	35.2
Elliptical Beam (Plan B and Plan C)	1.7° x 0.7°	39.2

The satellite receive system noise temperature and figure-of-merit $(G/T)_{sat}$ are estimated as follows:

It is anticipated that a TDA would be used as the front end low noise receiver. Although the uncooled paramp could be practicable in the time frame considered, the number of low noise receivers required in a multiple spot beam configuration is large and the paramp would then present reliability and weight problems.

TDA Noise temperature (N_F = 5.5 dB) = 740°K
TDA net gain = 12 dB
Down converter noise temperature (N_F = 8 dB) = 1535°K
Diplexer loss (0.8 dB) noise temperature = 50°K
Noise temperature of earth as seen by antenna = 290°K
Total system noise temperature:

$$T_s = 740 + \frac{1535}{16} + 50 + \frac{290}{1.2} = 1128°k = 30.5 \text{ dB/°K}$$

Combining this system noise temperature with the effective up-link satellite antenna gains of Table 3-8, the (G/T)'s of the various beams are obtained and are tabulated in Table 3-9.

Table 3-9 Satellite (G/T)

BEAM TYPE	SATELLITE (G/T) dB/°K
Spot Beam Plan (A)	+15.4
Medium Beam Plan (A)	+ 4.7
Elliptical Beam, Plan (B) and Plan (C)	+ 8.7

4. MAJOR TRAFFIC SATELLITE SYSTEM WITH ON-BOARD SWITCHING

The technical features with block diagrams of various system configurations are described. The system power trade-off equations are derived.

This chapter examines three satellite system models that use on-board time division beam switching techniques. The principles of operation are briefly described, the block diagrams of each model are given and the system power trade off equations are derived.

* * *

4. Major Traffic Satellite System with On-Board Switching

4.1 PRINCIPLES OF SDMA/SS-TDMA

The principles of operation of a satellite system with on-board time division switching among the spot beams are briefly explained.

The letters SDMA/SS-TDMA stand for Space Division Multiple Access/Satellite Switched - Time Division Multiple Access. Before explaining the satellite switcher system it is first necessary to briefly describe the "conventional" TDMA system. In the TDMA system, in its simplest form, all the earth stations communicating with each other would be located within one common satellite coverage antenna beam, using one common broadband transponder. The system would have to use digital modulation. Each earth station would receive from its terrestrial source the digital information at relatively slow but continuous bit rate, store the information in a buffer memory and send out to the satellite in a short burst at a high bit rate. This high bit rate transmission via the satellite would occupy the full transponder power and bandwidth during the burst. The bursts from the earth stations are arranged in an orderly manner on a time division basis such that no two bursts would overlap when the signals arrive at the transponder input. In order to ensure this orderly arrival of signals or bursts, a network synchronization scheme is necessary. Many techniques are feasible, and the currently favored scheme in Intelsat[18] appears to be one using a reference station with the rest of the network synchronizing to it. Each station would send a "preamble" data stream to the satellite immediately ahead of the communication data bits, measure the satellite round trip delay of its own station identification code which is part of the preamble, and compare the delay with the arrival of the reference station code. This is the feed back information necessary to enable each station to adjust its burst timing and achieve network synchronization. This scheme is decentralized and each station looks after itself in accessing the network. Centralized synchronizing schemes are feasible, but in the context of the Canadian application in which the number of major traffic centers is small, it is doubtful if there is any significant difference one way or the other in using a centralized control.

The need for time division switching on board the satellite arises because of the use of different spot beams for different geographical regions. There is now no common coverage beam and it is necessary to interconnect the beams in some fashion on-board the satellite. One way is to switch the signals from beam to beam in a cyclical manner. An array of switches on board the satellite, controlled by a mini-computer is necessary. The information bits are now organized on a one burst per destination basis and the on-board switching will route the bursts appropriately. Because frequency reuse is now feasible due to the isolation provided by the antenna spot beams illuminating different regions with or without further isolation due to polarization, the system is now a space division multiple access/satellite switched scheme over and above the TDMA scheme. Hence the system is called SDMA/SS-TDMA. In the SDMA/SS-TDMA network, the earth station burst timing now has to lock on also to the on-board switching system. One way of achieving this is to have a synchronizing interval during which the up-link signal is looped back to the down-link beam of the same region. Each earth station would send a unique code word of known length called the "sync burst" through this interval known as the "sync window". If the timing is correct, the code word would be looped back to the originating station untruncated. If the timing is incorrect one end or the other of the code word would be chopped off, or the whole word may be missing. Automatic search techniques are then used to re-adjust the earth station timing so that correct timing is achieved. Other synchronizing schemes are possible[19].

From the above it can be seen that the SDMA/SS-TDMA scheme is an extension of the "conventional" TDMA system. This gives flexibility in system evolution as, in general, the same techniques and earth segment facilities with appropriate modifications can be used if an existing satellite system uses TDMA and a replacement satellite adopts spot beams and on-board switching.

4.2 SATELLITE AND EARTH STATION CONFIGURATIONS

Three satellite system models with SDMA/SS-TDMA are presented, with block diagrams of the satellite and earth station configurations.

In 3.5, three rf channel allocation plans are presented for the on-board switched system corresponding to the three antenna beam plans. It is therefore necessary to examine typical composite space and earth segment configurations corresponding to these frequency plans. The models to be investigated are numbered as in Table 4-1.

Table 4-1

MODEL NUMBER	FREQUENCY PLAN	BEAM PLAN
U1	S1	A
U2	S2	B
U3	S3	C

A brief description of each of the models ensues.

Model U1

In this model, 8 very narrow spot beams are used. There are 8 transponders, each with a bandwidth of 125 MHz. This gives adequate system capacity for the beams with the densest traffic and no dual polarization frequency reuse is necessary. The overall system design can therefore be made very simple. The satellite and earth station configurations are illustrated in Figures 4-1(a) and 4-1(b) respectively.

Model U2

This model corresponds to Beam Plan (B) in which there are 14 uniform elliptical beams covering the whole of Canada. However, the locations of the major traffic centres are such that only 8 out of these 14 beams are sufficient. The system configuration is therefore designed based on 8 beams.

In this model, from considerations of interference as detailed in 3.5, the transponder bandwidth is constrained to 62.5 MHz. It is therefore necessary to have dual polarization frequency reuse in order to obtain adequate system capacity for the beams with very dense traffic. This means that, apart from the redundant equipment, the system configurations for both the space and earth segments are duplicates of two identical subsystems, one for vertical* and one for horizontal* polarization. The block diagrams of Model U2 for the space and earth segments are given in Figures 4-2(a) and 4-2(b) respectively.

Model U3

This model corresponds to Beam Plan (C), which has only 4 beams covering all the major traffic centers. Again, from considerations of interference, spectrum availability, beam size and traffic requirements in the various beams, each beam would need to employ dual polarization frequency reuse, with a transponder bandwidth of 125 MHz. The block diagrams are given in Figures 4-3(a) and 4-3(b) for the space and earth segments respectively.

In all the three models, up-link power control, as explained in 4.3.3, is used. To cater for a propagation reliability of 99.99 percent of the time, the rain attenuation data presented in the previous study[14] indicates that a 10 dB range of control is needed. This control is achieved by receiving a satellite beacon signal at the earth station and calibrating this signal against propagation loss.

In a time division multiple access system, as is the case in all the models considered above, demand assignment can be easily implemented by appropriate control of the burst length from each of the earth stations. The increase in earth station equipment complexity and cost is not significant. However, there is a need to interface with the traffic switching center in regard to the changing patterns of satellite channel capacity demands, but it is outside the scope of the present study to examine this interface aspect.

**Note*: The terms vertical and horizontal have no particular meanings except to indicate two orthogonal axes in space.

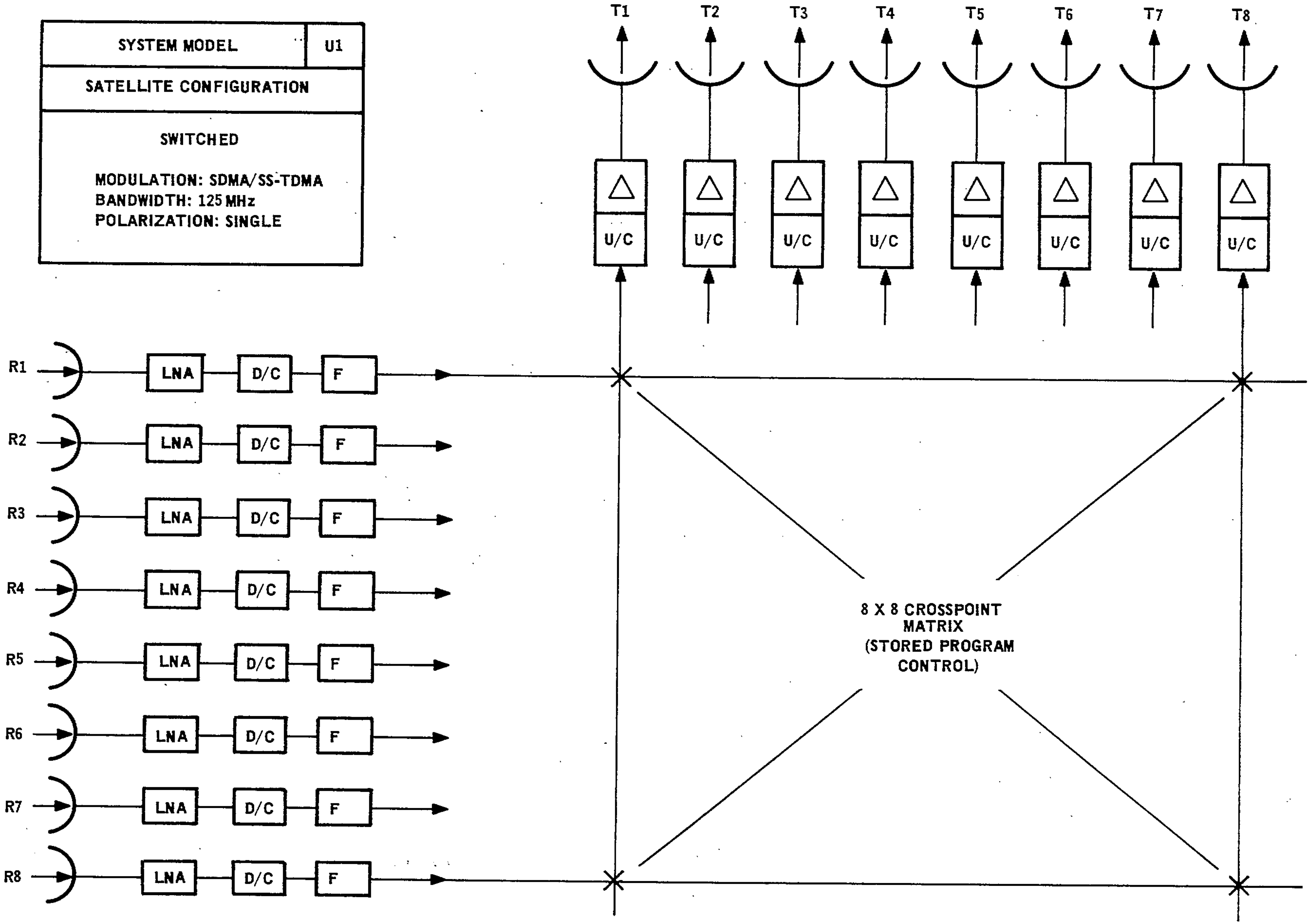

Figure 4-1(a) Model U1 - Satellite Configuration for SDMA/SS-TDMA

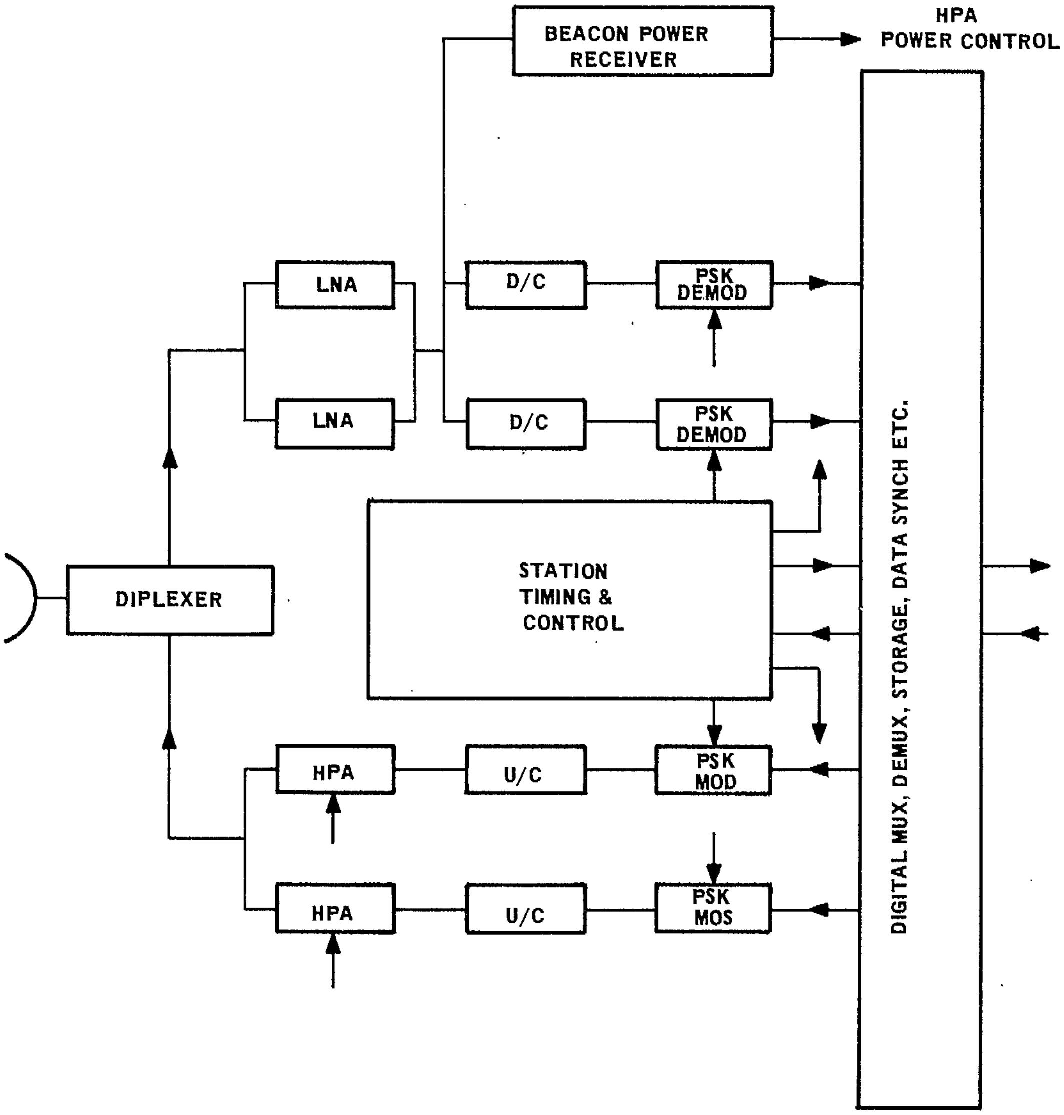

Figure 4-1(b) Model U1 - Earth Station Configuration for SDMA/SS-TDMA

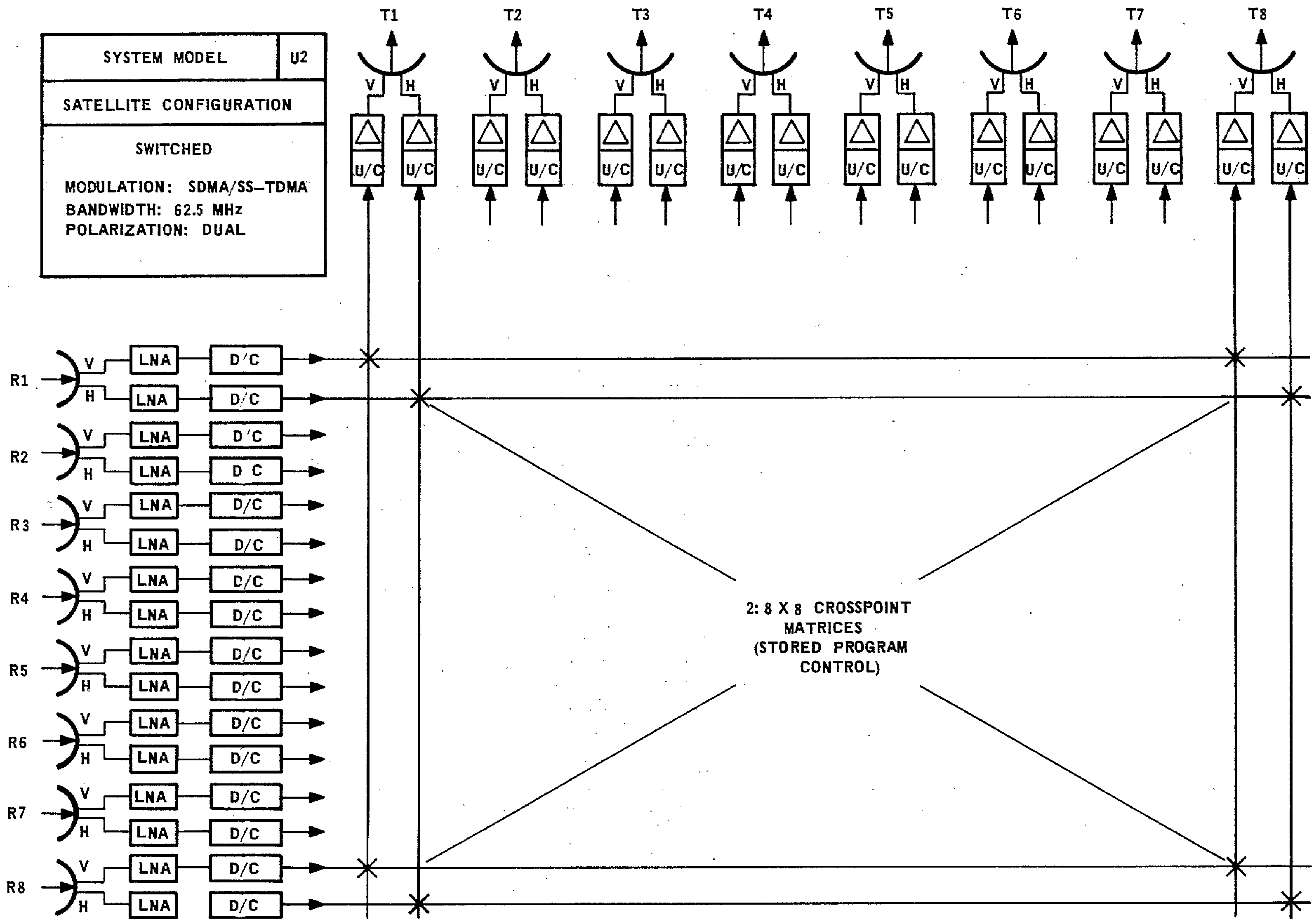

Figure 4-2(a) Model U2 - Satellite Configuration for SDMA/SS-TDMA

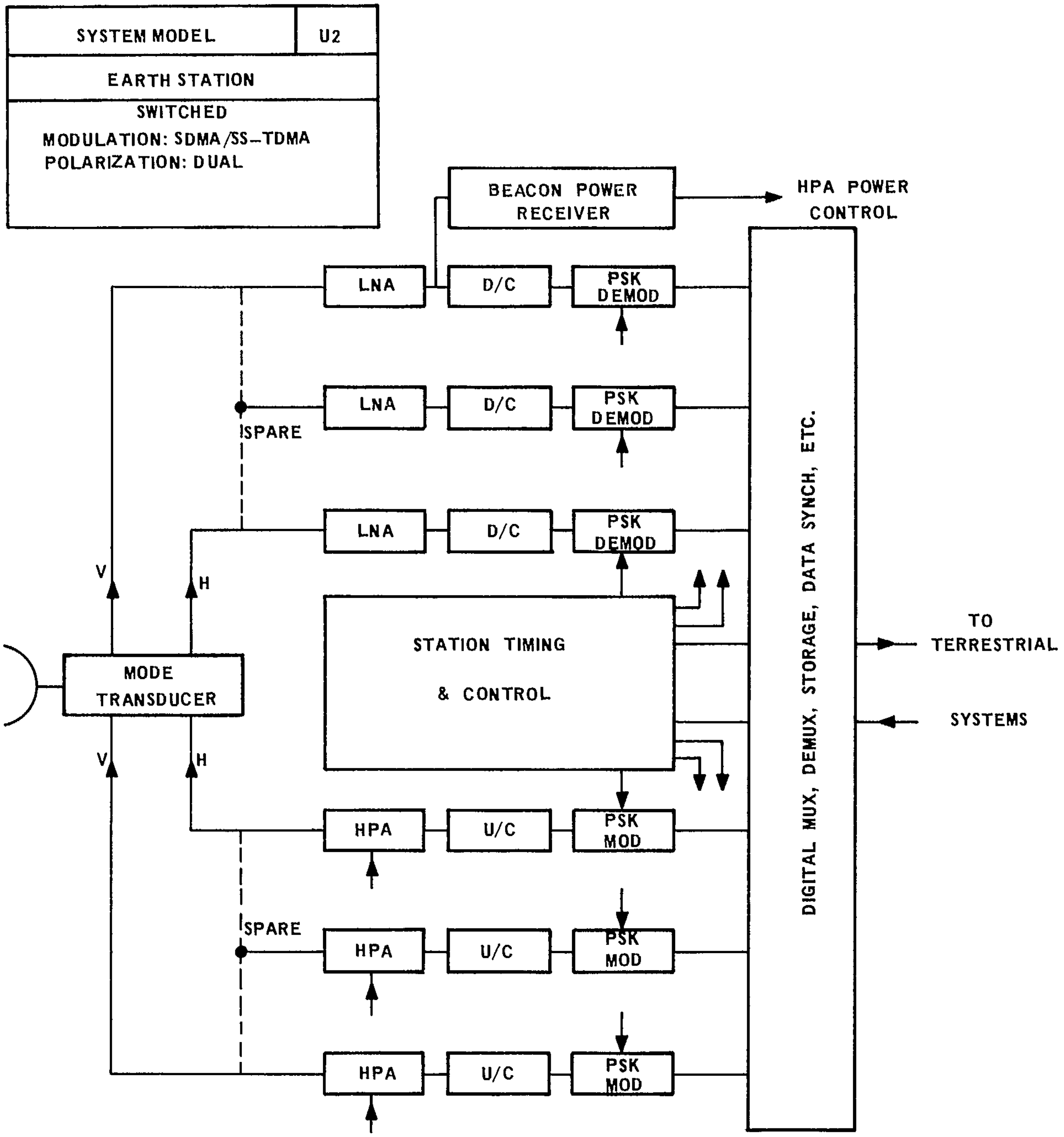

Figure 4-2(b) Model U2 - Earth Station Configuration for SDMA/SS-TDMA

SYSTEM MODEL	U3
SATELLITE CONFIGURATION	
SWITCHED MODULATION: SDMA/SS-TDMA BANDWIDTH: 125 MHz POLARIZATION: DUAL	

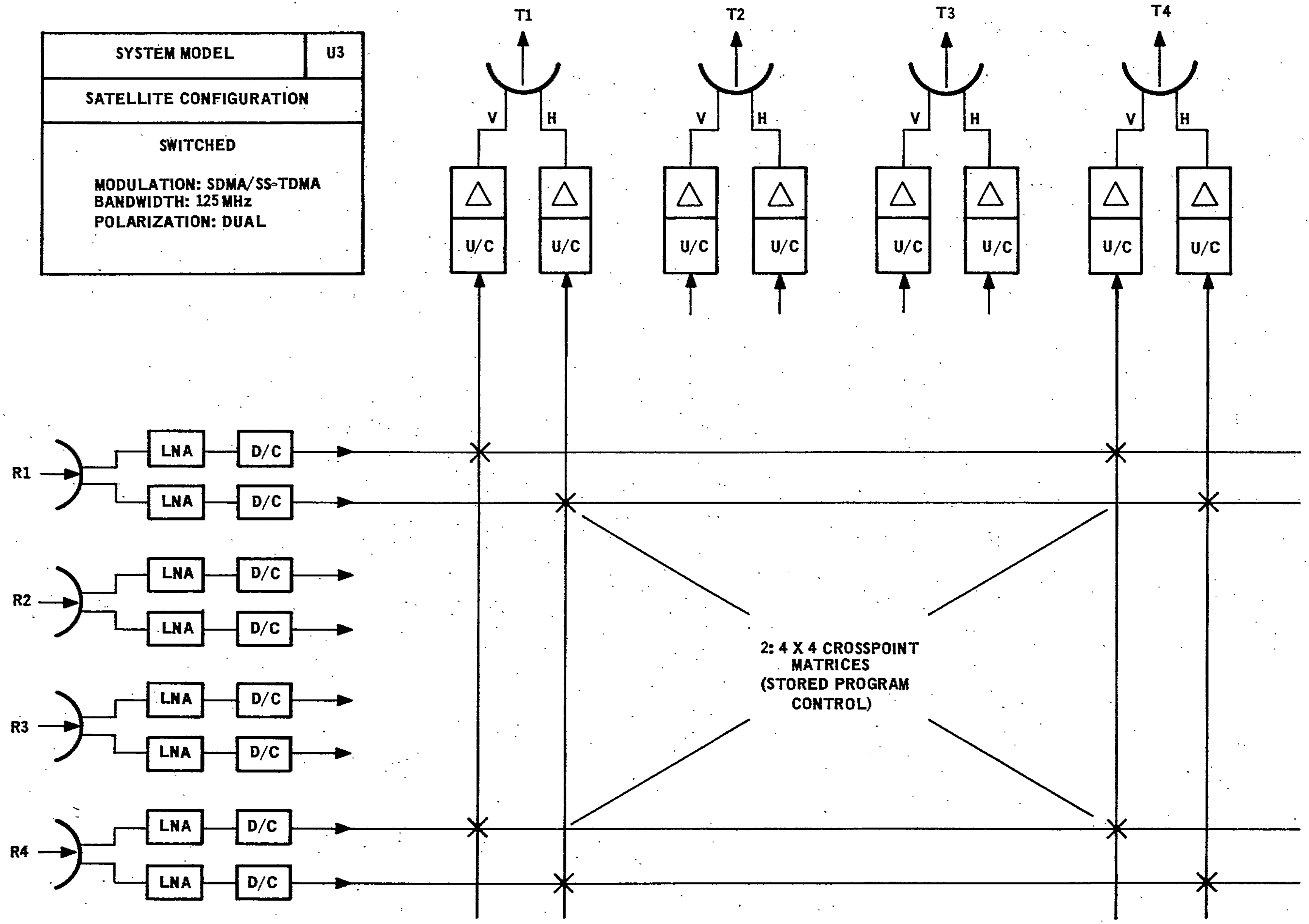

Figure 4-3(a) Model U3 - Satellite Configuration for SDMA/SS-TDMA

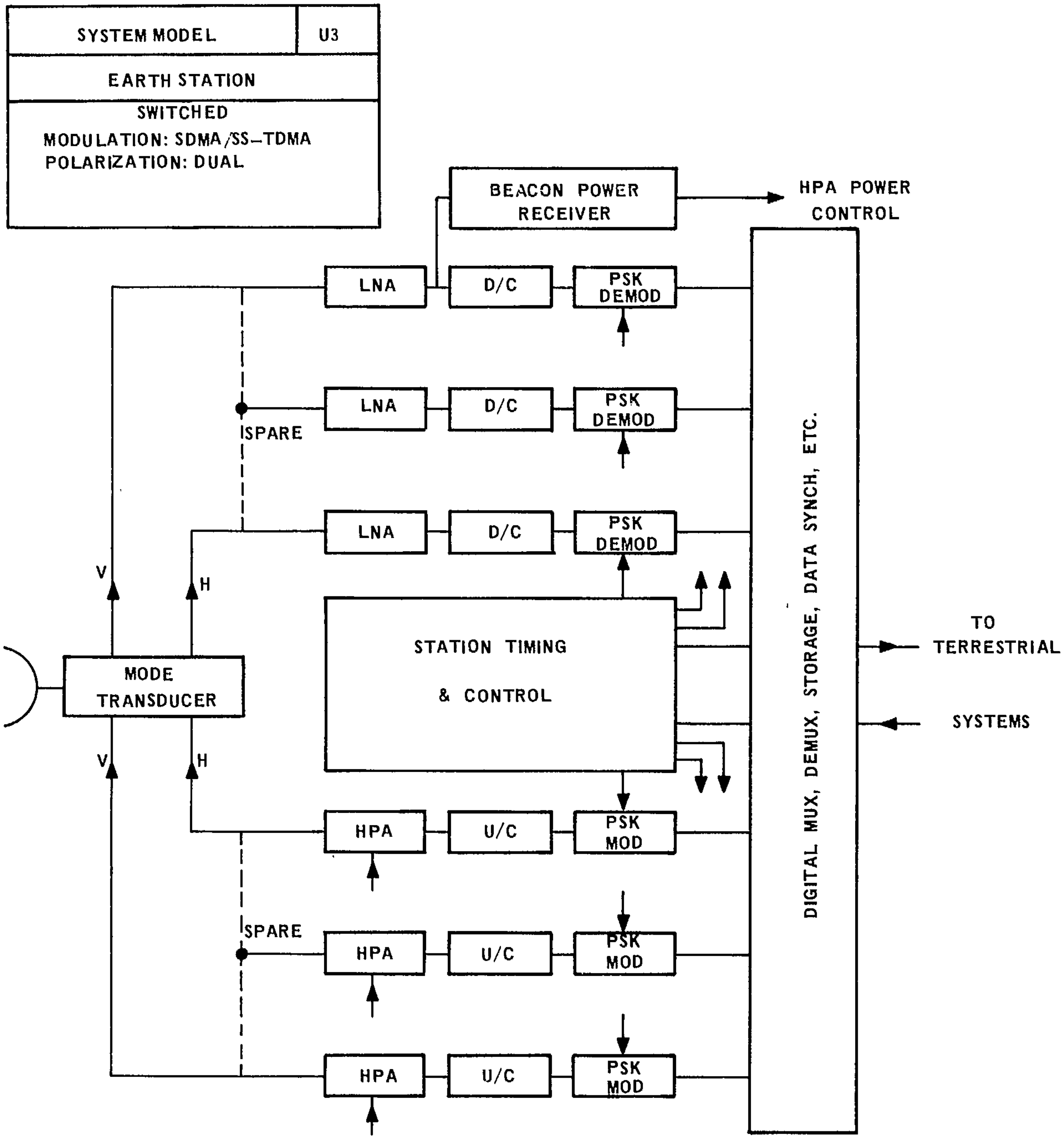

Figure 4-3(b) Model U3 - Earth Station Configuration for SDMA/SS-TDMA

4.3 TECHNICAL SYSTEM TRADE-OFF RELATIONSHIPS

The technical system trade-off equations are derived.

This section considers the technical factors and system equations that inter-relate the up-link and down-link power, bandwidth, noise and interference of a digital satellite system. System capacity is linked to bandwidth through a bandwidth utilization index (N/B).

* * *

4.3.1 CHOICE OF MODULATION

It is shown that 4-phase coherent phase shift keying is the optimum choice.

In a digital satellite communication system, it is well established that phase shift keying (PSK) with coherent detection is the most efficient in power and bandwidth utilization. Phase shift keying can be 2-phase (binary), 4-phase, 8-phase and so on. In a system disturbed purely by Gaussian noise, the theoretical energy per bit to noise density ratios (E/N_o) for any given bit error rate are identical between the 2- and 4-phase systems. For a bit error rate of 1×10^{-4}, the E/N_o required is 8.4 dB for either modulation. The 4-phase system however uses only half the bandwidth. In such a case there is no merit in choosing a 2-phase system. However, in a practical system, there is an implementation margin to be added. Here the 2-phase system has a slight advantage, say 0.5 dB. Further, in an interference environment, as is the case with multiple spot beams and frequency reuse, the 2-phase system again is more resilient, as evident from Table 4-2 which shows the increase in carrier-to-noise ratios needed to maintain an error rate of 1×10^{-4} in the presence of various degrees of interference. Table 4-2 is derived from 2.2.

Table 4-2 Increase in $(\frac{C}{N_oB})$ due to Presence of Interference, at Bit Error Rate of 1×10^{-4}

(Two Interference Entries)

	INCREASE OF $(\frac{C}{N_oB})$ FOR $\frac{C}{I}$ = VALUES BELOW			
	$\frac{C}{I} = \infty$ dB	$\frac{C}{I} = 18$ dB	$\frac{C}{I} = 15$ dB	$\frac{C}{I} = 12$ dB
2ϕ CPSK	0	0.3 dB	1.0 dB	2.0 dB
4ϕ CPSK	0	1.5 dB	2.7 dB	4.7 dB

It can be seen that for a typical C/I = 15 dB, the 2-phase system has an advantage of about 1.6 dB.

Although the above advantages in power utilization of the 2-phase system are significant, frequency spectrum is a valuable resource and the doubling of system capacity between 4- and 2-phase cannot really be compensated for by the slight advantage in system power. The most compelling argument against the 2-phase technique in the current system configurations, however, is that the 2-phase PSK system does not give adequate channel capacity in the dense traffic antenna beams illuminating Toronto and Montreal.

This leaves for consideration the 4- and 8-phase schemes. The 4-phase system satisfies the traffic capacity requirements and hence is a prime candidate. In the previous study[14], a direct comparison was made between the 4- and 8-phase schemes for major traffic communications. It was shown, from Figures 8-14 and 8-15 of the report of that study[14] that given equal bandwidth, the 8-phase system has a 50 percent more capacity than the 4-phase system. However, given equal power, (but no bandwidth constraint), the 4-phase system has 150 percent more capacity than the 8-phase modulation. In the particular models studied here, since the 4-phase system gives adequate capacity within the bandwidths available, it is therefore obvious that 4-phase PSK is the better choice. There is no need to consider higher levels of phase modulation such as 16-phase, as the power penalty is even more severe. An added important reason to choose the 4-phase scheme is its better resilience to interference, than the higher levels of phase modulation. Hence the 4-phase PSK is the optimum choice.

In the presentation in the next few subsections, the 2-phase system will be included for comparison purposes.

4.3.2 PCM-CPSK SYSTEMS EQUATIONS

The coherent PSK system equations under power and bandwidth limitations are combined to derive the critical values of carrier-to-noise ratios. The concept of bandwidth utilization index is introduced.

There are two basic equations which govern a digital radio link using coherent phase shift keying (CPSK). These are:

a) Under Power Limitation

$$10 \log_{10} R = C/T - E/N_0 - k - M_i \qquad (4.1)$$

b) Under Bandwidth Limitation

$$R = \frac{B \log_2 (L)}{1.2} \qquad (4.2)$$

where R = transmission bit rate (b/s)

C/T = system carrier to noise temperature ratio (dB)

E/N_0 = energy per bit to noise density (dB)

k = Boltzman's constant dBW/°K

M_i = implementation margin which accounts for the difference between theory and actual practice, and includes the effects of intersymbol interference, TWT non-linearities, etc. (dB)

L = number of phase positions in phase shift keying (2, 4, 8, etc)

1.2 = bandwidth allowance factor.

For a coherent PSK system in the presence of white gaussian rf noise, the theoretical value of E/N_0 for any particular bit error rate is well known[23]. In a digital radio system, for design purposes, the threshold value of bit error rate is taken as 1×10^{-4}. This gives approximately a noise of 50 000 pWop in a telephone channel[20]. This noise is also roughly the threshold noise of an FDM-FM system and is therefore a reasonable criterion. Table 4-3 gives the values of E/N_0 for a bit error rate of 1×10^{-4} for coherent PSK systems with 2- and 4-phase modulations. The same table also gives the implementation margins M_1. These margins were estimated from the measured results of tests conducted by the Kokusai Denshin Denwa Company (Japan) through Intelsat III[21].

COMSAT[22] (U.S.A.) has published certain figures which differ only fractionally from those in Table 4-3. The figures in Table 4-3 may therefore be used with reasonable confidence.

Table 4-3

FOR ERROR RATE OF 1×10^{-4}		
Lϕ CPSK	E/N_0 (dB)	M_i (dB)
2ϕ CPSK	8.4	3.5
4ϕ CPSK	8.4	4.0

Referring to equations (4.1) and (4.2), the optimum point of operation is when the transmission bit rates given under power limitation and bandwidth limitation coincide. That is:

$$10 \log_{10} \frac{B \log_2 L}{1.2} = C/T - E/N_0 - k - M_i,$$

or

$$\frac{C}{N_0 B} = \frac{C}{T} - k - 10 \log_{10} B = 10 \log_{10} \frac{\log_2 L}{1.2} + \frac{E}{N_0} + M_i. \quad (4.3)$$

Substituting the values of L, E/N_0 and M_i from Table 4-3 into equation (4.3) gives the "critical" values of carrier-to-noise ratio $C/(N_0B)$. These are presented in Table 4-4.

Table 4-4

MODULATION	CRITICAL C/N_oB (dB)
2ϕ CPSK	11.1
4ϕ CPSK	14.6

A system operating at these critical values of $C/(N_oB)$ would be using both power and rf bandwidth optimally. Below the critical value, the operation would be in the "power limited" region. Coversely, above the critical value, the operation would be in the "bandwidth limited" region. A practical system cannot, however, operate under the "power limited" region for the following reasons:

a) The system is operating at the threshold bit error rate of 1×10^{-4} at all times. This is not acceptable.

b) Any system fade must be immediately accompanied by a corresponding reduction in system bandwidth in order to maintain the same $C/(N_oB)$. This is not practical. Further, even if this were practical, a reduction in bandwidth means a reduction in transmission bit rate. This is not acceptable for a normal communications system.

Hence, all digital systems must operate in the "bandwidth limited" region with adequate fade margins to give the desired service availability.

Since the digital system must operate under a "bandwidth limited" region, there is a fixed relationship between system capacity and bandwidth. This is best expressed as a bandwidth utilization index in terms of "number of telephone channels per MHz bandwidth, (N/B)"*.

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*This method of expressing bandwidth utilization efficiency was, to the best of our knowledge, first used in an internal Bell-Northern Research study.

In a terrestrial PCM system, the standard now adopted is 64 kb/s per voice channel of 4 kHz. However, in the multiplexing process, framing and synchronizing (stuffing) bits are added. It is envisaged that for the satellite system under study, the Bell T1 (24 channels) and T2 (96 channels) PCM systems would be largely used as the basic systems for the digital telephone network. For the T2 svstem the average is 66 kb/s per 4 kHz voice channel. For a time division multiple access (TDMA) satellite network, with or without on-board switching, additional bits are necessary for the burst synchronization preamble word, guard time, etc. These would raise the bit rate per voice channel to approximately 67.5 kb/s.

The system transmission bit rate would therefore be:

$$R = 67.5N \times 10^3 \text{ b/s.} \tag{4.4}$$

Substituting equation (4.4) into equation (4.2) gives the bandwidth utilization index as:

$$\left(\frac{N}{B}\right) = 12.35 \log_2 (L). \tag{4.5}$$

where L = Level of phase modulation
(L = 2, 4, 8, etc.)

N = Number of voice channels

B = Bandwidth in MHz

Table 4-5 gives the values of (N/B) for 2-, 4- and 8-phase modulations.

Table 4-5 Bandwidth Utilization Index

MODULATION	CHANNEL PER MHz (N/B)
2φ CPSK	12.35
4φ CPSK	24.70
8φ CPSK	37.05

From Table 4-5, it can be seen that an antenna beam with total transponder bandwidth of 125 MHz would, with 4ϕ CPSK modulation, yield a total capacity of:

$$24.70 \times 125 = 3087 \text{ voice channels.}$$

This is approximately the peak system capacity required of each of the Toronto and Montreal spot beams which carry the densest traffic streams.

* * *

4.3.3 UP-LINK AND DOWN-LINK SYSTEM EQUATIONS

Up-link power control is shown to be advantageous. The basic up-link and down-link system power trade-off equations are derived.

In the multiple spot beam and frequency reuse environment, both up-link and down-link interferences are important considerations. However, the carrier-to-interference ratios are affected differently by up-link and down-link fades. In an up-link fade, it is generally the wanted carrier that suffers the fade. The interference comes from widely dispersed beams and therefore it is reasonable to expect negligible coincidence in fades. The interference levels therefore would remain sensibly constant. Hence the C/I of the particular up-link beam suffers. Increasing the normal (no fade) up-link power from every beam does not help, as this leaves the net mutual C/I unchanged. The only practical way is to have up-link power control. In the previous study[14] at 15 GHz, the fade for 0.01 percent of the time is given as approximately 10 dB. Up-link power control therefore need only be no more than a range of 10 dB. With this up-link power control, both the up-link carrier-to-thermal noise ratio and carrier-to-interference ratio may be taken as sensibly constant under all up-link fade conditions.

A fade in the down-link attenuates both the carrier and interference equally. Hence a down-link fade does not affect the C/I. The fade only affects the down-link carrier-to-thermal noise ratio and it is necessary to design the satellite down-link power with adequate margin to overcome this fade. From reference 14, the down link fade for 0.01 percent of the time is given as 6 dB, at a frequency of 12 GHz. Hence the down-link fade margin will be taken as 6 dB.

Before the system equations can be derived, there is yet one more relationship that needs to be established. This is the ratio of up-link carrier-to-noise temperature $(C/T)_u$ to down-link carrier-to-noise temperature $(C/T)_d$. The most critical down-link power requirement is during a 6 dB down-link fade. Under this circumstance, it is generally accepted that the up-link thermal noise contribution should be negligible. It is therefore considered reasonable that in such a 6 dB down-link fade, the up-link $(C/T)_u$ should be at least 10 dB better. That is to say, under median conditions, $(C/T)_u$ is better than $(C/T)_d$ by 10-6 dB, or 4 dB.

From the above considerations, the system link equations may be written as follows:

$$(C/T)_d = EIRP_{sat} - L_d + G/T_{earth} \quad dBW/°K \qquad (4.6)$$

$$(C/T)_u = EIRP_{earth} - L_u + G/T_{sat} \quad dBW/°K \qquad (4.7)$$

$$(C/T)_d = (C/T)_u - 4 \quad dBW/°K \qquad (4.8)$$

where

$(C/T)_d$ = Carrier-to-noise temperature, down-link

$(C/T)_u$ = Carrier-to-noise temperature, up-link

$EIRP_{sat}$ = EIRP, satellite

$EIRP_{earth}$ = EIRP, earth station

G/T_{sat} = G/T, satellite

G/T_{earth} = G/T, earth station

L_u = Propagation loss, up-link = 208.2 dB

L_d = Propagation loss, down-link = 206.7 dB

Rearranging equations (4.6) through (4.8) and putting $L_u - L_d = 1.5$ dB,

it follows that:

$$EIRP_{earth} = (EIRP_{sat} + G/T_{earth}) - G/T_{sat} + 5.5 \qquad (4.9)$$

From equation (4.6), it is easy to show that

$$\frac{C}{N_oB}_d = (C/T)_d - k - 10 \log (B) = EIRP_{sat} - L_d + G/T_{earth}$$
$$- k - 10 \log (B)$$

or:

$$(EIRP_{sat} + G/T_{earth}) = \frac{C}{N_oB}_d + 10 \log (B) - 21.9 \text{ dB} \qquad (4.10)$$

where

$\frac{C}{N_o B_d}$ = carrier-to-thermal noise, down-link (non fade condition)

k = Boltzman's constant = 228.6 dBW/°K

B = RF Bandwidth, Hz.

Equations (4.9) and (4.10) are the two basic system equations that link the power trade-off between the space and earth segments. To apply these equations to any particular satellite system model, it is necessary to assign values to the terms G/T_{sat}, $(C/N_oB)_d$ and bandwidth (B) associated with that model. In section 3, both G/T_{sat} and B are given for each model. The values of $(C/N_oB)_d$ may be derived as below.

Table 4-4 gives the "critical (C/N_oB)". To these must be added the increase in (C/N_oB) as given in Table 4-2, due to the presence of interference. The total may be termed the overall "threshold (C/N_oB)" and is given in Column I of Table 4-6. This threshold (C/N_oB) would be reached during a 6 dB fade in the down-link. Since under such a fade condition, the up-link carrier-to-thermal noise would be 10 dB higher than that of the down-link and would degrade the overall noise by about 0.4 dB, therefore the down-link (C/N_oB) under faded condition would be 0.4 dB better than the threshold values of Column I. These faded down link (C/N_oB) values are given in Column II. The non-faded down-link $(C/N_oB)_d$ are derived from Column II by simply adding the 6 dB margin and are given in Column III. The values in Column III are those used in the system trade-off.

Columns IV and V are up-link (C/N_oB) and total overall (C/N_oB) under non-fade conditions. These are simply additional information and are not used in the system trade off.

Inserting the values of $(C/N_oB)_d$ of Column III into equation (4.10), it is now possible to express the bracketed term ($EIRP_{sat}$ + G/T earth) of this equation as a constant for constant bandwidth. Tables 4-7a and 4-7b are tabulations of this bracketed term for 4ϕ CPSK and 2ϕ CPSK respectively.

For any particular satellite model under consideration, both the modulation (i.e., 4ϕ CPSK) and the transponder bandwidth are predetermined from the constraints of capacity need, available bandwidth, interference, etc, as discussed earlier. Hence, for each model, the term ($EIRP_{sat}$ + G/T_{earth}) is fixed. However, within this fixed bracketed quantity, there is a cost trade-off between satellite power and earth station G/T. This trade-off will be performed in section 7. A possible range of $EIRP_{sat}$ versus G/T_{earth} for a system with 4ϕ CPSK and interference ratio C/I = 15 dB is

Table 4-6 Carrier-to-Noise Ratios for Various Fading and C/I Conditions

	COLUMN I Threshold $(\frac{C}{N_oB})$				COLUMN II Faded Down Link $(\frac{C}{N_oB})$				COLUMN III Non-Faded Down-Link $(\frac{C}{N_oB})$				COLUMN IV Up-Link $(\frac{C}{N_oB})$ (Non-Variant)*				COLUMN V Total Non-Faded Overall System $(\frac{C}{N_oB})$			
Mod C/I =	∞dB	18dB	15dB	12dB	∞dB	18dB	15dB	12dB	∞dB	18dB	15dB	12dB	∞dB	18dB	15dB	12dB	∞dB	18dB	15dB	12dB
2φCPSK	11.1	11.4	12.1	13.1	11.5	11.8	12.5	13.5	17.5	17.8	18.5	19.5	21.5	21.8	22.5	23.5	16	16.3	17.0	18.0
4φCPSK	14.6	16.1	17.3	19.3	15.0	16.5	17.7	19.7	21.0	22.5	23.7	25.7	25.0	26.5	27.7	29.7	19.5	21.0	22.3	24.2

*Note: Up-link has power control.

Table 4-7a Values of $(EIRP_{sat} + G/T_{earth})$ for 4ϕ CPSK with 6 dB Fade Margin and Threshold $P_E = 1 \times 10^{-4}$

TRANSPONDER BANDWIDTH B MHz	$(EIRP_{sat} + G/T_{earth})$ dBW			
	$\frac{C}{I} = \infty$	$\frac{C}{I}$ = 18 dB	$\frac{C}{I}$ = 15 dB	$\frac{C}{I}$ = 12 dB
15.625	71.1	72.6	73.7	75.7
31.25	74.1	75.6	76.7	78.7
62.5	77.1	78.6	79.7	81.7
125.0	80.1	81.6	82.7	84.7

Table 4-7b Values of $(EIRP_{sat} + G/T_{earth})$ for 2ϕ CPSK with 6 dB Fade Margin and Threshold $P_E = 1 \times 10^{-4}$

TRANSPONDER BANDWIDTH B MHz	$(EIRP_{sat} + G/T_{earth})$ dBW			
	$\frac{C}{I} = \infty$	$\frac{C}{I}$ = 18 dB	$\frac{C}{I}$ = 15 dB	$\frac{C}{I}$ = 12 dB
15.625	67.6	67.9	68.6	69.6
31.25	70.6	70.9	71.6	72.6
62.5	73.6	73.9	74.6	75.6
125.0	76.6	76.9	77.6	78.6

plotted in Figure 4-4. In the system trade-off in subsequent sections, a system with C/I of 15 dB will be used even though the interference considerations of section 3 indicate that an overall C/I of 18 dB is achievable. It is preferable to take a conservative approach, in view of the lack of precise knowledge of many factors at this point in time.

Referring to equation (4.9), it will be noted that the term G/T_{sat} is constant for any particular antenna beam plan, and its value is given in Table 3-9. It is now obvious that the earth station radiated power ($EIRP_{earth}$) is constant for each model, regardless of the outcome of the system trade-off of satellite radiated power and earth station G/T. This surprising fact simplifies system trade-off. It should be noted that this $EIRP_{earth}$ is that value under normal non-fade conditions. Under an up-link fade, an up-link power control of up to 10 dB is required. The cost of high-power amplifiers with this reserve power is taken into consideration in the economic trade-off of chapter 7.

The above completes the basic technical considerations of a SDMA/SS-TDMA satellite system. The comparison between the models of this system and between the SDMA/SS-TDMA technique and the non-switch satellite configurations will be done in chapters 7 and 8.

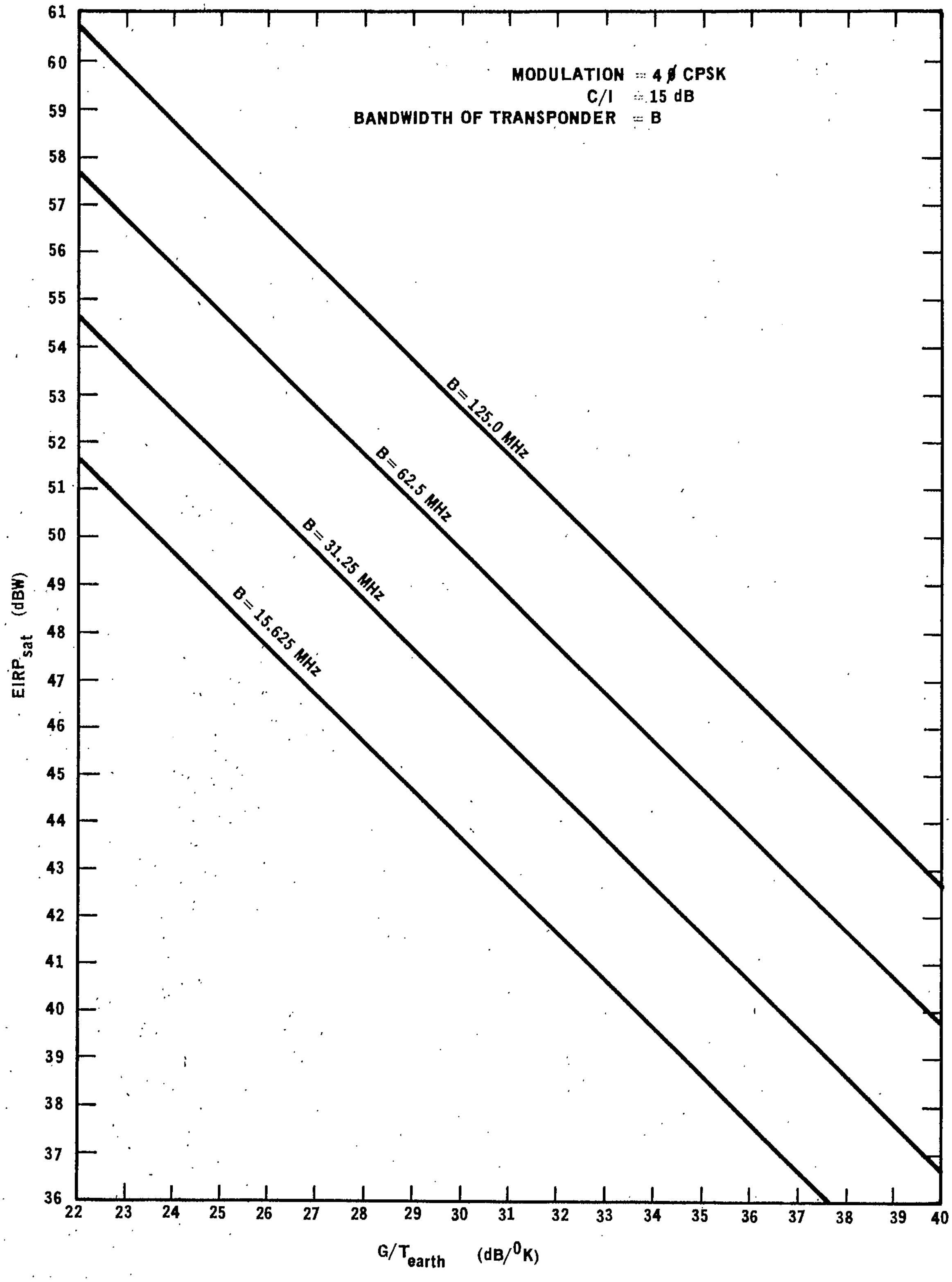

Figure 4-4 Satellite Radiated Power vs Earth Station G/T

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5. MAJOR TRAFFIC SATELLITE SYSTEM WITHOUT ON-BOARD SWITCHING

Three modulation schemes will be considered, namely PCM-PSK-TDMA, FDM-FM-FDMA and Single-Channel-Per-Carrier FM. The principles of operation, the space and earth segment configurations and technical parameters governing the system trade-off will be presented.

In chapter 4, the satellite models considered use on-board time division switching for interconnection from antenna beam to antenna beam. In this chapter however, the models to be examined will have no such switching. For routing the signals between the antenna beams, physical interconnections, either permanent or time patched through telecommand, will be used. The satellite transponders will in effect be transparent to the communications signals. Modulation can therefore be of any type consistent with the bandwidth and power available. In this chapter both digital and analog modulations will be examined. In the digital schemes, as has been indicated in chapter 4, pulse code modulation with 4-phase PSK will be used since this is optimum for the capacity needs and available bandwidth. Multiple access could be either FDMA or TDMA. In the case of analog modulation the two schemes considered will be frequency division multiplex with frequency modulation and frequency division multiple access (FDM-FM-FDMA) and single-channel-per-carrier frequency modulation. The FDM-FM-FDMA scheme is examined because it is the technique used today. Both the conventional system as well as a system with speech compression will be evaluated. Speech compression is carried out through the lincompex technique. Although the speech compression scheme considered here has not been fully proven, the system concept appears valid and the use of such technique appears to hold potential for better bandwidth and power utilization. With speech compression and its better tolerance to higher system interference and noise it is feasible to adopt the single-channel-per-carrier technique for major traffic communications. Hence, the single-channel-per-carrier system with speech compression will be investigated.

The following modulation techniques will be considered in sections 5.1, 5.2 and 5.3 respectively:

- PCM-PSK with both FDMA and TDMA, depending on beam configuration.
- FDM-FM-FDMA, with and without speech compression.
- Single-Channel-Per-Carrier, with speech compression.

Space and earth segment configurations optimum to each modulation technique will be presented.

5. Major Traffic Satellite System Without On-Board Switching

5.1 PCM-PSK SYSTEM

The principle of operation, the space and earth segment configurations and technical system parameters will be presented.

In the non-switch satellite models considered in this chapter, the satellite segment is of course transparent to the modulation technique. The principle of operation of a digital network with multiple satellite spot beams, the space and earth segment configurations and the technical system parameters are discussed in 5.1.1, 5.1.2 and 5.1.3.

* * *

5. Major Traffic Satellite System Without On-Board Switching

5.1 PCM-PSK System

5.1.1 PRINCIPLE OF OPERATION

The principle of operation of a PCM-PSK-TDMA network using multiple satellite spot beams will be explained.

The PCM-PSK-TDMA system postulated for the non-switched satellite models is basically the same as the "conventional" TDMA systems described briefly in chapter 4, that is to say, each signal burst would contain the preamble word followed immediately by the information bits. However, unlike any TDMA system that uses a single coverage beam, the multiple spot beam configurations considered here require a slightly different TDMA synchronizing technique. In the conventional single beam TDMA system, each transmitting earth station can receive back via satellite its own preamble word in every burst, plus the preamble word or station identification code of a reference station. Synchronization is achieved by measuring the time difference between the station's own identification code word looped back via the satellite and the identification code of the reference station. If the timing is found to be incorrect, the station timing and control equipment would automatically adjust the transmit burst timing until correct synchronization is achieved. This technique in effect is equivalent to each earth station making frequent periodic satellite range measurements (say three times per second) and automatically adjusting the burst timing to compensate for the satellite movement or change in propagation. This scheme, as is obvious, is non-centralized controlled. It is a technique well proven in Intelsat experiments[18].

In a multiple beam satellite system, if the up-link signals from one beam are routed to the down-links of all the other beams except to the down-link of the beam which illuminates the transmitting earth station, then that earth station will never see its own signal bursts via the satellite. This results in some difficulty in TDMA synchronizing. One possible scheme is to measure the satellite movement through ranging techniques from three or more widely separated earth stations. From this information, the distance from the satellite to each earth station is computed at a computer centre and each earth station is instructed via a data link to time the signal bursts accordingly. This scheme would be centralized controlled. As far as is known, this technique has not yet been tried out experimentally. An alternative scheme would be to interconnect an rf channel from the up-link to the down-link of the same beam. This interconnection is justified, at any rate in the case of beams illuminating more than one earth station, and communications are assumed needed between these stations.

This is especially true in the case of the larger beam models such as Beam Plan (C). In this alternative scheme one earth station would act as a reference station as in the conventional TDMA system and hence the reference station identification code can be received by all. In addition, each earth station can now loop back via the rf channel mentioned above, measure the round trip time of its own preamble code word and compare against the reference station code word. Synchronization can now be achieved as already described. This scheme would be non-centralized controlled.

It is beyond the scope of this study to delve into the pros and cons of a centralized system versus non-centralized system. From the cost point of view the difference in cost may not be significant, considering that the network has only 12 earth stations. Since the satellite configurations and beam plans considered do require at least one rf channel interconnected between the up-link and down-link of the same beam, and since the non-centralized synchronization scheme is closest in concept to the conventional TDMA scheme that has been tested out by Intelsat, this non-centralized scheme will be postulated for the present study.

The PCM-PSK-FDMA scheme, if used, is straight forward and presents no multiple access timing and synchronizing problems. The PCM-PSK signals are simply routed from point to point in a continuous stream.

5.1.2 SATELLITE AND EARTH STATION CONFIGURATIONS

The space and earth segment configurations will be presented.

In the PCM-PSK-TDMA system for the non-switched satellite configuration, three space and earth segment models will be presented, corresponding to the three frequency and antenna beam plans developed in chapter 3. These will be designated as models W1, W2 and W3 corresponding to frequency plans N1, N2 and N3 respectively. The block schematic diagrams are given in Figures 5-1, 5-2 and 5-3.

The significant difference between the switched and non-switched satellite configurations is that in the non-switched system, each satellite receive point (up-link) must somehow be strapped to each satellite transmit point (down-link). There are many ways of doing this and it is not practicable to evaluate all possible permutations and combinations. Hence a certain degree of engineering judgement must be exercised. From the point of view of system feasibility and economic viability, the outcome is not expected to significantly alter whatever technique is chosen. When it comes to actual hardware design, the matter may be different. The latter is not within the scope of this study.

In Figures 5-1, 5-2 and 5-3, the satellite receive and transmit interconnections are achieved via i-f filtering and combining as appropriate at i-f frequencies. This technique calls for a minimum of active electronic component on board with many of these components of common type, thus simplifying the redundant equipment requirement. Active components are assigned a redundancy of 30 percent in the cost evaluation in chapter 7. The actual arrangement of filtering is dictated by the associated frequency plans of chapter 3. The earth segment configurations are straightforward and conventional and redundant equipments are indicated. The degree of protection shown is considered reasonable.

It is to be noted that in models W1 and W2 as shown in Figures 5-1 and 5-2 respectively, multi-carrier operation of the satellite transponder TWT is postulated. With eight antenna beams carrying major route traffic in these models, and each beam having eight rf channels, it is not practicable to have single-channel-per-transponder configuration. Therefore, despite the back-off power penalty, multi-carrier operation is considered to be the only practical solution. With multi-carrier operation, there is no virtue in TDMA in these two models and full FDMA operation could be used. This would save the cost of the TDMA timing and control and buffer equipment at earth stations. The cost evaluations of chapter 7 in fact will be based on the FDMA approach.

In regard to demand assignment, this can be easily incorporated into the models of TDMA, just the same as in the SDMA/SS-TDMA case of chapter 4. The remarks in chapter 4 about the limitations of this study in this respect also apply here.

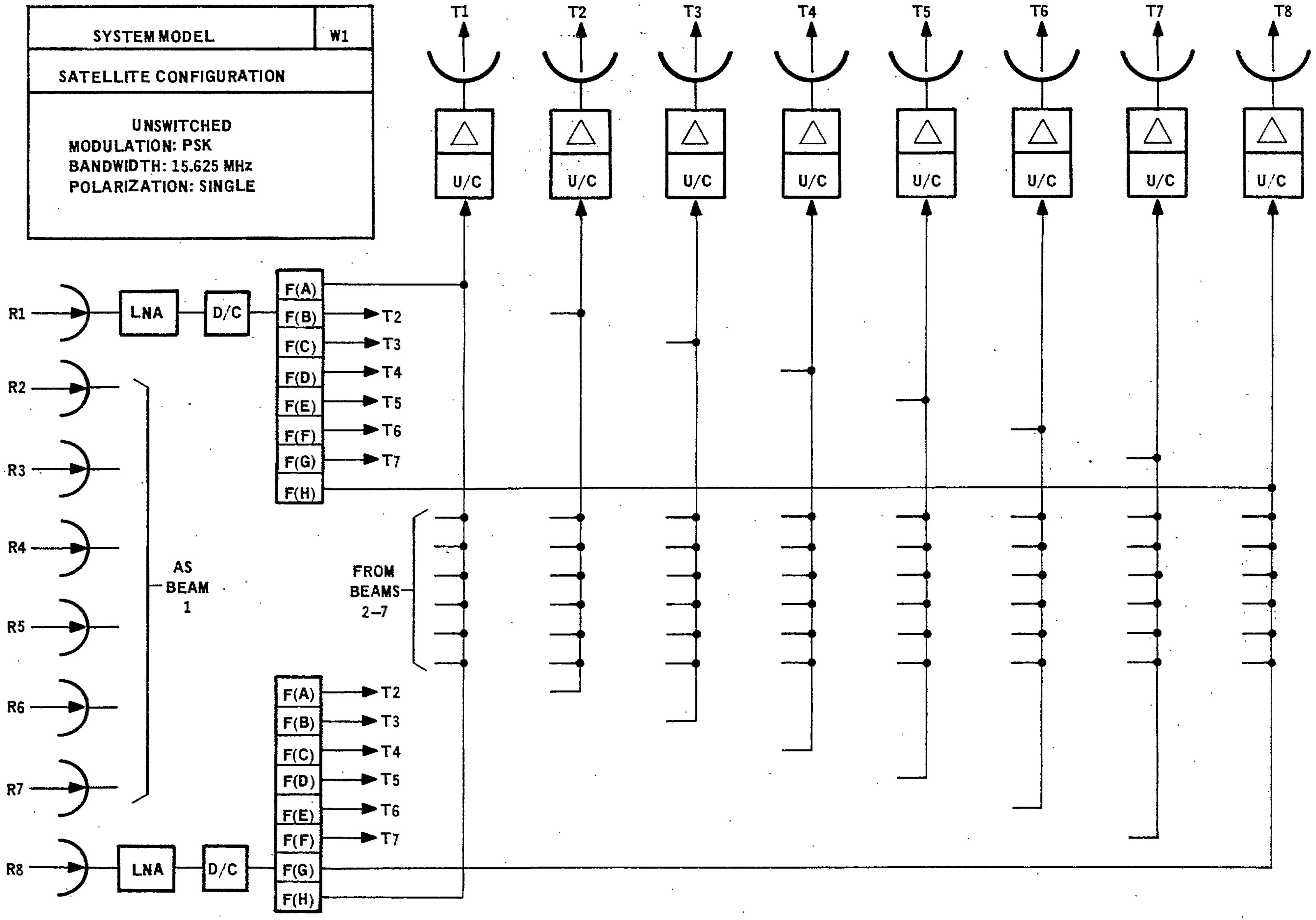


Figure 5-1a. Model W1 - Satellite Configuration

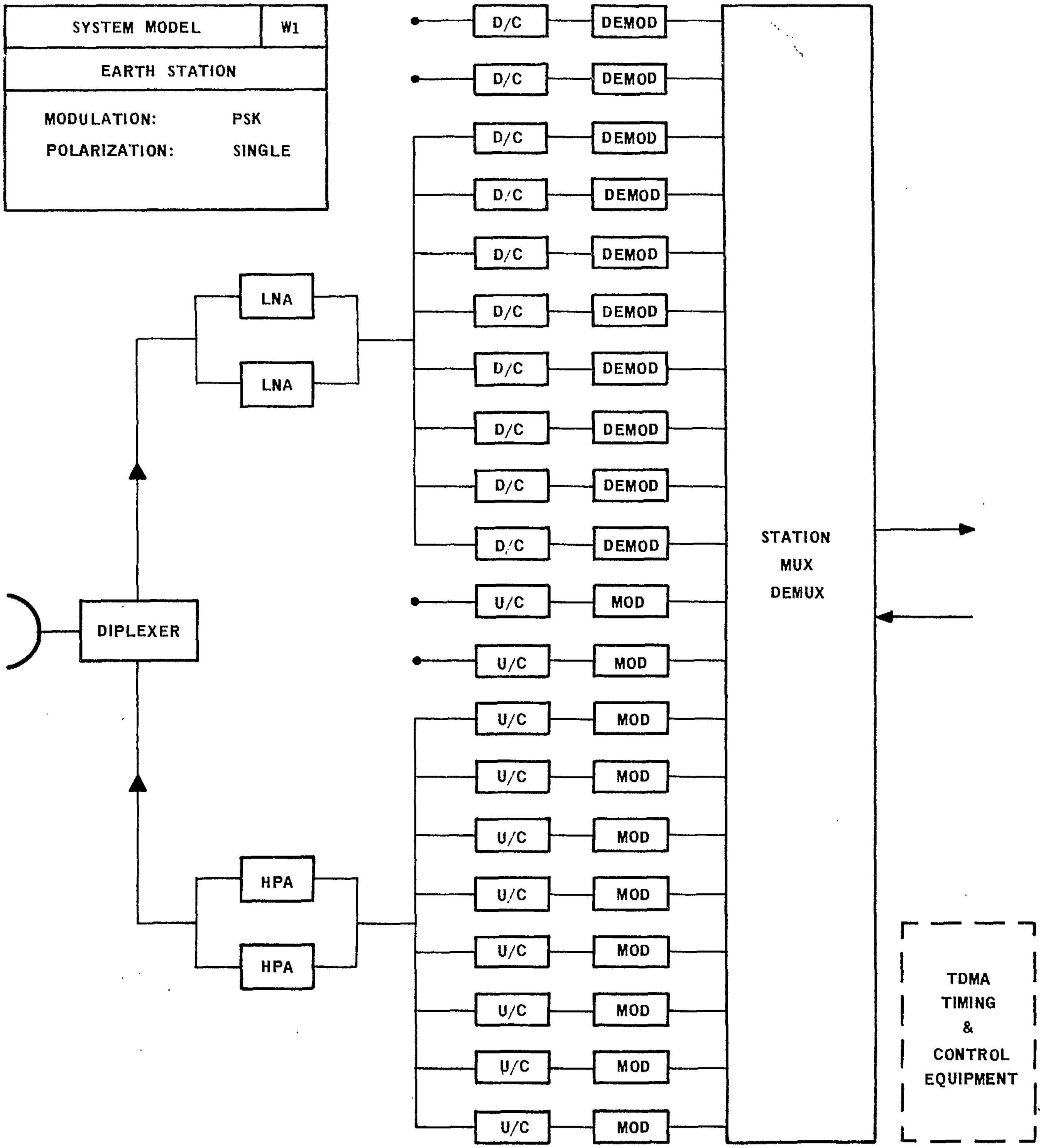


Figure 5-1b Model W1 - Earth Station Configuration

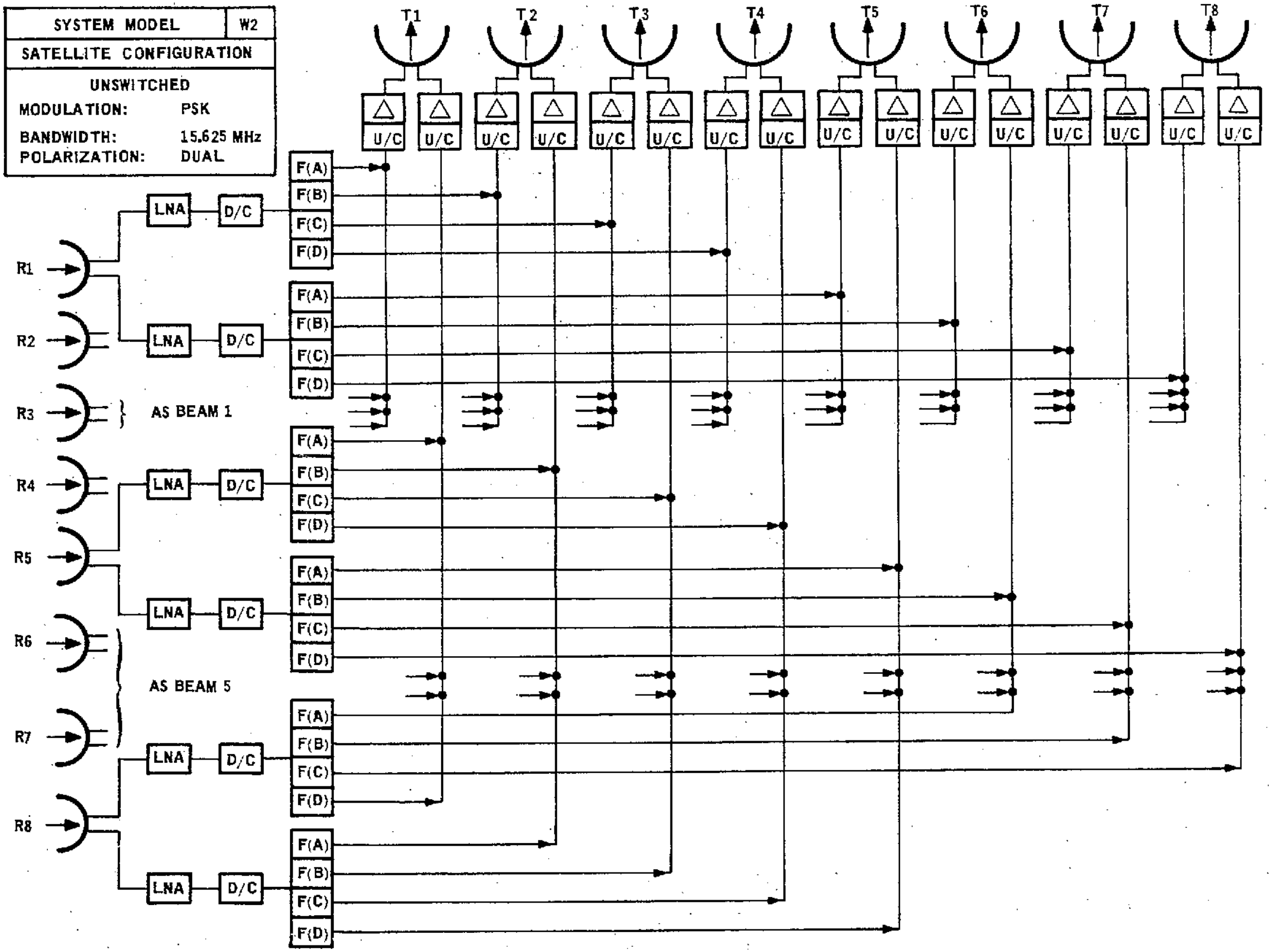


Figure 5-2a Model W2 - Satellite Configuration

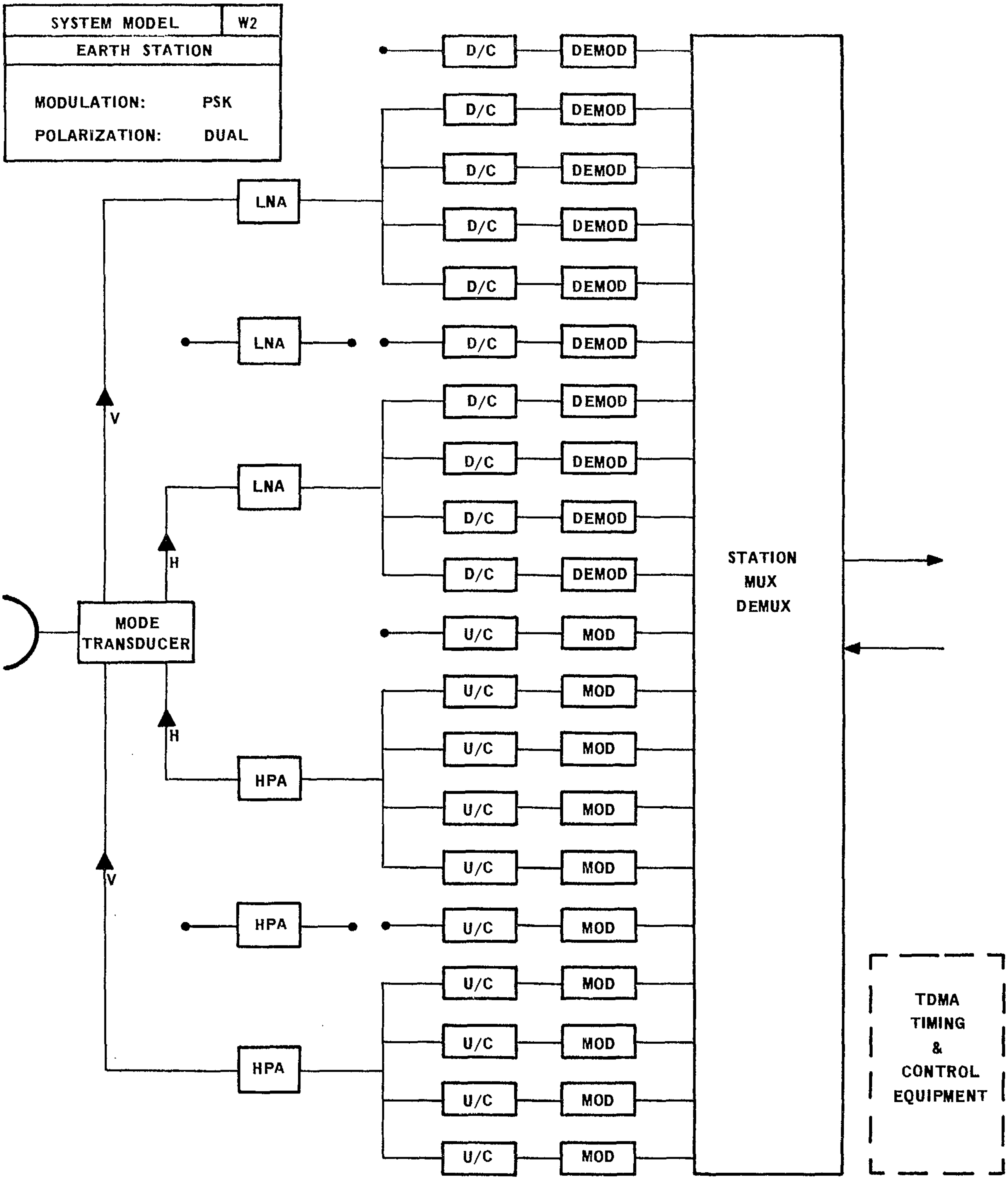
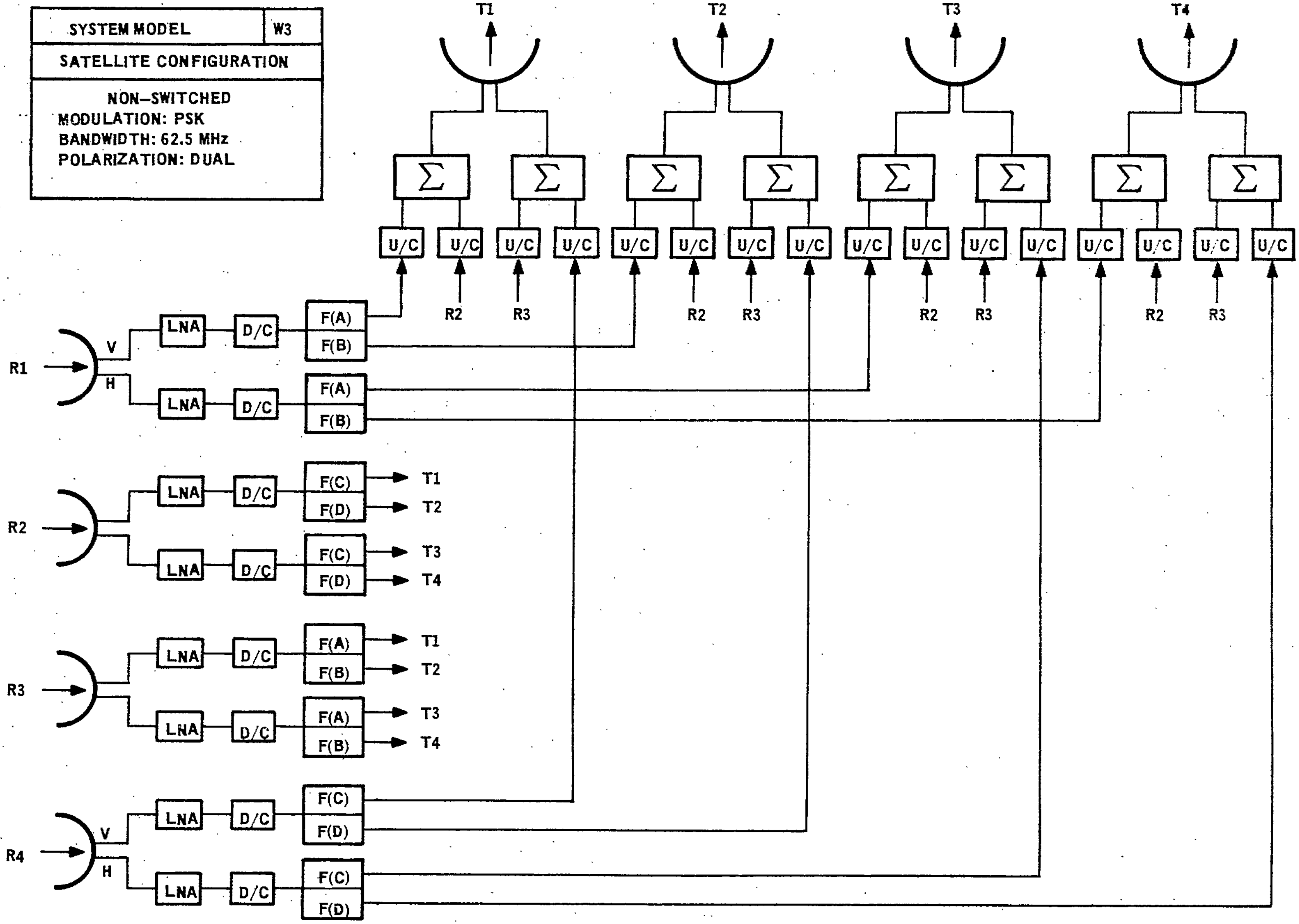


Figure 5-2b Model W2 - Earth Station Configuration



SYSTEM MODEL	W3
SATELLITE CONFIGURATION	

NON-SWITCHED
MODULATION: PSK
BANDWIDTH: 62.5 MHz
POLARIZATION: DUAL

Figure 5-3a Model W3 - Satellite Configuration

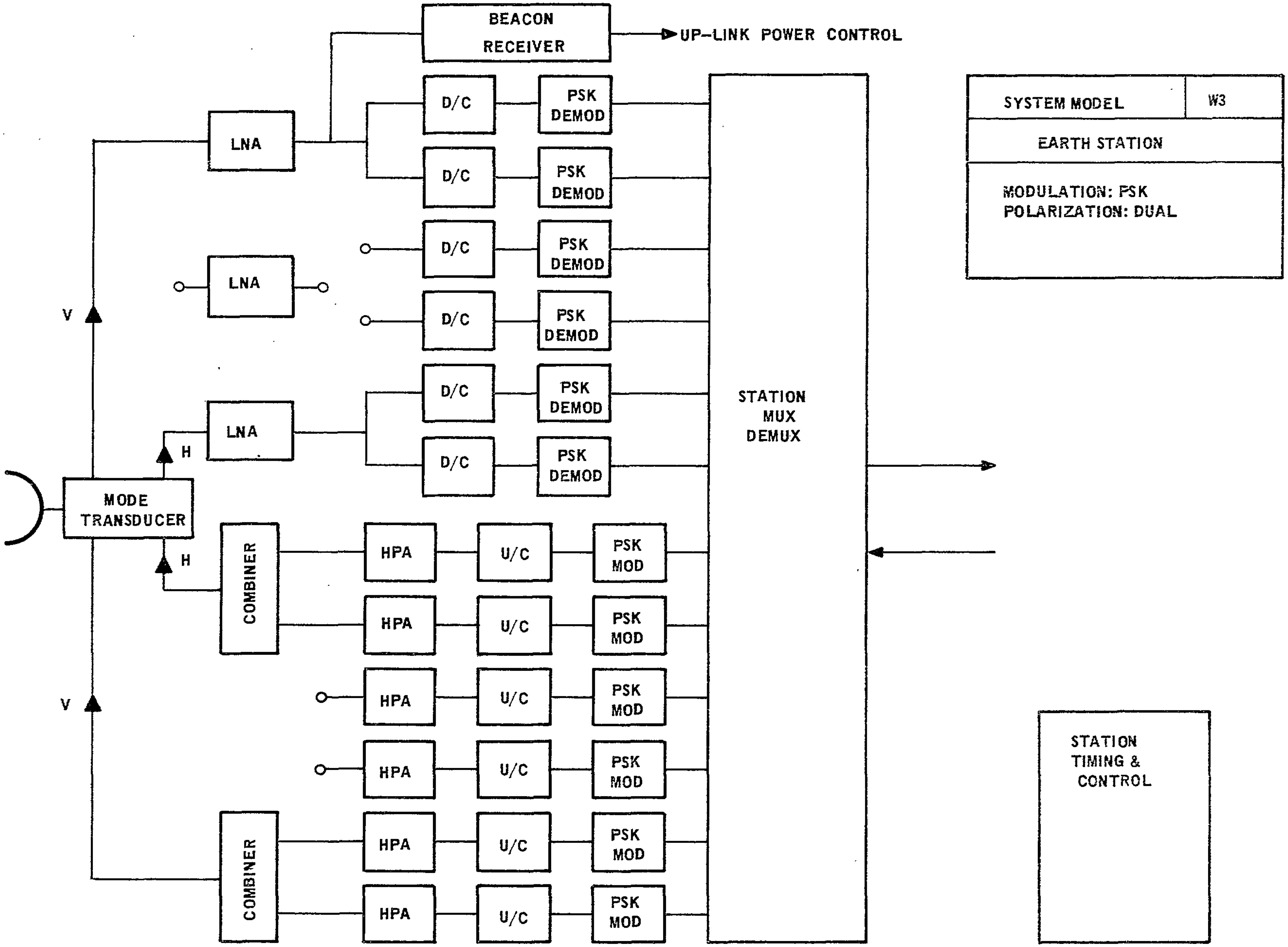


Figure 5-3b Model W3 - Earth Station Configuration

5.1.3 TECHNICAL SYSTEM TRADE-OFF RELATIONSHIPS

The system equations and parameters are the same as those given in chapter 4.

It is shown in chapter 4 that for a PCM-PSK system to cater for the capacity needs within the bandwidth available, the 4-phase coherent PSK system is the optimum. This conclusion is valid whether satellite switching is employed or not. Therefore all the system equations and parameters developed in chapter 4 are valid for the PCM-PSK system considered here for the non-switched models.

In models W1 and W2 in which multi-carrier operation in the transponder TWT is postulated, output power back-off of the TWT is necessary. From the data published by Westcott[30] it is quite easy to determine the optimum back-off for a system with many carriers. As shown in the report of the previous study[14], the optimum back-off occurs when the slope of the carrier-to-intermodulation curve is equal and opposite to the slope of the carrier-to-thermal noise curve. In this system, as in all other major route systems considered in this study, up-link power control is postulated, for reasons already given in chapter 4. Hence the most severe thermal noise contribution occurs when there is a down-link fade, which has a margin of 6 dB (see chapter 4). The optimum back-off is therefore determined such that with a down-link fade of 6 dB, the total thermal and intermodulation noise reaches the "critical S/N_0B" as defined in chapter 4. For the system here considered, this back-off is about 8 dB.

System economic trade-off will be presented in chapter 7. Each of the three configurations considered in this chapter will be evaluated in detail.

* * *

5.2 FDM-FM-FDMA SYSTEM

The principle of operation, the space and earth segment configurations and technical system parameters will be presented.

The FDM-FM-FDMA system is chosen for close examination because it is the most widely used technique at present, and may therefore be considered a baseline system against which other systems may be judged. In the subsections that follow, the principle of operation, the space and earth segment configurations and the technical system parameters will be presented.

* * *

5.2.1 PRINCIPLE OF OPERATION

The FDM-FM-FDMA technique is briefly reviewed.

The basic FDM-FM technique is well known and there is no need to explain it here. In a satellite system with one large coverage beam and many earth stations, some scheme of multiple access is obviously required. The frequency division multiple access (FDMA) technique is the most popular one at present. The most significant implication of FDMA is that many carriers will be amplified simultaneously in a common TWT amplifier and consequently, in order to achieve the necessary linearity, a certain amount of output power back-off is required. In a multiple beam satellite system, FDMA per se may or may not be present within any given beam, depending on whether a beam's coverage zone illuminates more than one earth station or not. Nevertheless, each up-link carrier is received, down converted and channelled by filters to an appropriate down-link subsystem. All the carriers going to a particular down-link beam can be amplified individually, combined and fed to the antenna. This would avoid the multiple carrier intermodulation problem. This, however, would need an inordinately large number of amplifiers and combining filters creating problems of excessive deadweight and equipment reliability. One alternative solution would be to employ a common amplifier in each down link operated in the back-off mode. With multi-carrier operation the amplifier now behaves in exactly the same manner as the conventional FDMA scheme in so far as the intermodulation and back-off requirements are concerned. Despite the back-off penalty, the FDMA mode is to be preferred on account of the equipment simplicity on board the satellite. Moreover, some degree of FDMA is unavoidable because certain beams cover more than one earth station. The FDMA model therefore will be used in this study.

In this study, the FDM-FM-FDMA system with and without speech compression will be evaluated. The conventional system without speech compression is well known to be quite inefficient from the system capacity point of view. In an attempt to overcome this, speech compression using the lincompex technique is postulated. The speech will be compressed in the transmit side and expanded correspondingly at the receive side. This is all done in the voice band 300 through 3400 Hz. In addition, a control channel is required between the transmit and receive lincompex equipment in order to convey the necessary information to control the far end expander. The lincompex equipment for high frequency circuits is available today, but its cost is prohibitively high for the large scale use postulated here. Much more economical techniques such as the use of delta modulation[29] may be feasible. The cost evaluations in chapter 7 assume the application of such techniques.

5.2.2 SATELLITE AND EARTH STATION CONFIGURATIONS

The space and earth segment configurations for the FDM-FM-FDMA non-switched models will be presented.

The frequency and beam plans for the non-switched satellite models are given in chapter 3. The frequency plans relevant to the FDM-FM-FDMA system are plans N1, N2 and N3. For each of the plans, there is a corresponding space and earth segment equipment configuration. These configurations are given in Figures 5-4, 5-5 and 5-6 and are designated models X1, X2 and X3. The space segment follows closely along the lines given in 5-1 for the digital modulation. The earth segment is, of course, quite different but is straightforward and conventional. As before, redundancy in the space segment on board each satellite is assumed to be 30 percent. The earth segment redundancy is as given in the block diagrams. For the same reasons discussed in chapter 4, up-link power control is postulated. This is done, as before, by monitoring a down-link beacon signal and then controlling the up-link transmitter amplifier output.

With FDM-FM-FDMA technique, demand assignment is difficult to implement. Demand assignment is therefore not considered here.

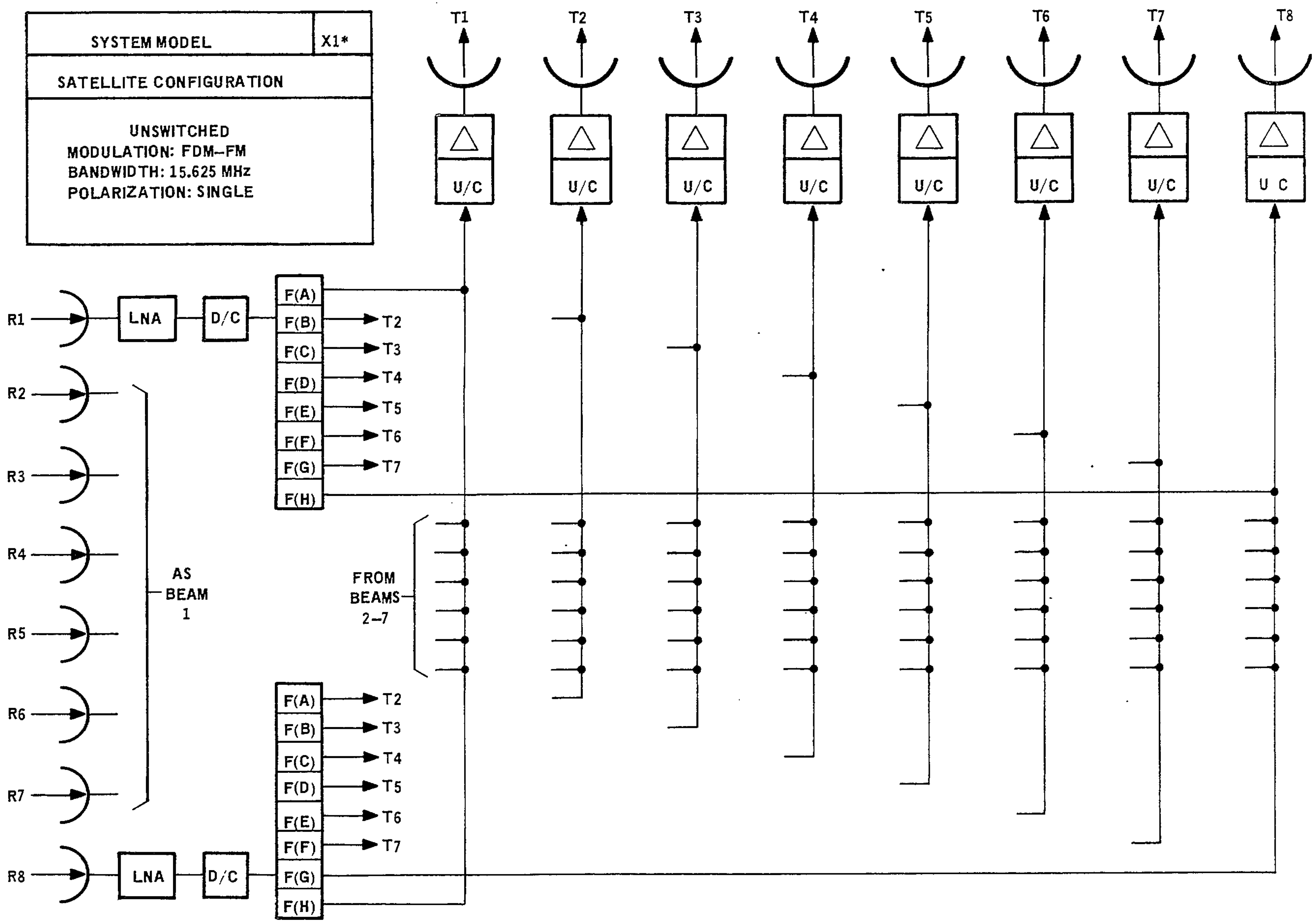


Figure 5-4a Model X1* - Satellite Configuration

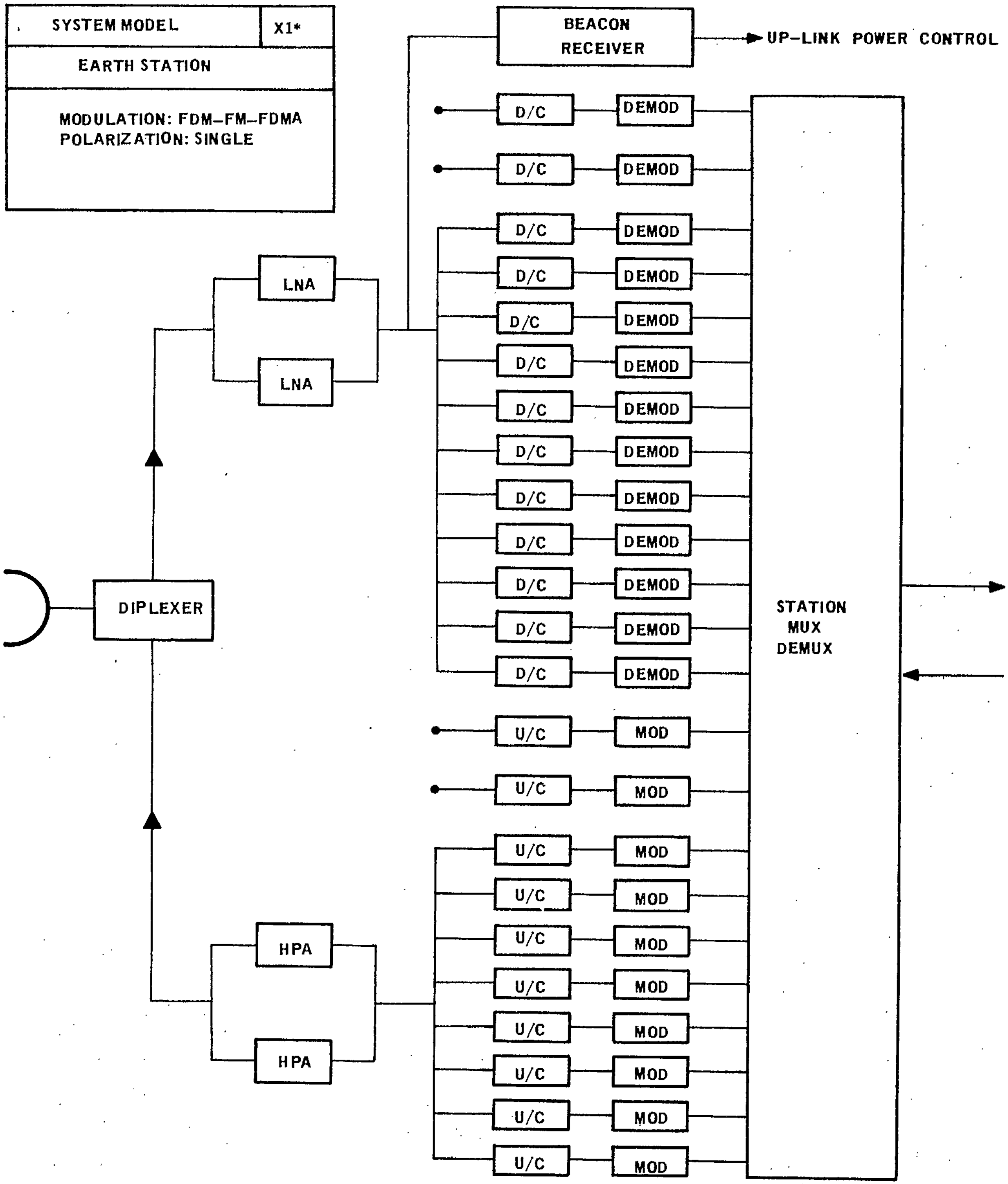


Figure 5-4b Model X1* - Earth Station Configuration

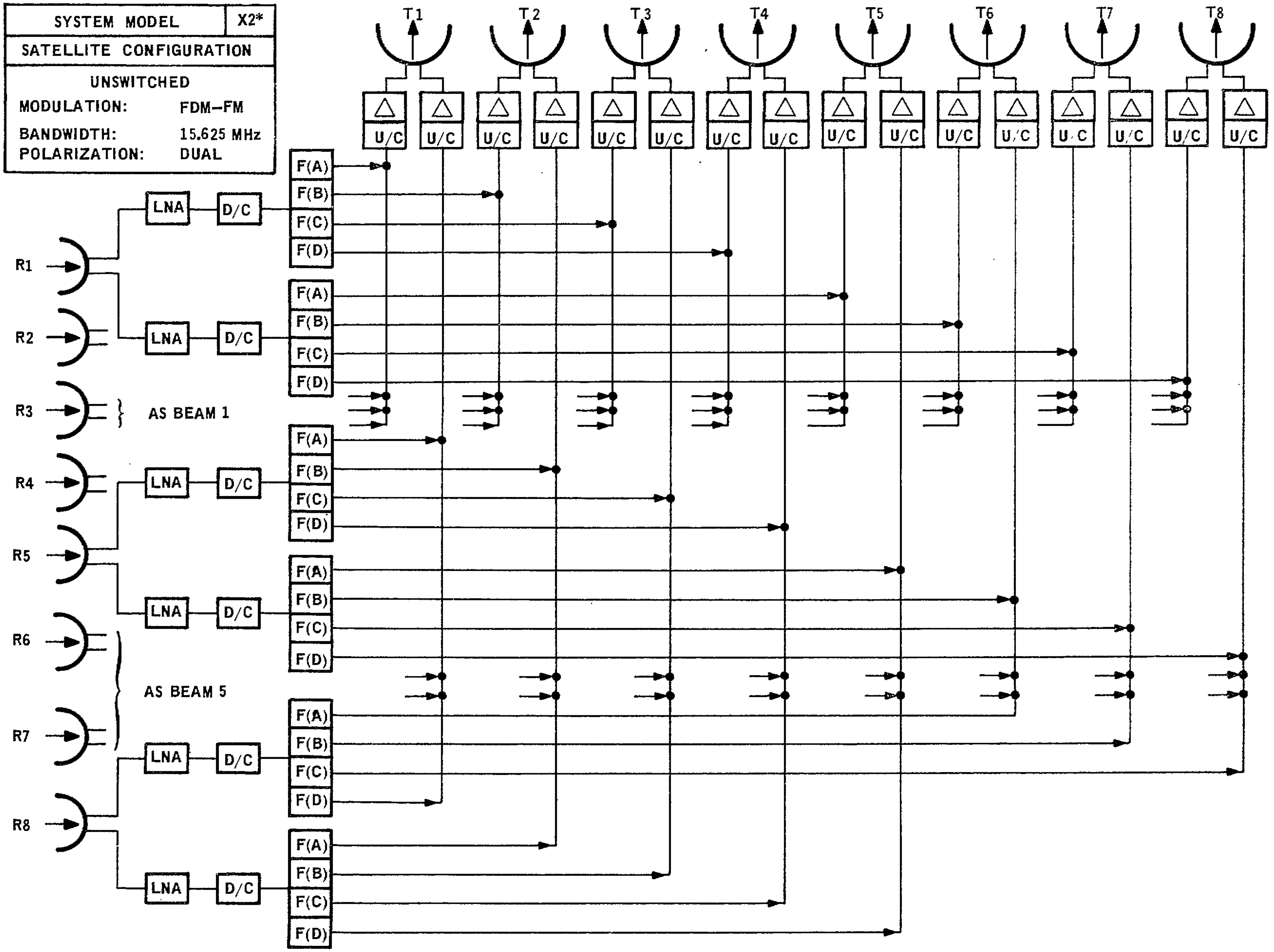


Figure 5-5a Model X2* - Satellite Configuration

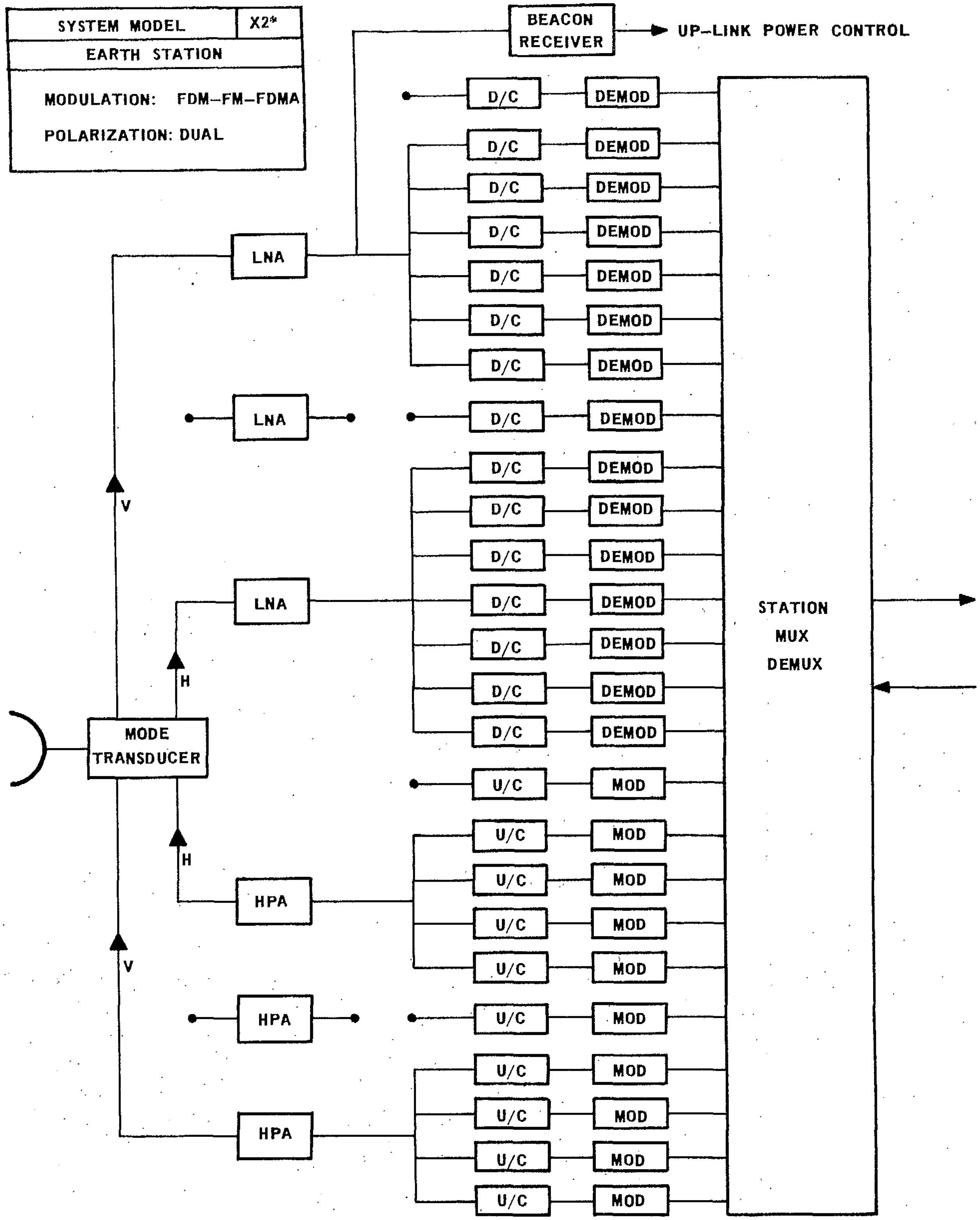


Figure 5-5b Model X2* - Earth Station Configuration

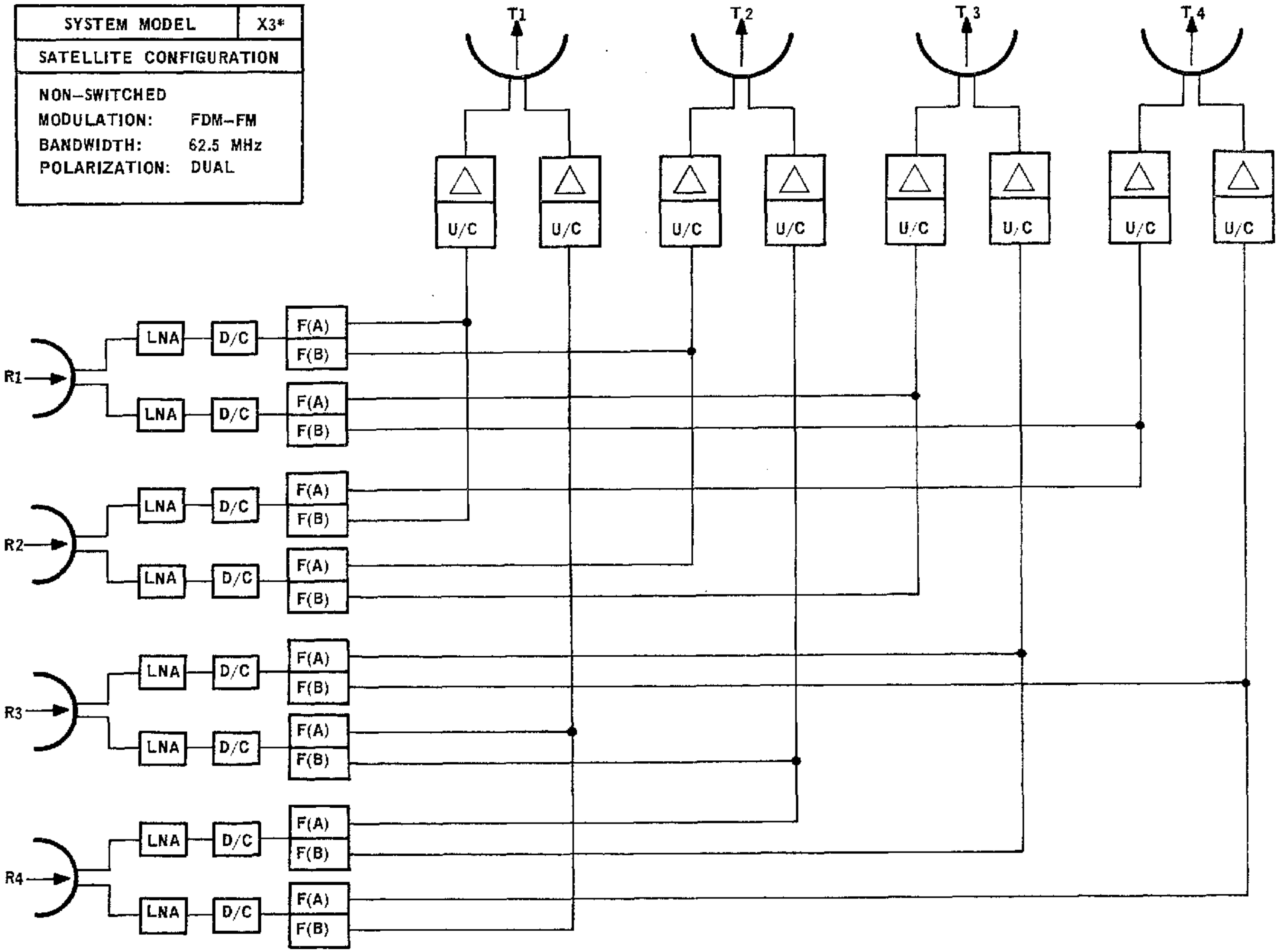


Figure 5-6a Model X3* - Satellite Configuration

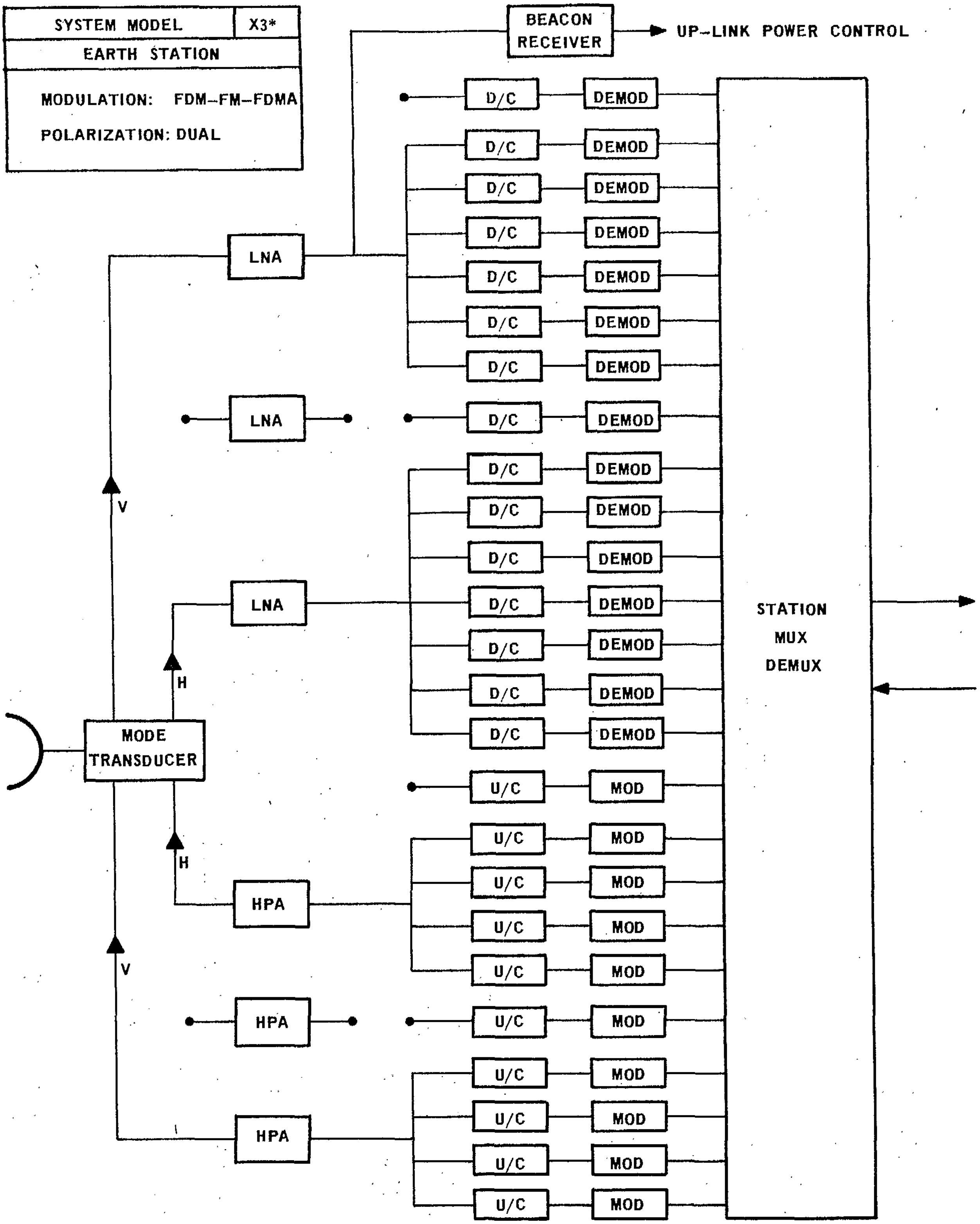


Figure 5-6b Model X3* - Earth Station Configuration

*   *   *

5.2.3 TECHNICAL SYSTEM TRADE-OFF RELATIONSHIPS

The system equations for an FDM-FM-FDMA network are derived, taking into account the interference and multi-carrier intermodulation effects.

In an FDM-FM-FDMA system, operated in a multiple beam and frequency reuse environment, interference and multiple carrier intermodulation are important considerations. In chapter 3, from considerations of antenna beam sizes and frequency and polarization plans, it is shown that interference would not be expected to exceed a level corresponding to C/I = 18 dB. Further, in chapter 2, the overall noise performance objective is stated to be 38 dBrnc0, with interference contributing half of this. The systems equations are now derived below that satisfy all these constraints. Both the conventional system without speech compression and the system with lincompex speech compression will be considered: Appendix C will show that the tolerable interference level is consistent with the noise budget allocation.

Conventional FDM-FM-FDMA

Overall design noise objective = 38 dBrnc0

∴ Interference noise allocation = 35 dBrnc0

Thermal and satellite intermodulation noise allocation = 35 dBrnc0

Now FDM peak loading is given by

$$P_s = -15 + 10 \log (N) + 10 \text{ dBm.}$$

where N = number of voice channels

10 = peak to rms factor for FDM signals.

The FM equation gives the test-tone to noise ratio (unweighted) as:

$$-S/N = P_s - 20 \log M + 10 \log 3100 - P_E - \left(\frac{C}{N_o}\right)_r \tag{5.1}$$

where M = Peak deviation index = $\frac{\Delta F}{fm}$

fm = Top baseband frequency of FDM signal

ΔF = Peak frequency deviation due to FDM signal.

P_E = pre-emphasis advantage = 4 dB

$(\frac{C}{N_o})_r$ = Required carrier-to-noise density.

The term $(\frac{C}{N_o})_r$ is also related to M and fm via the threshold equation* given in reference 33; i.e.,

$$(\frac{C}{N_o})_r = 10.5 + M_a + 10 \log fm + 12.9 \log (1+M) \qquad (5.2)$$

where M_a = systems fade margin = 6 dB for 99.99 percent systems reliability with up-link power control (see chapter 4).

Substituting:

$$S/N = 88 - 35 = 53 \text{ dB unweighted, and}$$

$$fm = 4200N \text{ (approximate)}, \qquad (5.3)$$

into equations (5.1) and (5.2) and combining these two equations, the following relationship results:

$$20 \log M + 12.9 \log (1+M) = 32.8 - M_a = 26.8$$

solving:

$$\therefore M = 6.3$$

Substituting this value of M into equation (5.2), and dividing fm by the number of voice channels (N) the carrier to noise density per channel is given by:

$$(\frac{C}{N_o}) = 10.5 + 6 + 36.2 + 11.1 = 63.8 \text{ dB-Hz}$$

The bandwidth per voice channel is given from Carson's Rule as follows:

**Note*: Interference is not generally considered to have an effect on threshold[32].

RF Bandwidth

$$B = 2fm\ (1+M)$$

$\therefore$ Bandwidth per voice channel

$$b = \frac{B}{N} = \frac{2fm}{N}(1+M).$$

since $\frac{fm}{N} = 4.2$ kHz

$\therefore$ bandwidth $b = 8.4\ (1+M)$;

substituting the value of M derived above:

$$\therefore b = 8.4(1 + 6.3)$$

or $b = 64$ kHz.

Hence required carrier-to-noise ratio is given by

$$(\frac{C}{N}) = 63.8 - 10 \log 64 \times 10^3.$$

$$= 15.8 \text{ dB.} \qquad (5.4)$$

This is the carrier-to-noise ratio at the input to the demodulator, under normal operational conditions. The pre-detection noise is made up partly by thermal noise and partly by multi-carrier intermodulation noise of the TWT. The optimum distribution can be found from the intermodulation noise characteristics of the TWT versus input back-off. If the number of carriers is large as it is here, the intermodulation is sensibly independent of the number of carriers[30].

As shown in Appendix B of the previous study report[14], the optimum TWT input back-off point is obtained when the slope of the carrier-to-intermodulation (C/IM) curve is equal and opposite to the slope of the carrier-to-thermal noise (C/N) curve. From typical TWT characteristics as measured by Westcott[30], the optimum output back-off is found to be about 6.8 dB.

Speech Compressed FDM-FM-FDMA System

For the speech compressed system equipped with lincompex units, as shown in Appendix C, the required test-tone-to-noise ratio is about 15 dB. As before, half of the noise will be allocated to interference and half to thermal and intermodulation effects. Hence the thermal and intermodulation contributions would give:

$$\frac{S}{N} = 18 \text{ dB}$$

where $\frac{S}{N}$ = test-tone-to-noise ratio (unweighted).

For the lincompex unit using delta modulation as the compressor, reference 29 indicates that the speech volume at the output is maintained sensibly constant at zero dBm(0). However, in a normal telephone circuit, conversation is actively present for considerably less than half the total circuit holding time. It is assumed that on the average the activity is about 25 percent. Hence the average (rms) power in a baseband comprising N channels may be taken to be -6 + 10 log N dBm(0). Assuming that the baseband signal is still Gaussian in nature, the peak factor is 10 dB, as before. Hence the peak system loading is given by:

$$P_s = -6 + 10 \log N + 10 \text{ dBm}(0).$$

$$= 4 + 10 \log N \text{ dBm}(0) \qquad (5.5)$$

It should be pointed out that this system loading postulates the use of either a common channel interoffice signalling (CCIS) or a signalling scheme in which the signalling tones do not pass through the lincompex units. This would mean modifications to the current single frequency (SF) or multifrequency (MF) signalling schemes. Now, using equations (5.1), and (5.2) and (5.3) and substituting the above appropriate expressions of S/N, P_s, and M_a (= 6 dB), the following results:

$$20 \log M + 12.9 \log (1+M) = 0.2$$

Solving for M:

$$M = 0.72$$

As before, dividing equation (5.2) by N, the carrier-to-noise density per channel is obtained as below:

$$(\frac{C}{N_o}) = 10.5 + 6 + 36.2 + 3 = 55.7 \text{ dB-Hz}.$$

Bandwidth per voice channel is given by:

$$b = 8.4\ (1+0.72)$$

$$= 14.5 \text{ kHz}.$$

The carrier-to-noise ratio is:

$$\frac{C}{N} = 55.7 - 10 \log 14.5 \times 10^3 = 14.1 \text{ dB.}$$

$$= 14.1 \text{ dB.}$$

This carrier-to-noise is made up of contributions of thermal noise and multi-carrier intermodulation noise. The optimum TWT output back-off is found to be about 5.6 dB.

* * *

5.3 SINGLE-CHANNEL-PER-CARRIER SYSTEM

The use of speech compression technique makes it feasible to use a single-channel-per-carrier system. This system could have potentials of easy implementation of demand assignment giving network flexibility. The principle of operation, the space and earth segment configurations and the systems equations are presented.

As discussed earlier, with the use of speech compression utilizing the lincompex principle, it is possible to tolerate a high level of interference and intermodulation noise in the voice baseband. A high capacity scheme using single-channel-per-carrier could be feasible. The single-channel-per-carrier approach lends itself easily to demand assignment operation. Further, the earth station receive figure-of-merit (G/T) and up-link transmit EIRP are directly proportional to the number of channels served by that earth station. It is feasible therefore to closely match earth station costs to the capacity needs at each geographical location. Hence it is considered worthwhile to examine the technical and economic merits of such a scheme. It should, however, be pointed out that the lincompex speech compression technique, using the delta modulation scheme postulated in this study, has not been fully tested and further work in this area is necessary.

* * *

5.3 Single-Channel-Per-Carrier System

5.3.1 PRINCIPLE OF OPERATION

The principle of operation is straight forward, but the network management requires the system to be locked to a pilot frequency.

The principle of operation of the single-channel-per-carrier system is straight forward. Such a technique is well developed and is being applied in the Intelsat network in the SPADE system, which uses PCM and digital modulation as against the analog compressed FM system postulated here. The lincompex used is the same as that employed in connection with the FDM-FM-FDMA scheme. On account of the very narrow rf bandwidth proposed, it is necessary to provide an rf reference pilot to which the whole network is locked. A similar technique is also used in SPADE. One rf pilot has to be provided per transponder. With multiple beam operation in which there are many separate satellite up- and down-converters in the various beams, in order that the pilot frequency drift and hence the total network frequency drift be all in synchronism, it is necessary that the local oscillators in the up- and down-converters on board the satellite be phase locked to each other. This necessitates some complications in hardware design and may pose some reliability problems.

One major advantage of the single-channel-per-carrier technique is the ease of routing traffic and implementation of demand assignment. This is simply done by selecting an appropriate pair of transmit and receive frequencies. Through filtering in the satellite, the selected carriers will be automatically channelled to the appropriate satellite antenna beams. In effect, the satellite performs as a frequency division switch among the spot beams.

To really get the best advantage of a single-channel-per-carrier system, the normal telephone talker activity factor should also be exploited. This means that the rf carrier should be turned on only when speech or signalling tones are present. This is the well known voice activation system which is also used in the SPADE network. Voice activation will therefore be used here in this study.

* * *

5.3.2 SATELLITE AND EARTH STATION CONFIGURATIONS

The space and earth segment configurations of the single-channel-per-carrier speech compressed FM system are presented. An earth segment rf pilot frequency locking scheme is explained.

The space segment configuration for each of the antenna beam plans is the same as the FDM-FM-FDMA case considered previously. The earth segment configurations, however, are quite different. Since there is a carrier for each channel, and since the number of modulators and demodulators (modems) must equal the number of carriers, each earth station would therefore be equipped with the number of modems equal to its channel capacity. The block diagrams of the space and earth segment configurations are shown in Figures 5-7, 5-8 and 5-9. These are designated models Y1, Y2 and Y3, corresponding to frequency plans N1, N2 and N3 respectively (see chapter 3). The earth segment is unconventional and warrants some brief explanation as given below.

Refer to Figure 5-10 for a block diagram of a simplified transmit and receive subsystem at an earth station. The whole design approach is aimed at solving the network frequency stability problem.

The first up- and down-converters are fed by a common local oscillator whose frequency is exactly midway between the up-link and down-link rf channel center frequencies. The transmit and receive intermediate frequencies (i-f's) are therefore 1.15 GHz. Consider first the receive subsystem. The 1.15 GHz i-f (designated IFI) is mixed with a second local oscillator to produce another i-f called IF2 centered around 70 MHz. This second local oscillator is phase locked to a received pilot frequency. The IF2 containing a multitude of individual carriers, each frequency modulated by a single channel, is further mixed with individual crystal controlled third local oscillators to produce a third i-f called IF3, which may be 10.7 MHz. This IF3 is at a fixed frequency and will go through a narrow band mechanical or crystal filter. Hence the crystal controlled third local oscillator acts as the channel selection element. The narrow band filtered IF3 then goes to the demodulator. From the output of the demodulator, the voice baseband signal then goes to the voice activation squelch control unit and thence to the lincompex equipment.

In the transmit path, speech is compressed by the lincompex and is routed via the voice activation device to the fm modulator which produces an i-f of 10.7 MHz. This is up-converted by the same crystal controlled local oscillator as used in the receive path, to IF2 at the 70 MHz band. The next up-conversion stage is from 70 MHz to 1.15 GHz. This is done by a transmit local oscillator that is phase locked to the receive local oscillator VCO such that a negative frequency error correction in the receive local oscillator VCO would result in a positive frequency correction in the transmit local oscillator VCO. The 1.15 GHz i-f is then up-converted by the common rf local oscillator to the 14 GHz band. The reason for the above opposite frequency error correction between the receive and transmit paths is to cancel the frequency drift in the common rf local oscillator. As this local oscillator operates at the very high frequency range of around 13 GHz, it is not easy to maintain the required stringent frequency stability, and an error cancelling scheme such as the one proposed is needed.

The above described frequency control scheme would also cater for the satellite transponder frequency drift, provided the errors caused to the network up-link and down-link frequencies are in the same direction. This requires that the frequency translating oscillators on board the satellite be phase locked to each other.

For demand assignment operation, the third crystal controlled local oscillators which function as channel selectors would be replaced by remote controlled programmable frequency synthesizers. Demand assignment can therefore be easily implemented.

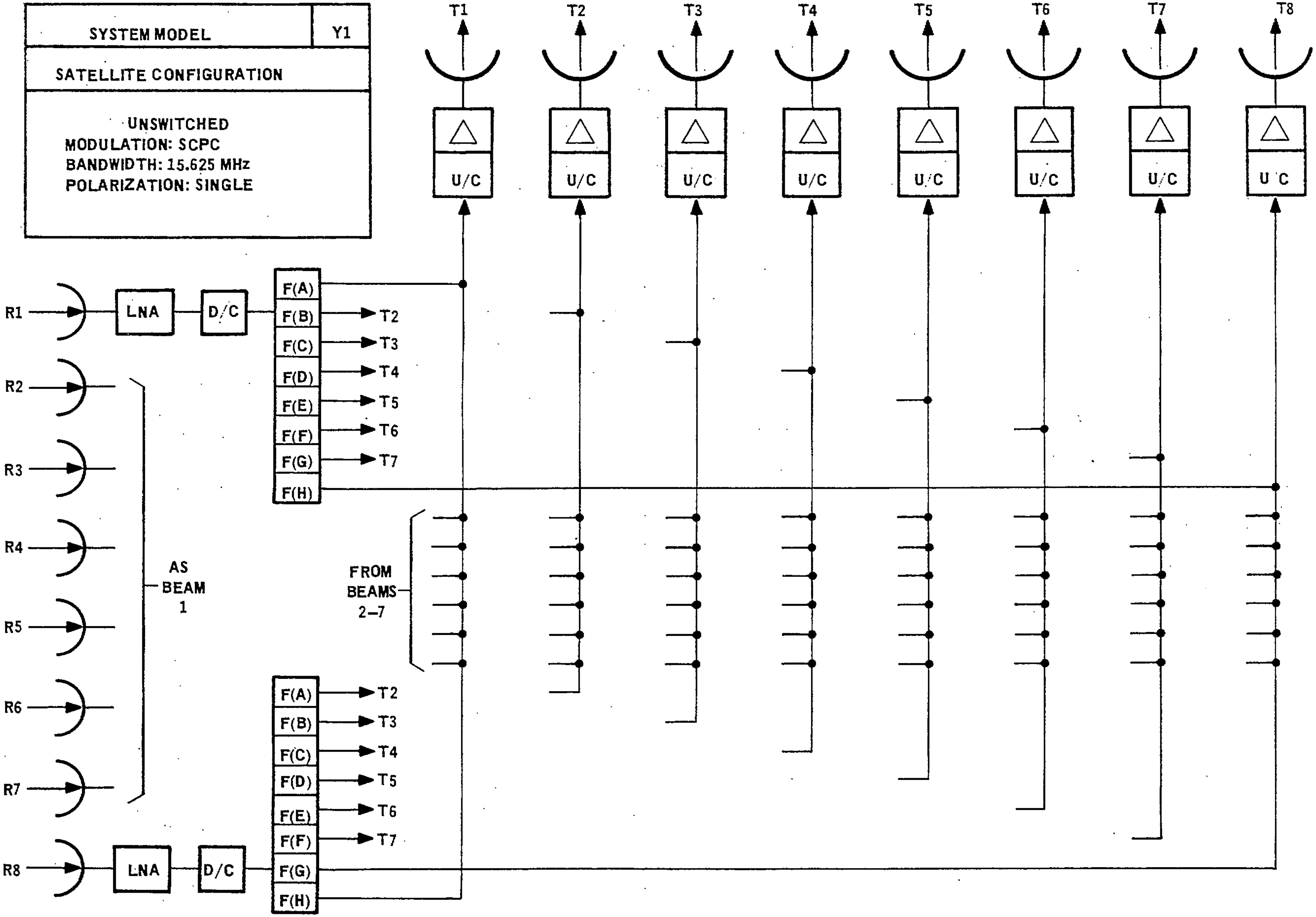


Figure 5-7a Model Y1 - Satellite Configuration

SYSTEM MODEL	Y1
EARTH STATION	
MODULATION: SCPC	
POLARIZATION: SINGLE	

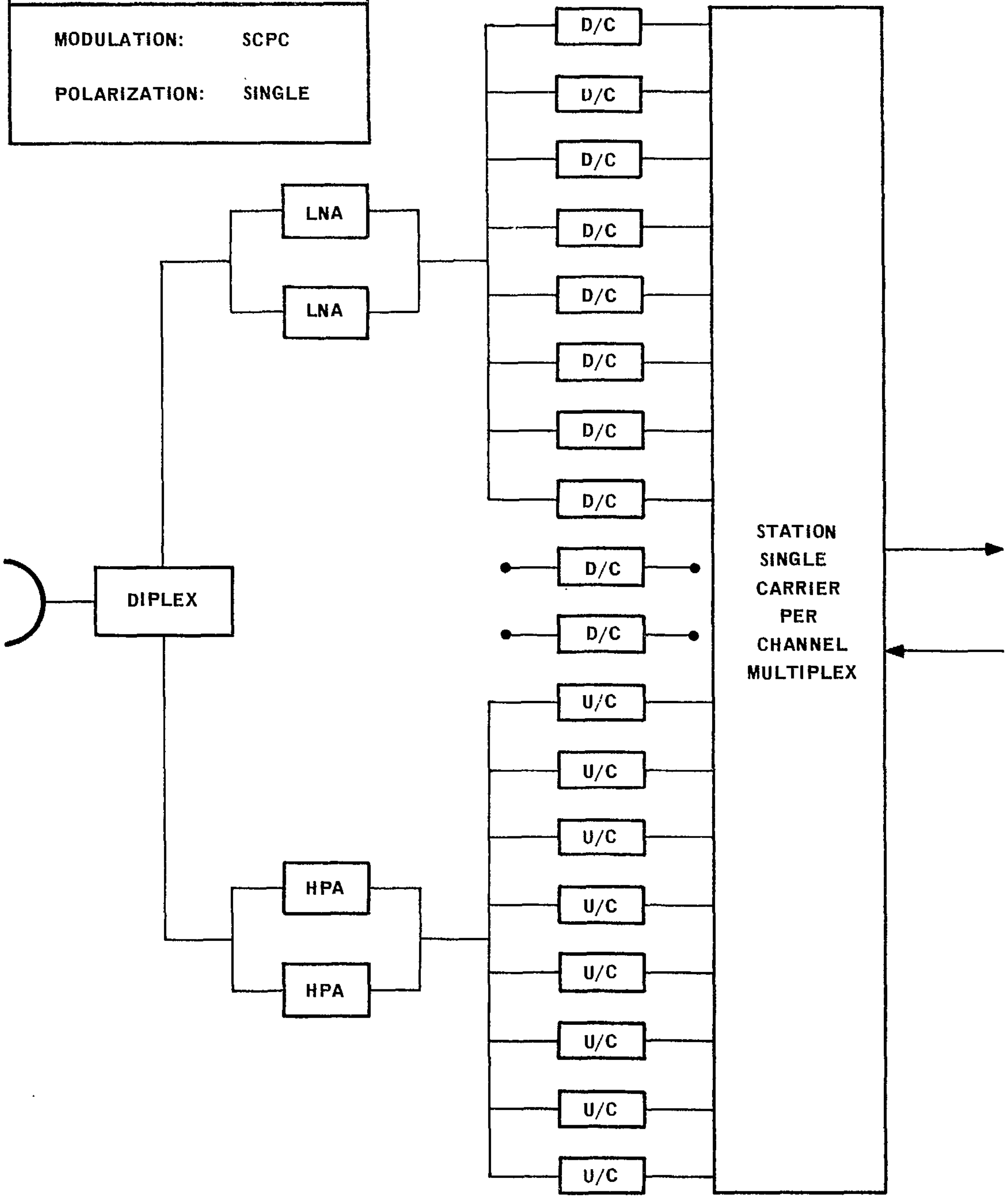


Figure 5-7b Model Y1 - Earth Station Configuration

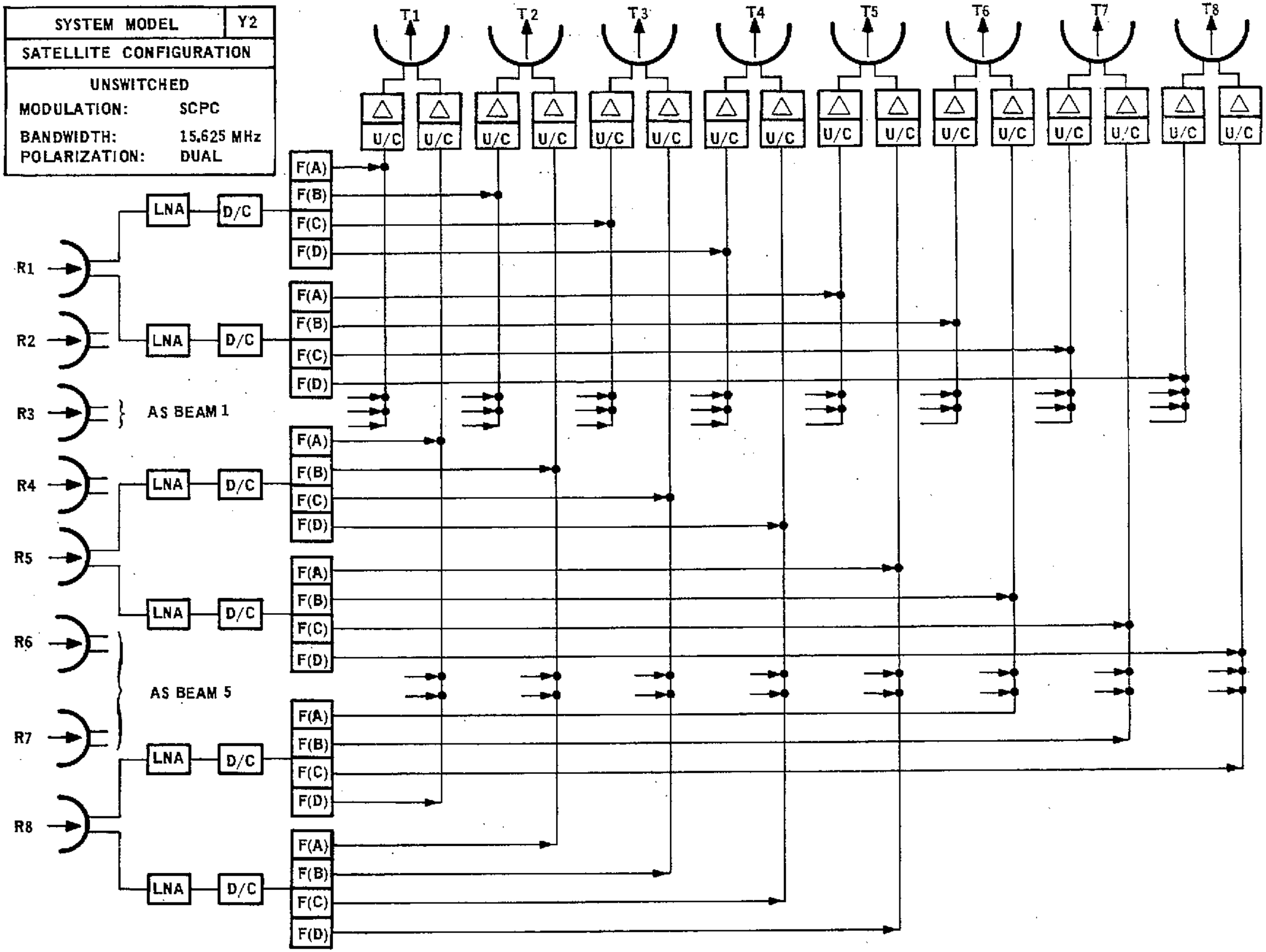


Figure 5-8a Model Y2 - Satellite Configuration

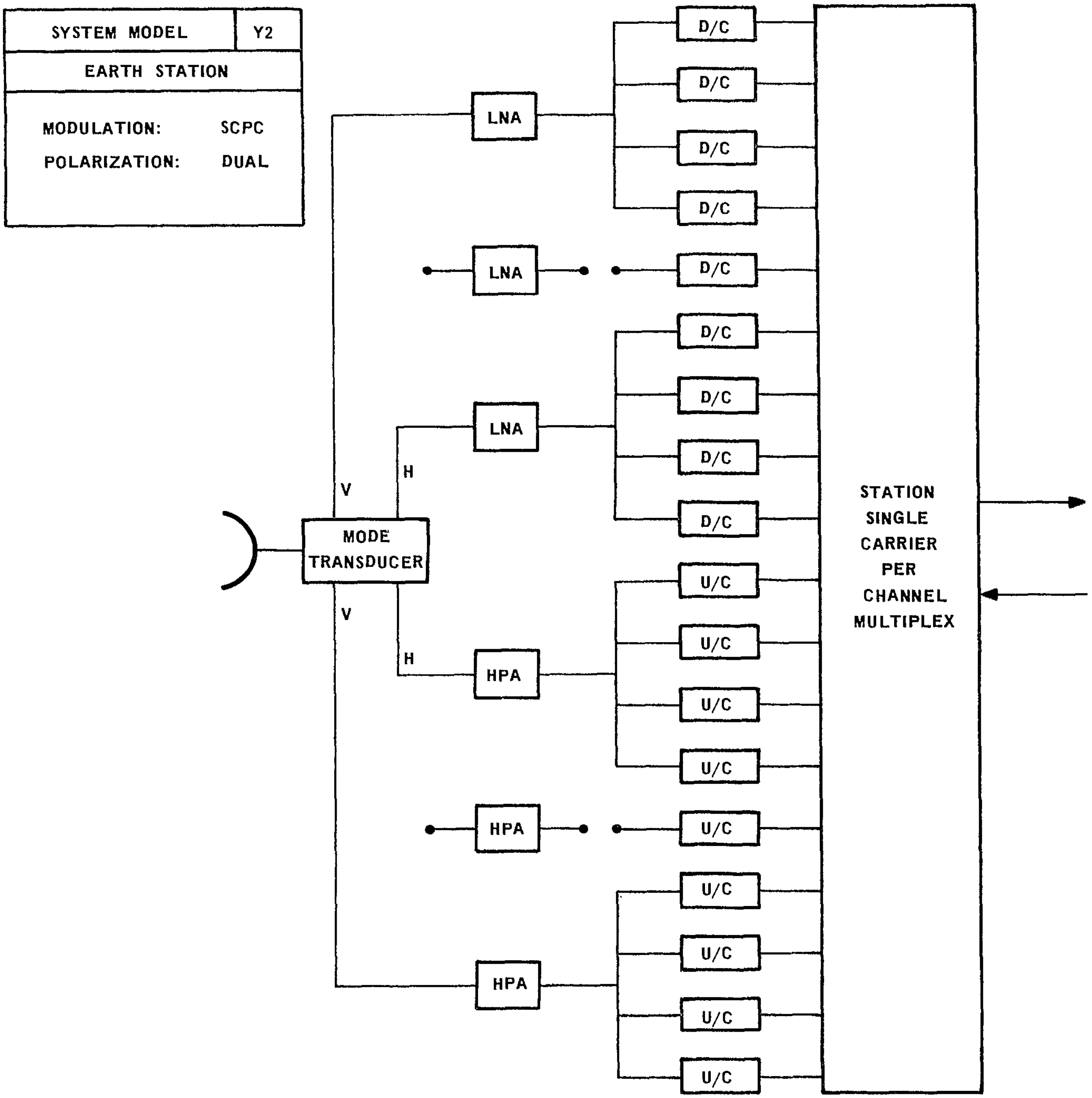


Figure 5-8b Model Y2 - Earth Station Configuration

SYSTEM MODEL	Y3
SATELLITE CONFIGURATION	
NON-SWITCHED	
MODULATION:	SCPC
BANDWIDTH:	62.5 MHz
POLARIZATION:	DUAL

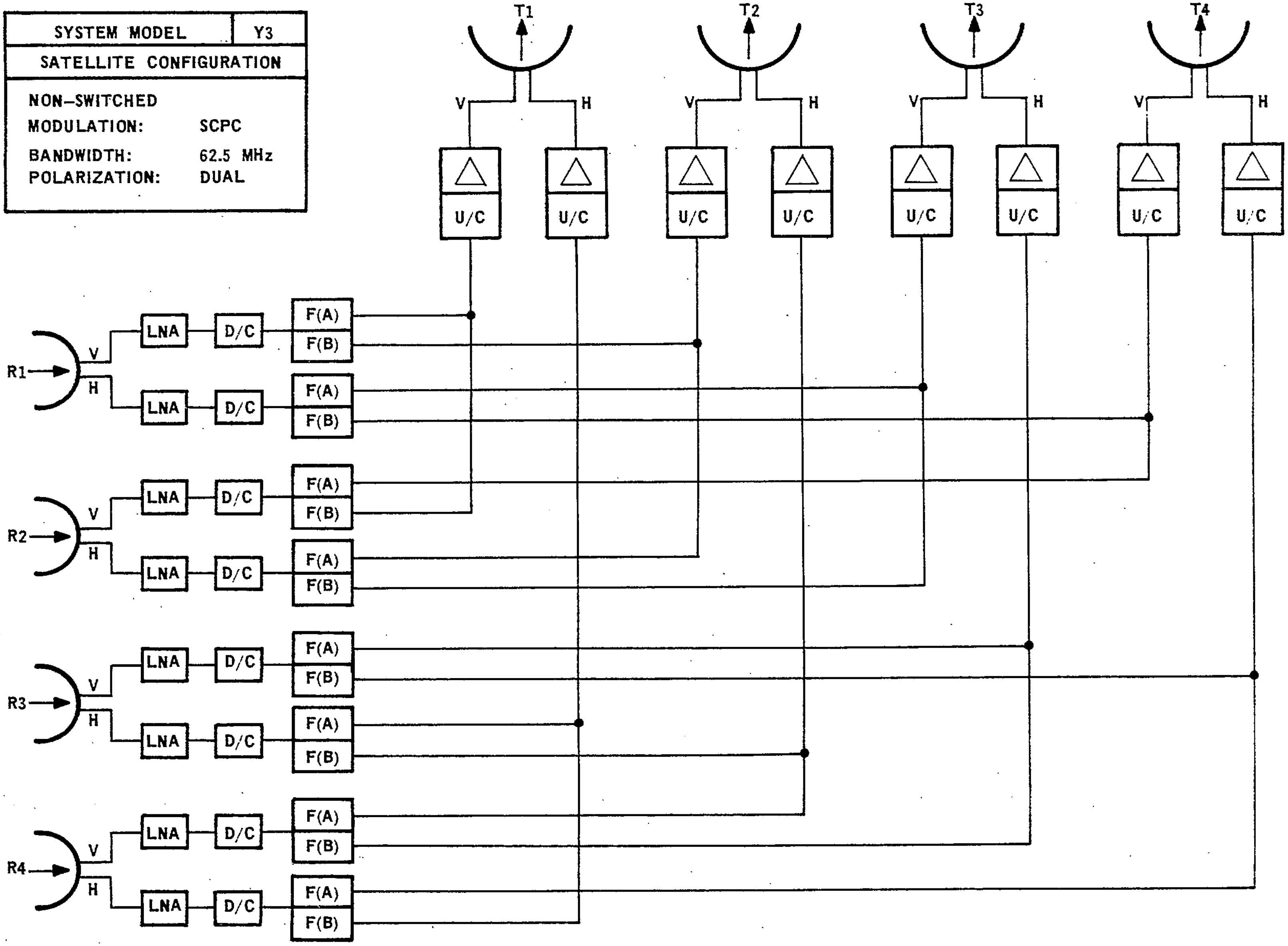


Figure 5-9a Model Y3 - Satellite Configuration

SYSTEM MODEL	Y3
EARTH STATION	
MODULATION:	SCPC
POLARIZATION:	DUAL

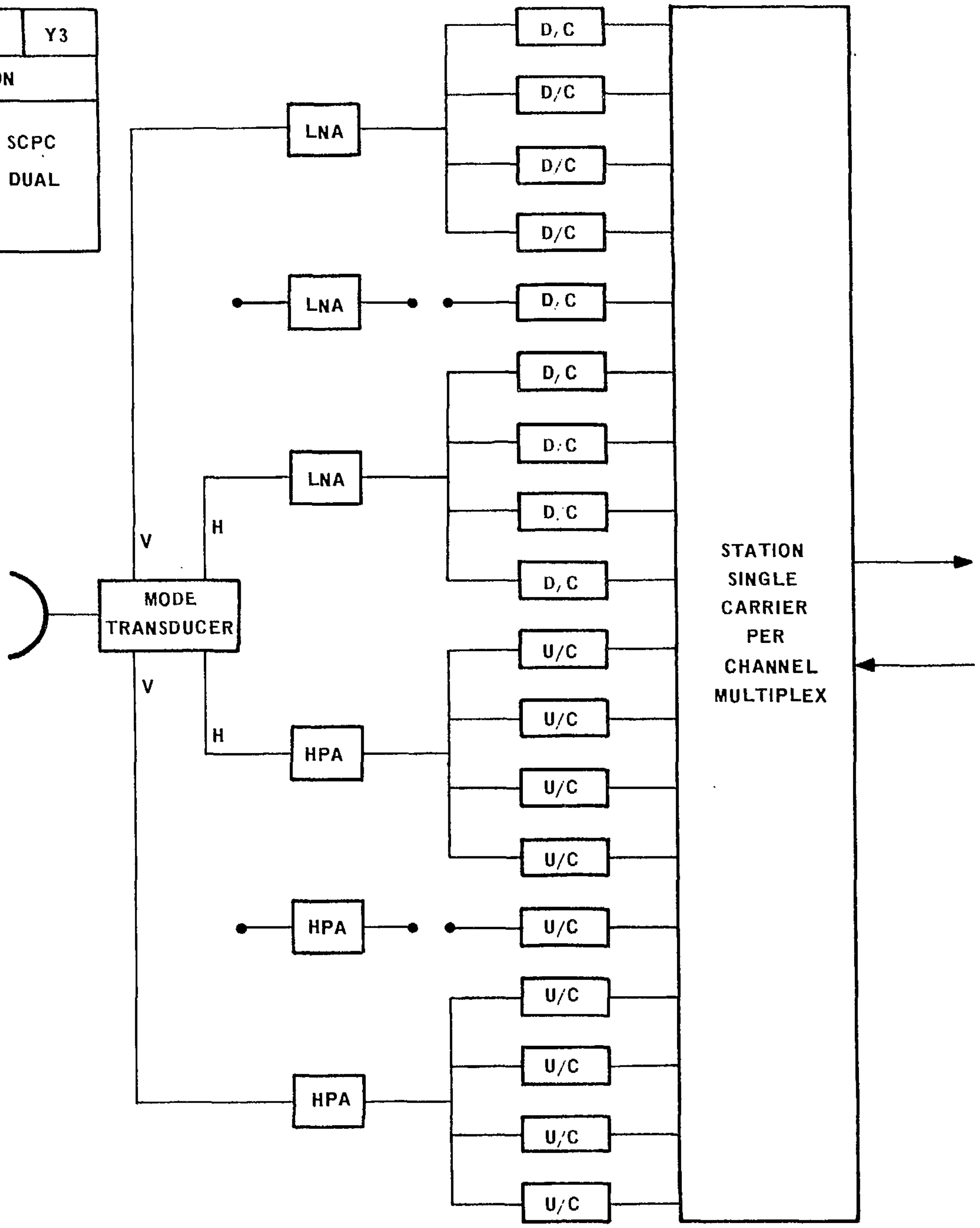


Figure 5-9b Model Y3 - Earth Station Configuration

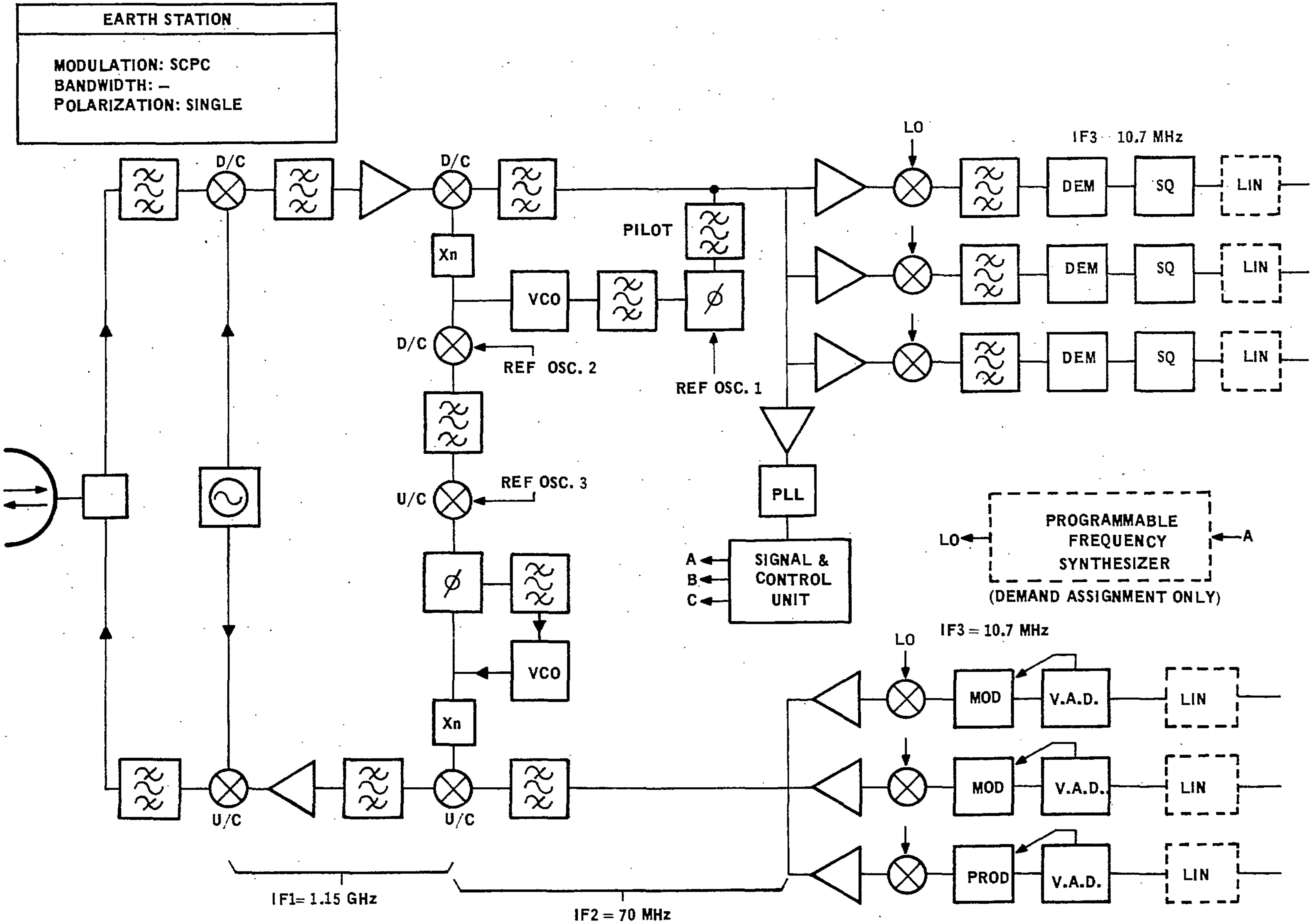


Figure 5-10 Earth Station

*　　*　　*

5.3.3 TECHNICAL SYSTEM TRADE-OFF RELATIONSHIPS

The systems equations are derived which will enable system trade-off to be carried out in later chapters.

Just as in the speech compressed FDM-FM-FDMA case, the required overall test-tone-to-noise ratio is 15 dB (unweighted). Allowing, as before, 50 percent of the noise budget to interference, the permissible noise due to thermal and multi-carrier intermodulation is therefore -18 dBm(0).

The two equations that need to be satisfied in a single-channel-per-carrier FM system are the fm equation that relates test-tone-to-noise ratio to the system parameters and the threshold equation[33]. These equations are as follows:

$$S/N = 10 \log \frac{3}{2} + 20 \log M + \left(\frac{C}{N_o}\right) - 10 \log fm + P_E \qquad (5.7)$$

where $\frac{S}{N}$ = test-tone-to-noise ratio (unweighted) = 18 dB

M = Modulation Index = $\frac{\text{Peak test tone deviation}}{fm}$

fm = voice bandwidth (= 3.4 kHz)

$\left(\frac{C}{N_o}\right)$ = Carrier-to-noise density before demodulator

P_E = pre-emphasis advantage (= 4 dB say).

$$\left(\frac{C}{N_o}\right)_{thr} = 14 + 10 \log (fm). \qquad (5.8)$$

Substituting $fm = 3.4 \times 10^3$ Hz into equation (5.8):

$$\left(\frac{C}{N_o}\right)_{thr} = 49.3 \text{ dB-Hz}. \qquad (5.9)$$

For a 99.99 percent system reliability due to propagation effects, the required down-link fade margin has been established as 6 dB, provided up-link power control is used, as has been the case throughout the previous discussions. Hence, the carrier-to-noise density at pre-detection is given approximately by the threshold carrier-to-noise density $(C/N_o)_{thr}$ plus 6 dB, as up-link noise contribution can be made negligibly small.

$$\therefore \left(\frac{C}{N_o}\right) = 49.3 + 6 = 55.3 \text{ dB-Hz} \qquad (5.10)$$

Substituting this $\left(\frac{C}{N_o}\right)$ and other parameters into equation(5.7):

$$20 \log M = -7.8$$

Solving for M:

$$M = 0.41$$

The i-f bandwidth required, from Carson's Rule*, is:

$$B = 2fm\ (1+M)$$

$$= 2 \times 3.4\ (1+0.4) = 9.5 \text{ kHz.}$$

The pre-detection total carrier-to-noise ratio is given by:

$$\left(\frac{C}{N}\right) = 55.3 - 10 \log 9.5 \times 10^3 = 15.5 \text{ dB.}$$

The above parameters are derived based on the assumption that the rf carriers are on all the time. Now with voice activation, the carriers are turned on only during the speech spurts. This results in an intermodulation advantage[30] of about 4 dB plus a power conservation advantage of also about 4 dB (i.e., activation factor 40 percent). Taking into account these voice activation advantages, the required TWT output back-off is found to be about 4.8 dB. This yields an effective down-link thermal carrier-to-noise density ratio of 58.3 dB-Hz. The bandwidth, the down-link carrier-to-noise density and TWT output back-off are the systems parameters needed in the trade-off computations of chapter 7.

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* *Note*: The validity of Carson's Rule for this very narrow bandwidth may need further examination.

6. REMOTE COMMUNICATIONS

This chapter will discuss the merits of the single-channel-per-carrier scheme and present the space and earth sequent configurations and the systems equations for the remote communications network.

The traffic requirements and the number of earth terminals for the remote communications service have been defined in chapter 2 as being 1200 channels and 400 earth stations respectively. In chapter 3, the antenna beam plans for this service are developed. There are only two beam plans, instead of three, since for the remote service Plan (A) is the same as Plan (C). In Plan (A) for the remote service, there are 5 medium-size beams which are basically the educational television service beams developed in the previous study[14]. In Plan (B) for the remote service, there are a total of 14 beams.

In this chapter will be presented the relative merits of multi-channel versus single-channel-per-carrier operation, the space and earth segment configurations and the system equations. System economic trade-off will be done in chapter 7.

* * *

6. Remote Communications

6.1 RELATIVE MERITS OF MULTI-CHANNEL VERSUS SINGLE-CHANNEL-PER-CARRIER (SCPC) SYSTEMS

The above relative merits are discussed. The SCPC scheme is recommended.

Generally, any radio communications system, including communications via satellite, can be divided into multi-channel-per-carrier and single-channel-per-carrier types. Further, in either type, current modulation techniques used are either digital or analog.

Communications to remote areas immediately imply that earth terminals at each location would have very small capacity, perhaps a few channels. Indeed in the traffic models adopted for this study, each remote earth terminal would have an average of only 3 channels. The total space segment capacity is of the order of 1200 channels, as the estimated number of earth terminals is about 400.

Consider first the case of a multi-channel system with digital modulation. If time division multiple access (TDMA) is used, then each earth terminal must have the capability to transmit a wide band signal and drive the transponder to saturation. On the receive side, the station G/T must also be such as to receive the digital signal at full bandwidth even if the earth station is going to extract only one single channel. In other words, the capability of the station that is necessary is not related to the station capacity. Over and above these drawbacks there is the requirement of high speed logic and memory devices. For the type of capacities the remote stations are expected to handle, the above approach is obviously not economic. If a frequency division multiple access (FDMA) approach is taken the earth station capability can be related somewhat to the channel capacity but there is the very severe problem of digital multiplexing and demultiplexing, let alone the fact that an FDMA scheme suffers from the well known transponder back-off penalty. To solve the multiplexing and demultiplexing problem would require certain amounts of equipment standardization; as for example multiplexing may be in groups of 12 channels. This imposes severe inflexibility to the network and is also wasteful in space segment capacity. Further, multiplexing equipment for small capacity systems is costly.

If analog modulation is used for a multi-channel system, only the FDMA scheme is feasible. The above remarks about the multiplexing problems apply here also.

In contrast to the above, the single-channel-per-carrier (SCPC) system has the following advantages:

a) Earth station capability in terms of transmit EIRP and receive G/T is related to the basic requirement of only one channel. Additional channels can be implemented by merely adding on similar equipment. Hence earth station cost can be directly related to capacity requirement.

b) Interfacing is easy since this is done at voice baseband. There is no multiplexing problem.

c) Flexibility is excellent. Demand assignment and traffic routing is achievable easily be selection of appropriate transmit and receive frequencies.

d) Circuit diversity can easily be engineered giving effective system reliability without excessive redundant equipment provision.

The SCPC technique has one chief disadvantage; that is, the need for network frequency stability on account of the very narrow channel bandwidth. The solution, however, is not difficult and a workable scheme has been presented in chapter 5.

From the above considerations it is obvious that for the remote system application, the SCPC technique is the preferred choice. An FDMA scheme is the easiest to implement. Although this still suffers from the transponder back-off penalty, the problem is made less severe by the fact that now every carrier is of the same power level (ignoring minor differences in fades, etc). Modulation could be either digital such as the delta modulation or analog, such as compandored or lincompex frequency modulation.

From the point of view of space segment capacity, calculations (not presented here) have shown that there is very little difference between delta modulation and compandored frequency modulation, given the equivalent channel performance of the two techniques. The lincompex, however, is known to be superior in performance to the compandor. Hence it is considered that the lincompex system is potentially the best choice and this is the system analyzed in this study.

6.2 SATELLITE AND EARTH STATION CONFIGURATIONS

The space and earth segment configurations are presented. The space segment block diagrams are explained in detail.

Since there are only two antenna beam plans for the remote service, only two models are examined here. As already explained, only the single-channel-per-carrier (SCPC) technique with lincompex speech compression and frequency modulation will be considered. Voice activation, as explained in chapter 5, is also used. These two models are designated as Z1/Z3 and Z2. The schematic block diagrams of the space and earth segments are shown in Figures 6-1 and 6-2 respectively. The SCPC operation of the earth segment is exactly the same as that given in chapter 5 and need not be repeated here. The space segment configurations, however, require some explanation and this is given below.

Refer to Figure 6-1a for the space segment block diagram of model Z1/Z3. The rf channels from the five antennas are combined by filters to feed a common low-noise amplifier (LNA). Each rf channel is then down-converted to a common i-f band by individual down-converters. The i-f band is then split by filters into five segments. One segment is taken from each up-link beam and these segments are added together. A total of five such assembled bands are thus formed. These are then up-converted by five individual up-converters before amplification by a common TWT. From the TWT output, the five bands are split by five rf filters to feed the five down-link antennas. The above arrangement achieves the following:

a) Every up-link beam is interconnected via 6 MHz segments to every down-link beam. This allows single hop communication throughout Canada. Routing of the SCPC carriers from any beam to any other beam is determined entirely by appropriate frequency selection from the ground. This achieves the double objective of demand assignment and effective circuit routing. The 6 MHz segment bandwidth is adequate for the traffic demands postulated.

b) Excluding redundancy, only one common LNA and TWT are used. This saving in electronic equipment is, of course, at the expense of antenna filtering that is now necessary. Since a reduction of electronic equipment tends to improve reliability, this approach is considered worthwhile.

Next, refer to Figure 6-2a for the space segment block diagram of model Z2. This model has 14 spot beams, and frequency plan N2 (see chapter 3) allocates a total of eight rf channels in frequency reuse application. For this model, it is not practicable to interconnect every beam to every other beam as in the previous case. In chapter 2, it is stated that in most likelihood, the traffic needs of any remote station in the North would be to neighboring remote stations or to a southern major population center. Since the beam plan here has beams shaped elliptically with the major axis along the general North-South direction, therefore an up-link rf channel looped back to the same beam on the down-link would meet most of the traffic needs. If communication to a remote station of another beam, is required, then a double hop connection via southern stations would be necessary. This scheme is definitely not as good as that of Model Z1/Z3. However, under the circumstances, it is a reasonable compromise solution.

From Figure 6-2a it can be seen that in beams numbered 1 to 8, the eight rf channels (one from each beam) are combined by rf filters and are fed to a common LNA. The whole spectrum is then down-shifted to the 12 GHz band by a down-converter and amplified by a common TWT. From the output of the TWT, the filters separate out the rf channels again and feed one channel each to the down-link antennas. In this way, an up-link rf channel is looped back to the same down-link beam. A similar scheme is needed for the beams numbered 9 to 14.

SYSTEM MODEL	Z1/Z3
SATELLITE CONFIGURATIONS	
UNSWITCHED MODULATION: SCPC BANDWIDTH:31.2 MHz POLARIZATION: SINGLE	

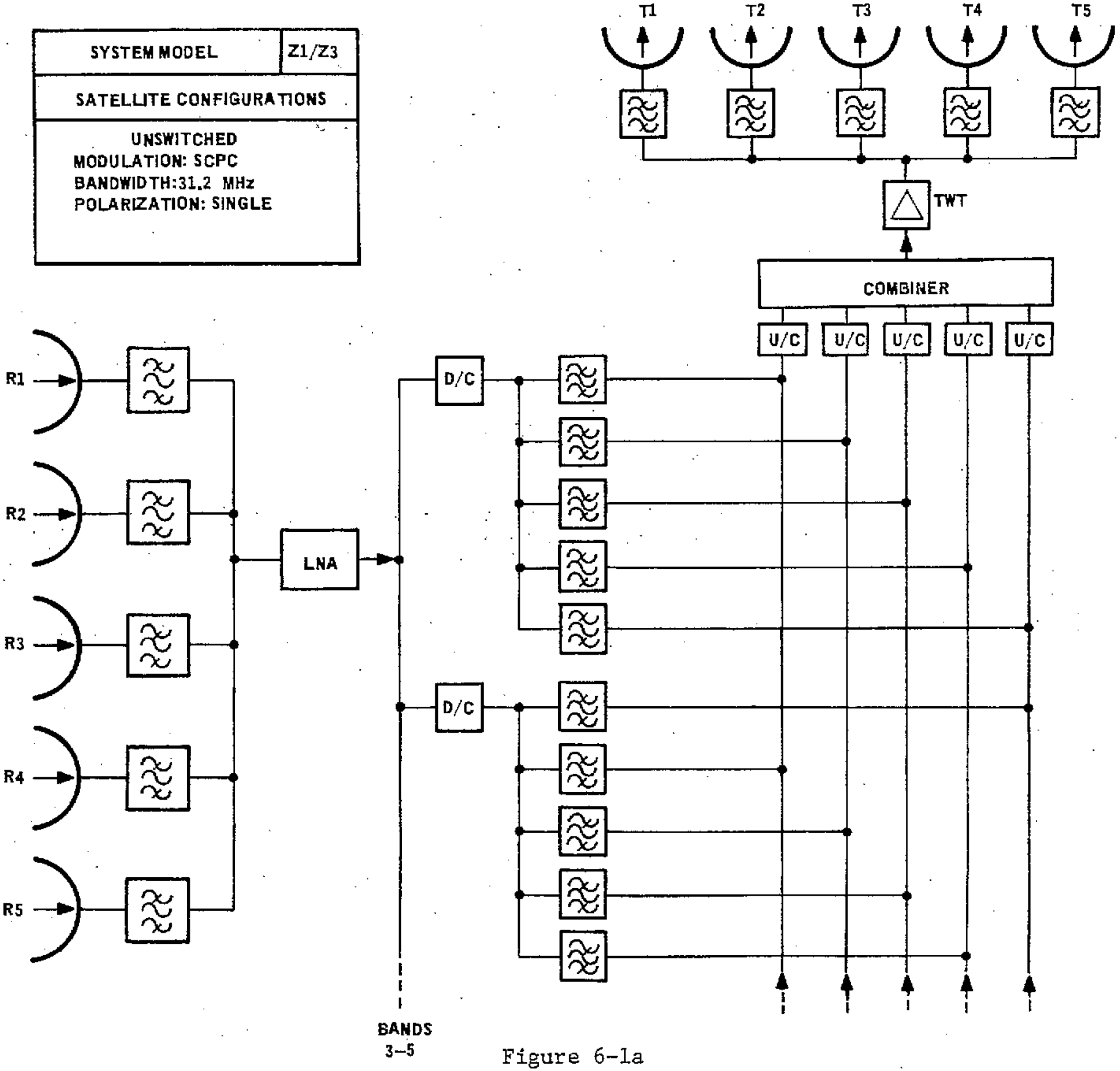

Figure 6-1a

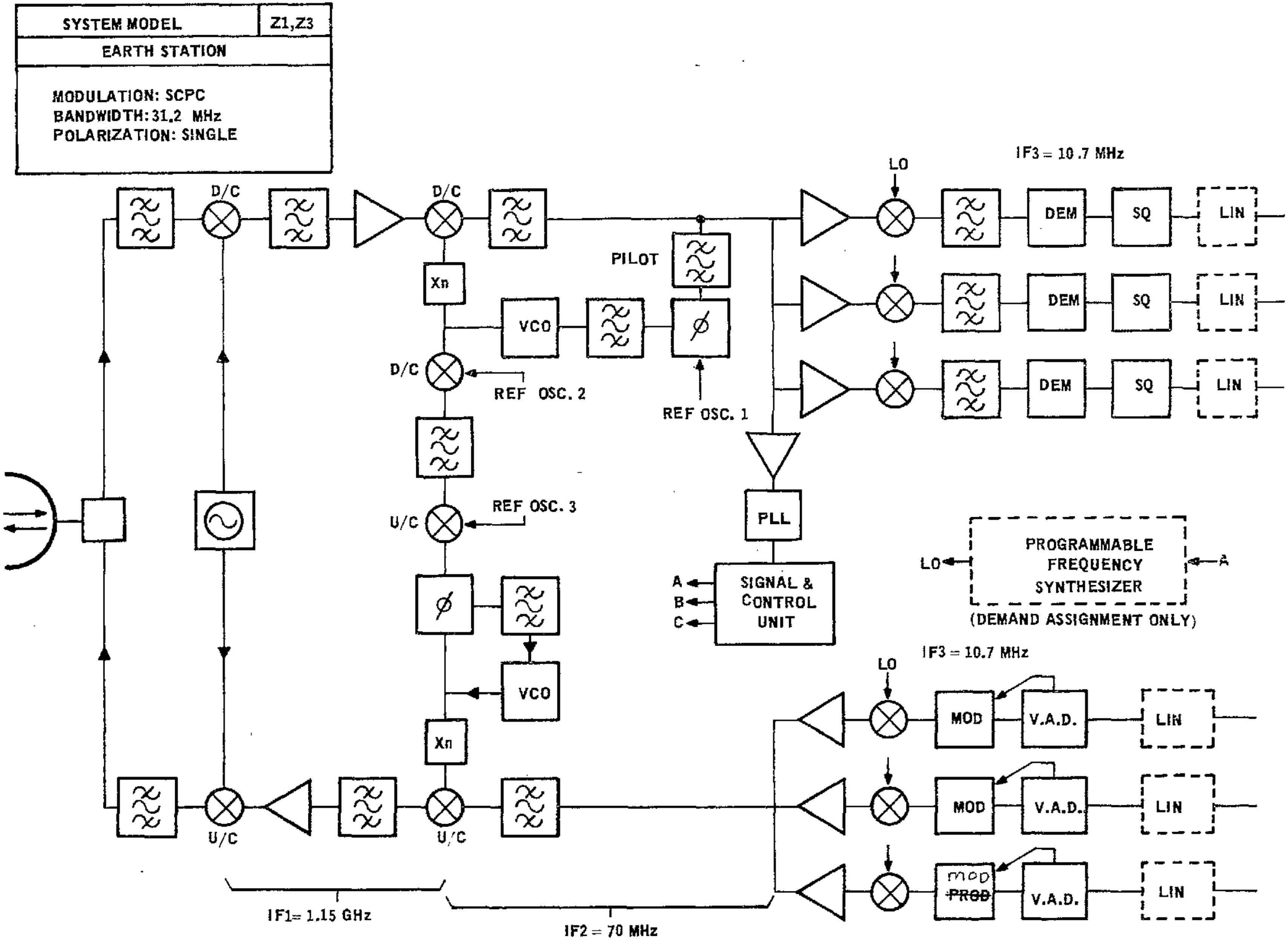
SYSTEM MODEL
Z1,Z3
EARTH STATION
MODULATION: SCPC
BANDWIDTH: 31.2 MHz
POLARIZATION: SINGLE
D/C
D/C
Xn
VCO
PILOT
D/C
REF OSC. 2
REF OSC. 1
REF OSC. 3
U/C
VCO
Xn
U/C
U/C
LO
IF3 = 10.7 MHz
DEM
SQ
LIN
DEM
SQ
LIN
DEM
SQ
LIN
PLL
SIGNAL & CONTROL UNIT
A
B
C
LO
PROGRAMMABLE FREQUENCY SYNTHESIZER
A
(DEMAND ASSIGNMENT ONLY)
IF3 = 10.7 MHz
LO
MOD
V.A.D.
LIN
MOD
V.A.D.
LIN
mop
~~PROD~~
V.A.D.
LIN
IF1 = 1.15 GHz
IF2 = 70 MHz

Figure 6-1b

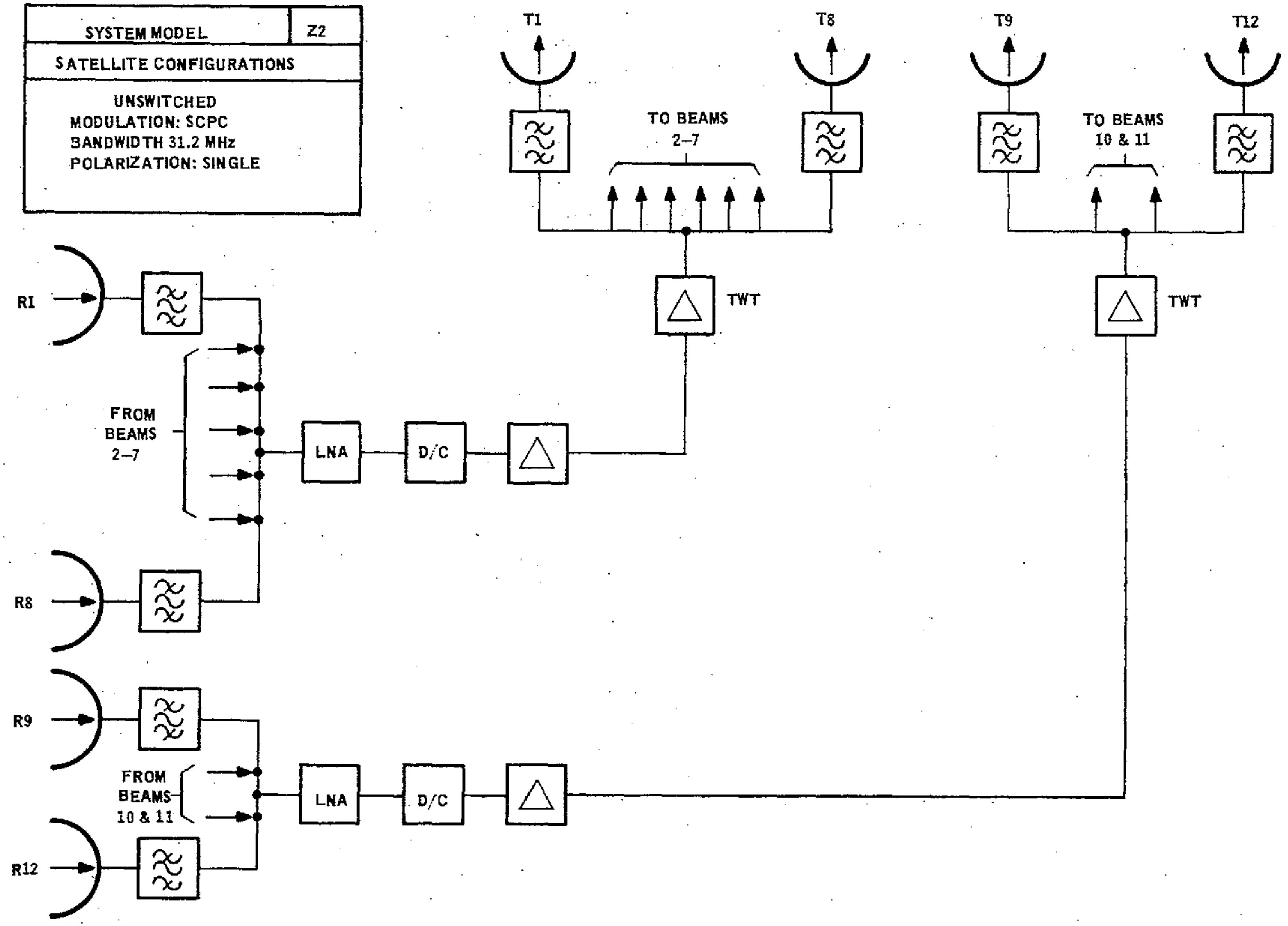
SYSTEM MODEL
Z2
SATELLITE CONFIGURATIONS
UNSWITCHED
MODULATION: SCPC
BANDWIDTH 31.2 MHz
POLARIZATION: SINGLE
T1
T8
T9
T12
TO BEAMS 2–7
TO BEAMS 10 & 11
TWT
TWT
R1
FROM BEAMS 2–7
LNA
D/C
R8
R9
FROM BEAMS 10 & 11
LNA
D/C
R12

Figure 6-2a

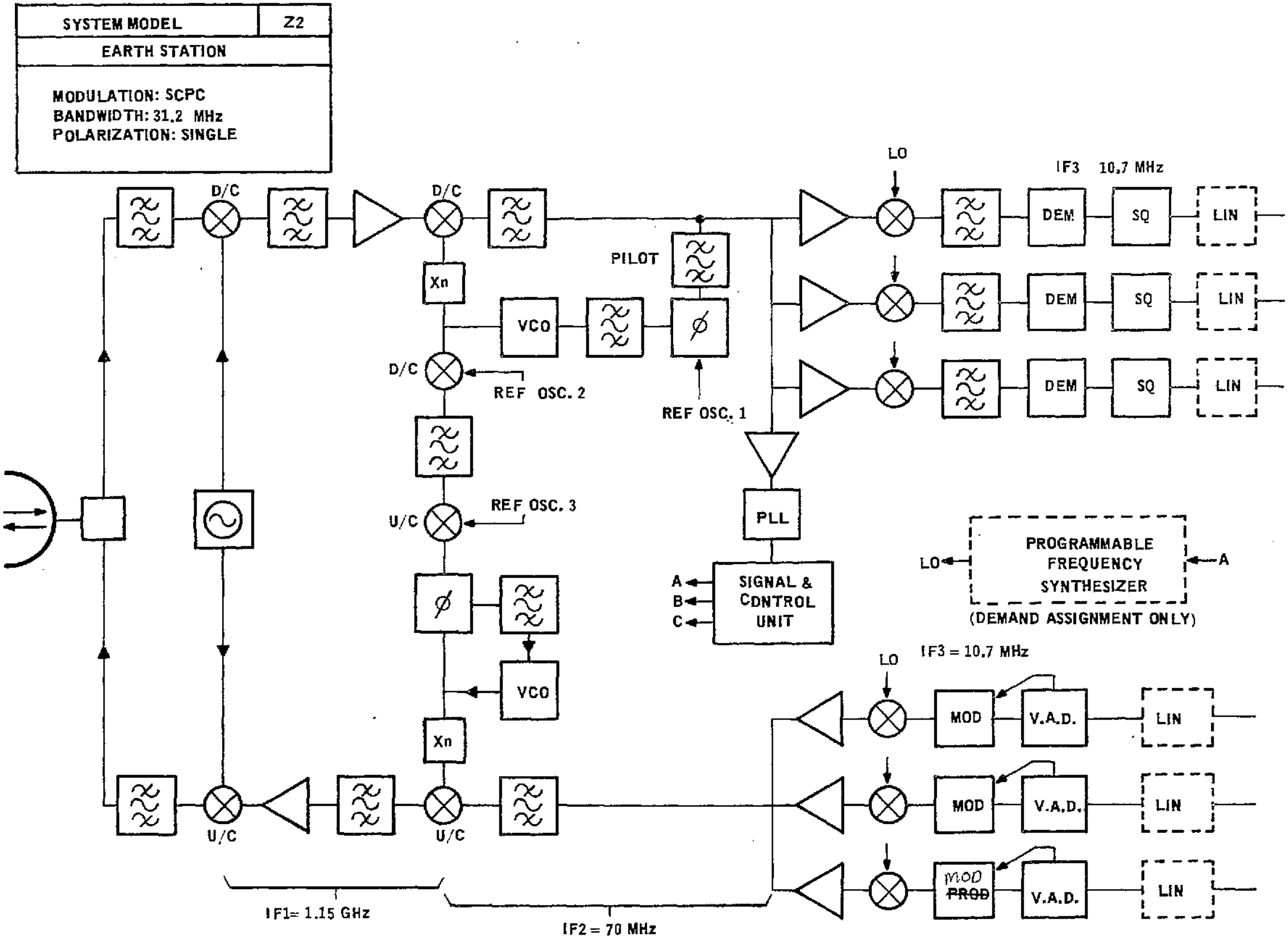
SYSTEM MODEL
Z2
EARTH STATION
MODULATION: SCPC
BANDWIDTH: 31.2 MHz
POLARIZATION: SINGLE
D/C
D/C
Xn
VCO
PILOT
D/C
REF OSC. 2
REF OSC. 1
REF OSC. 3
U/C
VCO
Xn
U/C
U/C
IF1= 1.15 GHz
IF2 = 70 MHz
LO
IF3 10.7 MHz
DEM
SQ
LIN
DEM
SQ
LIN
DEM
SQ
LIN
PLL
A
B
C
SIGNAL & CONTROL UNIT
LO
PROGRAMMABLE FREQUENCY SYNTHESIZER
A
(DEMAND ASSIGNMENT ONLY)
LO
IF3 = 10.7 MHz
MOD
V.A.D.
LIN
MOD
V.A.D.
LIN
MOD ~~PROD~~
V.A.D.
LIN

Figure 6-2b

6.3 TECHNICAL SYSTEM TRADE-OFF RELATIONSHIPS

The systems parameters needed for system trade-off are derived.

In chapter 2 it is stated that for the remote services a noise objective of 44 dBrnCo is adequate. With lincompex speech compression, this objective can easily be met or exceeded when the system has a test-tone to noise ratio of 15 dB (see Appendix C). Reducing this test-tone to noise ratio any further may not be feasible on account of the effect of noise on the link control channel of the lincompex system. Hence the same criterion as used in the SCPC system for the major traffic network as detailed in chapter 5 will be used. Equation (5.9) developed previously may therefore be applied here.

For the remote services, as indicated in the previous study[14], a propagation reliability of 99.9 percent is considered adequate. Again, from the previous study, the down-link fade margin corresponding to this reliability is estimated to be 4.7 dB. For the remote service no up-link power control is postulated. Hence, under normal operation and using equation (5.9), the pre-detection carrier-to-noise density ratio is given by:

$$(\frac{C}{No}) = 49.3 + 4.7 = 54.0 \text{ dB-Hz.}$$

Using equation (5.7) and substituting the other parameters as given in chapter 5 into this equation:

$$20 \log M = -6.5$$

Solving,

$$M = 0.47$$

The required i-f bandwidth, from Carson's Rule, is

$$\begin{aligned} B &= 2fm\ (1+M) \\ &= 2 \times 3.4\ (1+0.47) \text{ kHz} \\ &= 10 \text{ kHz.} \end{aligned}$$

The pre-detection total carrier to noise ratio is:

$$\begin{aligned} (\frac{C}{N}) &= 54 - 10 \log 10 \times 10^3 \\ &= 14 \text{ dB.} \end{aligned}$$

As in chapter 5, taking into account the intermodulation and power conservation advantages of voice activation, and using the well known TWT back-off characteristics, the optimum output back-off is determined to be 4.0 dB. The effective down-link carrier-to-noise density ratio with voice activation is then calculated to be 56.2 dB-Hz. The bandwidth, the down-link carrier to noise density and TWT back-off are the systems parameters needed in the trade-off computations of chapter 7.

7. System Economics

7.1 GENERAL

The economic analysis of the various system models has been carried out in terms of total annual charges. Each model has been optimized using a computer program to determine the most economical variant. The allowable range of variation of the technical parameters on either side of the optimum, for an increase of 5 percent in Total Annual Charge, has also been determined as an indicator of the degree of freedom available in each case.

The previous discussions have covered the technical aspects of multibeam communications satellite systems, and have identified and detailed the technical constraints. Within these constraints there remains a large range of choice between many possible configuration variants and many interrelated technical parameters. The final choice must be made on the basis of lowest cost for service provided. As in the previous study, Total Annual Charges have been used as the economic indicator; although the currently most popular indicator is the Present Worth of Annual Charges (PWAC), it is preferred to use Total Annual Charges, assuming 1972 dollars throughout, since this simplifies the process and provides valid comparative figures without making any assumptions about inflation rates, profits, taxes, etc.

Since space segment costs increase with satellite EIRP, and earth segment costs increase with earth station G/T, the sum of space and earth segment costs has a minimum at some particular value of earth station G/T. A computer program has been compiled which locates this minimum point for any desired system, and this has been used to determine the preferred variant of each system configuration. The final choice may be influenced by other factors (such as maximum satellite weight allowable, or maximum earth station antenna desirable) and to enable this to be done the key parameters have been determined over a range of Total Annual Charges between the minimum and an arbitrarily chosen level of 5 percent above minimum.

The following paragraphs 7.2 and 7.3 consider the component costs for the space and earth segments respectively. These constitute fixed inputs to the program. The results obtained when the appropriate variable inputs for the various system configurations were fed into the program are given in 7.4.

* * *

7.2 SPACE SEGMENT

The total cost of the space segment is made up of development, production, and launch costs.

The total cost of the "space segment" is made up of two major items,

1) Satellite Production (including start-up costs and hardware)

2) Launch costs

Satellite manufacturing is discussed in 7.2.1 with some comments on development costs.

Spacecraft launch costs are considered in 7.2.2.

*   *   *

7.2.1 SATELLITE COSTS

Total spacecraft cost is made up of development costs (amortized) and production costs. Development of three satellites of the type considered in this study is expected to amount to $10 million.

Spacecraft production hardware is assumed to cost $10 000 per pound for a set of three spacecraft in the Thor-Delta launch class.

Development

The development of a spacecraft includes acquisition or development of the required technology, design and development of engineering models, and prototype production and testing. Program direction and overhead charges must also be allowed for.

The total cost of development of the class of satellite with which we are concerned was considered in the previous study (9.2)[14].This work has been reviewed in the light of recent information and appears to be still valid. The conclusions can be summarized as follows:

By 1977, satellite technology in the 12 GHz through 15 GHz region will have been largely developed through ATS, Communications Technology Satellite, and U.S. domestic and military programs, as well as terrestrial systems. The necessary refinement to antenna, solar array and body stabilization techniques will also have been accomplished during some of these programs. It is therefore concluded that, because much of the basic technology necessary for the development of a body stabilized satellite of the weight class envisaged would have been done, the development costs associated with a program initiated in 1975/85 would be appreciably less than those necessary to develop a spacecraft when most of the technology was not available and most of the available components not flight qualified.

The communications subsystem constitutes about 30 percent of the spacecraft cost and will require specific development. The remaining 70 percent is made up of housekeeping and power subsystems which are assumed to be available on payment of royalties of 10 to 15 percent; a further 10 to 15 percent is estimated to be needed for integration.

In view of the factors mentioned, it may be expected that the development program would cost between 40 and 50 percent of a program to develop the spacecraft from scratch. Intelsat IV, which may be considered typical, cost \$22 million to develop, and hence the development cost for the new satellite would be around \$10 million. If a figure of \$2.5k per pound of hardware is used (as suggested by NASA officials for the appropriate weight class and technology status in a repetitive case) a program producing three satellites would be expected to cost about \$9 million, to which royalties need to be added, making a total again around \$10 million. This figure may therefore be used with some confidence.

Production

Production cost has been estimated on a per-pound basis. While at first sight this method of calculation would seem somewhat arbitrary and a poor substitute for detailed costing, it has proved in practice to be the best method for predicting the cost of complex equipment such as spacecraft, and is used extensively by both NASA and COMSAT[26].

In the previous study the figure of \$10 000 per pound of spacecraft total weight was used for production cost estimates. On this basis, three 1000-pound satellites would cost \$30 million. A subsequent paper published by COMSAT[26] derives a square root relationship between cost and weight ($0.9 \sqrt{W_s}$ for non-recurring costs and $0.22 \sqrt{W_s}$ for recurring costs, where W_s = total satellite weight.) Using these figures, three 1000-lb spacecraft would be estimated at about \$49 million. The COMSAT figures, however, are based largely on Intelsat spacecraft (which are abnormally costly because of the conditions of international participation) and scientific satellites (which have high production costs because of the small numbers involved - usually one off). In view of these factors, it is felt that for our purposes the figure of \$30 million for three spacecraft is more realistic, and will again be used. Further confirmation that this figure is reasonable is afforded by the recent Western Union order of three HS333 satellites from Hughes Aircraft Corporation (HAC).

It was mentioned that the COMSAT study derived a square root relationship between cost and mass of complete integrated satellites. This function cannot be used here since it would lead to a situation where services taking up a small proportion of the total weight would be at an unreal disadvantage when their cost liability was calculated. Since we have assumed that the satellite design finally arrived at will be approximately optimized for Thor-Delta launch, it is legitimate (and in fact imperative) to use a linear cost/weight relationship within the spacecraft, even though a different function may apply to total spacecraft weight when considering alternative launch vehicles.

Production costs of the communication subsystem required for a particular service utilizing the various system configurations have been determined by breaking the subsystem down into a number of components each of which has a unit weight and a unit power consumption assigned to it. The power consumption figures are used to calculate the weight of the power subsystem. All the unit weights are then multiplied by the appropriate quantities to give the total payload weight contribution, from which is derived the total spacecraft weight assignable to the particular service. The parameters used are listed in Table 7-1.

Table 7-1

ITEM	W(lbs)	P(Watts)
L.N.A.	0.5	1.0
Downconverter complex	1.0	2.0
Upconverter + TWTA	6.4	*
Upconverter - (SS output)	5.4	*
Crosspoint (each)	0.2	0.02
On-board C.P.U.	20.0	20.0
R.F. filter (each)	0.5	--
Horn feed	2.0	--
Payload/Satellite Weight Ratio		0.45
Upconverter + TWTA dc efficiency η		0.25
Upconverter (SS output) efficiency η		0.10
Harness, coax and fittings		10% of Payload
Antenna mass density		0.2 lb/ft^2
Space ratio (redundancy)		0.3

* Calculated from dc conversion efficiency (7)

* * *

7.2.2 LAUNCH COSTS

A figure of $8000 per pound in synchronous orbit has been allowed for launch costs.

Throughout this study the assumption has been made that a Thor-Delta launch vehicle would be used. In practice this has not acted as a constraint, since all the systems considered have involved spacecraft weight allowances below the maximum permissible for Thor-Delta launch. It is assumed that other services such as CTV or ETV would be added to make the spacecraft weight up to the Thor-Delta capability (see Statement of Work, chapter 1).

The predicted launch cost in 1977 for this vehicle is $7 to 8M, with a synchronous orbit capability of 1000 to 1200 pounds. A conservative estimate of launch cost per pound is therefore $8000, and this figure has been used in the calculations. The conservative view has been taken since the final design of satellite may not end up with a launch weight which is exactly optimum for the vehicle.

* * *

7.3.1 FIXED COSTS

In the case of the large (major traffic) earth stations, a capital allowance of $280 000 has been made to cover buildings, civil works, power, test equipment, documentation and other fixed costs.

The allowance for small (remote) stations is $20 000.

The fixed costs associated with large earth stations vary widely, depending on terrain, architectural and aesthetic considerations, accessibility of reliable prime power, sharing of the site or buildings with other services, and other factors. Large Intelsat station building and site costs are typically $900 000. Domestic stations in the unattended category have been estimated to require $80 000 in building and site costs.

Power back-up will be required, but since reliable prime power should be available on all the Canadian sites the facilities will be in the form of standby generators and supply continuity equipment. $50 000 has been allocated.

Finally, an allowance of $150 000 had been made for installation, test equipment, spares and documentation, bringing the total "fixed" costs to $280 000 for each large station.

In the case of the remote stations, a figure of $20 000 has been used as an average fixed cost. This assumes that simple equipment enclosures or shelters would be used, or the equipment would share existing buildings.

* * *

7.3.2 ANTENNA COSTS

The earth station antenna size and complexity varies with earth station G/T. The rules used for estimating the antenna cost are stated.

The costs of various sizes of antenna were determined during the previous study (see 9.3.2)[14]. The rules used for estimating antenna cost as a function of aperture were derived. These rules are repeated below:

- Up to 17 feet diameter:

Spun reflector
No tracking
Cost = $17.25\ D^{1.85}$ dollars each (qty. 50+)
where D is the reflector diameter in feet

- 17 to 30 feet diameter:

Conventional fabrication
Simple tracking (step-track)
Cost = $65\ 000 + 330\ D^{1.5}$ dollars each
where D is the reflector diameter in feet

- 30 to 60 feet diameter:

Conventional fabrication
Precision tracking
Cost = $200\ 000 + 670\ D^{1.53}$ dollars each
where D is the reflector diameter in feet

These rules have again been used in the cost program compiled for this study. Some caution should be exercised when considering antennas in the vicinity of 30 feet diameter. Tracking systems which might be described as "semi-precision" may bring the cost of the larger antennas below the figure derived from the above relationship. Since the change from step to precision tracking occurs at a point where some of the system configurations are approaching their economic optima, this factor is of importance; for this reason the system economic optimum point should be examined in each case and where it falls at the point of changeover, this should be borne in mind. Further reference to this matter will be made in section 8 where the comparative economics of different systems is discussed.

* * *

7.3.3 EQUIPMENT COSTS

The components of the earth station equipment have been lumped together into "black boxes" defined in such a way as to permit the computer program to assemble earth stations for the various system configurations. The cost figures used for these component "black boxes" are given.

The price of an earth station varies considerably depending on the system configuration chosen. In order to obtain reasonable comparative estimates without designing a multiplicity of possible stations from scratch, a number of subsystem "black boxes" were defined and used as building blocks. Although some variation in subsystem costs occur between applications, these variations are minor and a typical cost figure has been assigned to each. The computer then has only to take these figures (or in some cases, calculate them from other input parameters) and multiply them by the number of units required in each case.

The equipment building blocks chosen, and the fixed costs in dollars assigned to each, are given below:

LNA (uncooled parametric amplifier) :	30 000	
LNA (remote station) (TDA) :	2 000	
Down-converter/demodulator :	12 000	
Modulator/up-converter :	14 000	
Mode transducer, waveguide, etc., (large stations) :	30 000	
TDMA timing and control equipment :	190 000	
Multiplex (analog) :	1 000	per channel
Multiplex (digital) :	750	per channel
Lincompex unit :	500	per channel
SC/C interface cost (large station) :	50 000	
SC/C channel unit	1 500	per channel

The above figures are derived from current sources and are typical. The last item (single-channel-per-carrier channel unit) is perhaps questionable, since this is a completely new item. The cost figure that has been used assumes that by 1977 the necessary technology will be widely available, and the quantity of units involved will run into tens of thousands. If the quantities involved are appreciably less, and if full development costs were to be amortized over these, the unit cost would be much higher - perhaps $5000 each. On the other hand, if similar units are manufactured on a very large scale (e.g., for the U.S. domestic system) the unit cost might well be less.

The rules used for variable costs are given below (the calculations are carried out within the program):

1) Antenna (See 7.3.2):

If $D_{E/S} < 17$, $C_A = 17.25\ (D_{E/S})^{1.85} : 1000$

If $17 < D_{E/S} < 30$, $C_A = [65\ 000 + 330\ (D_{E/S})^{1.50}] : 1000$

If $30 < D_{E/S}$, $C_A = [200\ 000 + 670\ (D_{E/S})^{1.53}] : 1000$

where $D_{E/S}$ is earth station antenna diameter (feet)

C_A is earth station antenna cost ($ thousands)

2) Power Amplifier:

If $1 < P < 20$ watts, $C_{HPA} = 10$

If $20 < P < 300$ watts, $C_{HPA} = 20$

If $300 < P < 1000$ watts, $C_{HPA} = 35$

If $P > 1000$ watts, $C_{HPA} = 40$

where P is transmitter rated power (watts)

C_{HPA} is transmitter cost ($ thousands).

No allowance has been made for backhaul facilities from the earth stations. This is because the use of exclusive frequency bands removes the interference problems encountered in the 4GHz/6GHz bands. This problem compels earth stations in the latter bands to locate remotely from the cities they serve, whereas in this case no constraint exists and the stations may be sited adjacent to their switching centres.

7.4 TOTAL SYSTEM COSTS

The method used to compute total system costs is explained.

In order to find the total relative cost of each system, a computer program was compiled. The inputs to this program were in two groups: data common to all systems, and data peculiar to particular systems. The former group contained such items as unit costs, and the latter such items as spacecraft antenna gain and required quantities of various units.

The program carried out the basic system calculations and computed the total capital costs of earth and space segments, total annual charges for each, and the sum total for each system. Development charges were not included. The calculations were repeated for a range of earth station (G/T)'s from 20 dB/°K to 40 dB/°K for the heavy route stations, and from 16 dB/°K to 36 dB/°K for the remote stations. The basic operation of the program is described below:

Simplified Program

Input: Earth Station system noise temperature (T), and set earth station (G/T) = 16 or 20

Calculate: $G_R = (G/T) + T$ (7.1)

Calculate: $G_T = G_R + 1.5$ (7.2)

Input: RF channels per satellite PA (N_S)

Input: Required $[\frac{C}{N_o B}]$ and bandwidth (BW)

Calculate (if $N_S = 1$): $EIRP_{SAT} = \frac{C}{N_o B} + 10 \log BW - (G/T) - 21.9$ (7.3)

Calculate (if $N_S > 1$):

$EIRP_{SAT} = 58.3 + 10 \log N_S - (G/T) - 21.9$ (SCPC Major) (7.4a)

$EIRP_{SAT} = 56.2 + 10 \log N_S - (G/T) - 21.9$ (SCPC Remote) (7.4b)

Input: Satellite antenna transmit gain (G_{SD}), number of carriers per transponder or up-converter (Q) and back-off factor (BOF)

Calculate: $P_o = Q.BOF.10^{(EIRP_{SAT}-G_{SD})/10}$ (7.5)

Input: RF channels per earth station PA (N_G) and satellite (G/T)

Calculate: (if $N_G=1$) $EIRP_{ES} = 5.5 + EIRP_{SAT} + (G.T) - (G/T)_{SAT}$ (7.6)

Calculate: (if $N_G>1$) $EIRP_{ES} + 73.4 + 10 \log N_G - 20.4 - (G/T)_{SAT}$ (7.7)

Input: Back-off factor (earth station) (BOF_{ES})

Calculate: $P_{TX} = N_G.BOF_{ES}.10^{(EIRP_{ES} - G_T + 14)/10}$ (7.8)

Input: Number of spot beams (X), number of sub-units per beam (A, B, C, M), unit power consumptions (P_{LNA}, P_{DC}, P_M, P_{CPU})

Calculate: $P_T = X(A.P_{LNA} + B.P_{DC}) + X.M.\ P_o/\eta_o + C.P_M + P_{CPU}$ (7.9)

Input: Antenna size (d_1 x d_2), area constant (K_1), mass density (ρ), unit weight of feed horn (W_F)

Calculate: $W_{AR} = 0.135\ P_T$ (7.10)

Calculate: $W_{ES} = P_T \div 12$ (7.11)

Calculate: $W_A = \pi.\rho.K_1.d_1 d_2 + X.W_F$ (7.12)

Input: Spare ratio (S), number of filters (N_{RF}), and unit weights (W_{LNA}, W_{DC}, W_M, W_{UC}, W_{CPU})

Calculate: $W_T = X(1 + S).(A.W_{LNA} + B.W_{DC}) + C.W_M +$

$X.M(1 + S)\ W_{DC} + W_{CPU} + N_{RF}.W_{RF}$ (7.13)

Calculate: $W_T = W_T \times 1.1$ (7.14)

Input: Ratio payload mass/satellite mass (Y)

Calculate: $W_{SAT} = (W_{AR} + W_{ES} + W_A + W_T)\ 1/Y$ (7.15)

Input: Cost/lb (Cpp)

Calculate: $C_{TS} = C_R + Cpp.W_{SAT}.N_{SAT}$ (7.16)

Calculate: $D_{ES} = 10^{(G_R-29)/10}$ (7.17)

Calculate: C_A (cost of earth station antenna) and C_{HPA} (cost of earth station HPA) in accordance with rules given previously

Input: Unit costs and quantities of earth station components (C_{LNA}, N_{LNA}, C_{DC}, N_{DC}, N_{HPA}, C_{UC}, N_{UC}, C_{CON}, C_{MUX}, C_{MODE})

Calculate:

$$C_{ES} = C_{LNA}.N_{LNA} + C_{DC}.N_{DC} + C_{HPA}N_{HPA} + C_{UC}N_{UC} + C_{CON} + C_{MUX} + C_{MODE} \quad (7.18)$$

Input: Number of earth stations (N_{ES}), fixed costs (C_{FIX}) number of voice SCPC channels (N_{VOC})

Calculate: $C_{TE} = (C_{FIX} + E_{ES} + C_A) \, . \, N_{ES} + N_{VOC}.N_{ES}.1.5$ (7.19)

Calculate: $C_{SA} = 0.18 \; C_{TS}$ (7.20)

Calculate: $C_{EA} = 0.28 \; C_{TE}$ (7.21)

Calculate: $C_{ANNUAL} = C_{SA} + C_{EA}$ (7.22)

Add 2 to earth station G/T and repeat for desired range.

Notes on Simplified Program

Equation	Note
(7.1)	Derives earth station receive antenna gain from (G/T) and T.
(7.2)	Adds 1.5 dB to earth station receive gain to allow for frequency difference between up-link and down-link.

Equation	Note
(7.3)	Calculates necessary satellite transponder EIRP to provide required carrier to noise ratio (C/N_0B). The constant (21.9 dB) is the difference between the down-link path loss and Boltzmann's constant.
(7.4)	As (7.3), in the single channel per carrier cases (configurations Y and Z). Derivation of the constant 57.3 dB may be found in chapter 5.
(7.5)	Calculates satellite transponder RF power output required.
(7.6)	Calculates the earth station EIRP from EIRP_{SAT}. Derivation of the constant 5.5 dB is given in section 4 (equation 4.9).
(7.7)	Carries out the earth station EIRP calculation for configurations Y and Z (see chapter 5).
(7.8)	Calculates the required earth station transmitter power rating, taking into account back-off factor, up/down link ratio and down-link margin (see chapter 4).
(7.9)	Calculates total power consumption of satellite.
(7.10)	Calculates weight of solar array from total power consumption.
(7.11)	Calculates battery weight from total power consumption.
(7.12)	Calculates antenna weight from dimensions, material density and feed horn unit weight.
(7.13)	Calculates total payload weight, taking into account redundancy ratio (taken as 1:3).
(7.14)	Adds 10 percent allowance for cabling, waveguide and fixtures.
(7.15)	Calculates total satellite weight.
(7.16)	Calculates total space segment cost.
(7.17)	Computes earth station antenna diameter from receive gain.
(7.18)	Sums earth station costs.
(7.19)	Calculates total earth segment costs (including single-channel-per-carrier modem units @ \$1.5k each).

Equation	Note
(7.20) and (7.21)	Computes annual charges for space and earth segments respectively.
(7.22)	Sums annual charges of space and earth segments.

Symbols Used in Program

G/T	Earth station (G/T) in dB/1°K.
T	Earth station system noise temperature in °K.
N_S	Number of voice channels per satellite P.A. (configurations Y and Z).
$\frac{C}{N_oB}$	Carrier to noise ratio (dB).
BW	Effective bandwidth (MHz).
Q	Number of rf carriers per transponder.
BOF	Back-off factor for satellite output stage
$(G/T)_{SAT}$	Satellite (G/T) in dB/°K.
N_G	Number of rf channels per earth station HPA.
BOF_{ES}	Back-off factor (earth station output stage).
X	Number of spot beams.
A	Number of LNA's per beam.
B	Number of down converters per beam.
C	Number of crosspoints per beam.
M	Number of transponders (or up-converters) per beam.
P_{LNA}	Power consumption of each LNA (W)
P_{DC}	Power consumption of each down converter (W)
P_M	Power consumption of each crosspoint drive (W)
P_{CPU}	Power consumption of each on-board processor (W)

ρ	Antenna mass density (lb/ft^2)
K_1	Conversion factor from aperture to surface area for antenna with F/D = 0.4
d_1, d_2	Major and minor dimensions of antenna (ft)
W_F	Weight of each feed horn (lb)
S	Spares (redundancy) ratio
N_{RF}	Number of rf (i-f) filters
W_{RF}	Weight of each rf (i-f) filter (lb)
W_{LNA}	Weight of each LNA (lb)
W_{DC}	Weight of each downconverter (lb)
W_M	Weight of each crosspoint and drive (lb)
W_{UC}	Weight of each upconverter (lb)
W_{CPU}	Weight of each on-board processor (lb)
Y	Ratio payload/satellite total weight
Cpp	Cost per pound ($1000's)
C_{LNA}	Unit cost of earth station LNA ($1000's)
C_{DC}	Unit cost of earth downconverter ($1000's)
C_{UC}	Unit cost of earth upconverter ($1000's)
C_{CON}	Unit cost of earth timing and control ($1000's)
C_{MUX}	Unit cost of earth multiplex ($1000's)
C_{MODE}	Unit cost of earth mode couples and waveguide ($1000's)
N_{LNA}	Number of earth station LNA's
N_{DC}	Number of earth station downconverters
N_{UC}	Number of earth station upconverters
N_{HPA}	Number of earth station rf power amplifiers
N_{ES}	Number of earth station in system

C_{FIX}	Fixed charges (buildings, power, installation, test, documentation etc.) ($1000's)
N_{VOC}	Number of voice channels (used in configurations Y and Z only)
$EIRP_{SAT}$	Satellite EIRP (dBW)
P_O	RF output power per transponder or upconverter (W)
$EIRP_{ES}$	Earth station EIRP (dBW)
P_{TX}	Earth station rf output power per HPA (W)
P_T	Satellite dc power consumption (W)
W_{AR}	Weight of solar array (lb)
W_{ES}	Weight of storage battery subsystem (lb)
W_A	Weight of satellite antenna (lb)
W_T	Weight of communications subsystem (lb)
W_{SAT}	Total satellite weight (in orbit) (lb)
C_{TS}	Space segment capital cost ($1000's)
C_R	Space segment development cost (if included) ($1000's)
N_{SAT}	Number of satellites
D_{ES}	Earth station antenna diameter (ft)
C_A	Earth station antenna cost ($1000's)
C_{HPA}	Earth station rf amplifier cost ($1000's)
C_{ES}	Earth station electronics total cost ($1000's)
C_{TE}	Total cost of earth segment annual charges ($1000's)
C_{EA}	Earth segment annual charges ($1000's)
C_{ANNUAL}	Total annual charges ($1000's)

The following sections present the results obtained when the program described above was used to compute the costs of the various multi-beam satellite systems under consideration.

* * *

7.4.1 MAJOR TRAFFIC: SWITCHED MODELS

The trade-off curves for the three SDMA/SS-TDMA models are given.

The space segment, earth segment, and total system annual charges for the SDMA/SS-TDMA on-board switched models (system configuration U_1, U_2 and U_3) are plotted as a function of earth station G/T in Figures 7-1, 7-2 and 7-3 respectively. Twelve earth stations, three satellites, and a total voice channel capacity of 24 000 (with ground equipment for 10 000 circuits) have been assumed. It will be seen that in each case a fairly pronounced minimum in the curve of total annual charges occurs at an earth station G/T value of 34 dB/°K. This corresponds with the point at which the earth station antenna changes from step to precision tracking in the cost equation. As mentioned previously, this point is likely to become less definite if and when "semi-precision" tracking systems evolve, and therefore the precise earth station G/T at which the minimum occurs in the curve may shift somewhat to the right. An indication of the effect of permitting a 5 percent in Total Annual Charges is given at the bottom of each curve.

The sensitivity of the total figures to changes of 20 percent up or down for the space segment of one of the configurations (U_3) is displayed in Figure 7-4, and for similar changes in the earth segment in Figure 7-5.

The sensitivity curves indicate how the Total Annual Charges change with variations in space and earth segment costs respectively. It should be noted that such changes affect the totals only; the optimum remains at the same point.

It will be noticed that the four-beam configuration (U_3) yields the least Total Annual Charges by a small margin, while the 8-beam arrangement (the heavy route contribution to the 14-beam total system U_2) is markedly more expensive. The sensitivity of the "U" system costs to the number of spot beams - disregarding traffic distribution factors - was examined, and the results confirm that for these configurations four beams in fact provide the most economic arrangement for a given total channel capacity even if other factors are ignored. More discussions of the relative merits and costs of the various beam models will be presented in chapter 8.

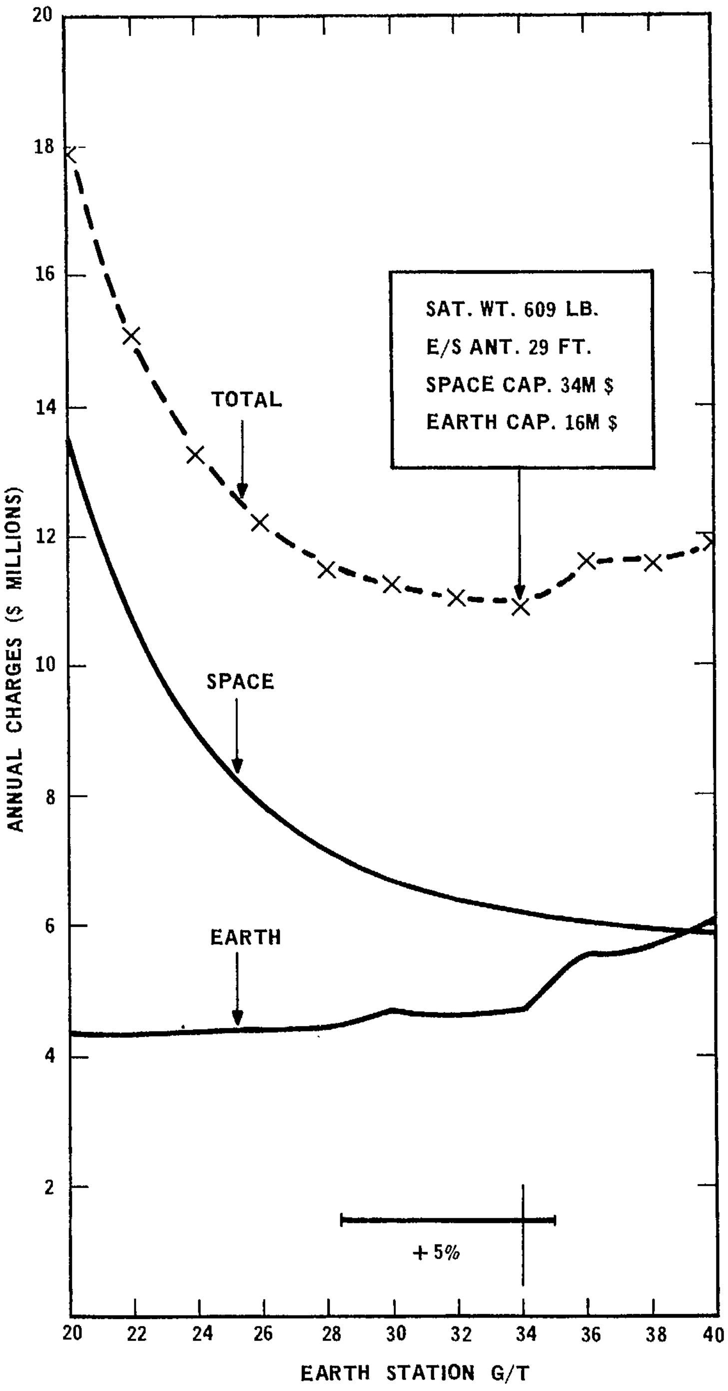

Figure 7-1 Configuration U_1, SDMA/SS-TDMA

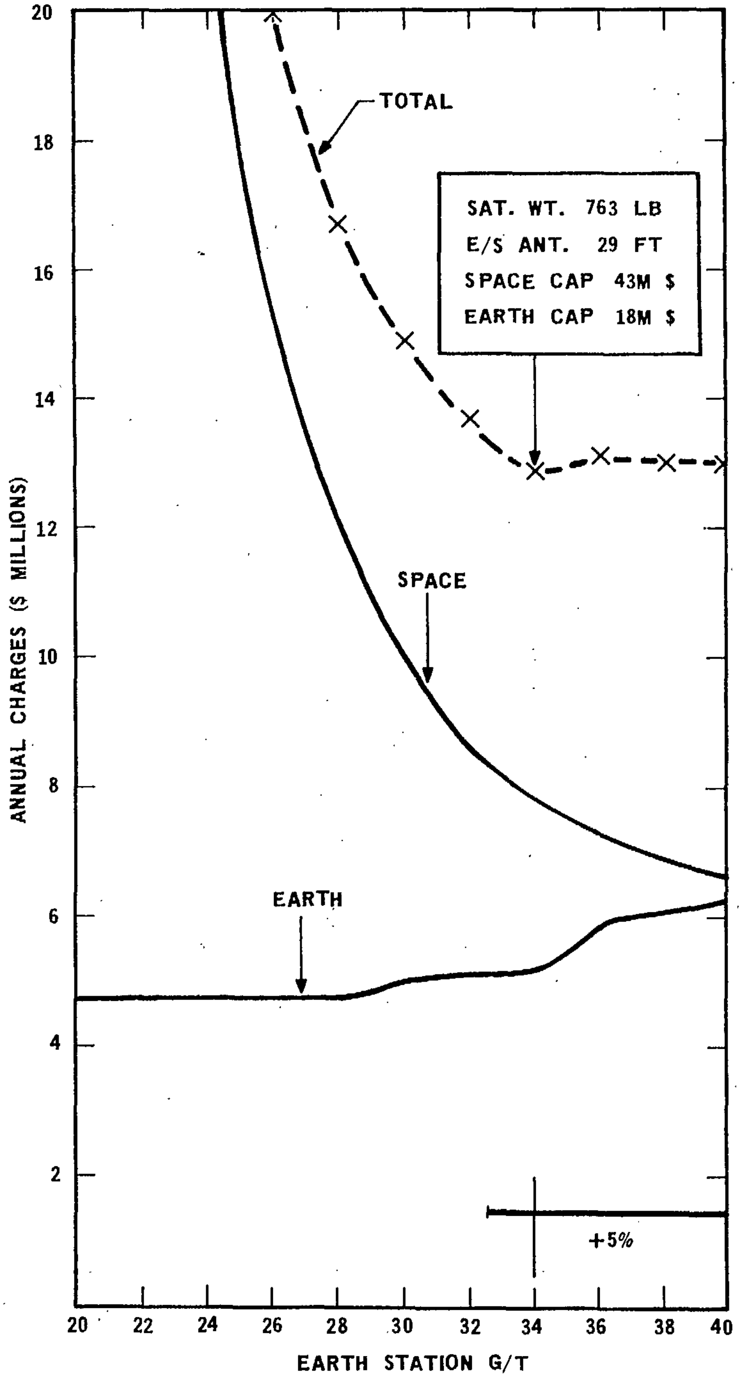

Figure 7-2 Configuration U_2, SDMA/SS-TDMA

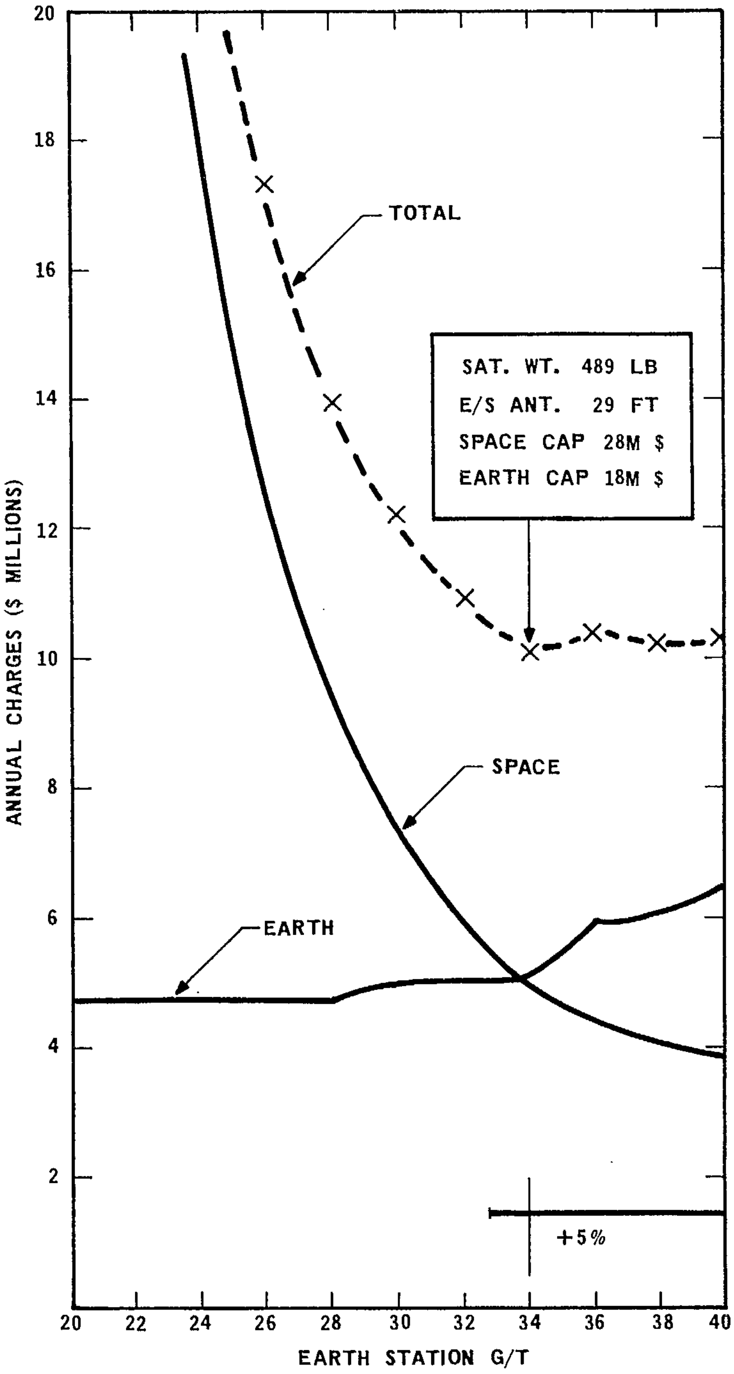

Figure 7-3 Configuration U_3, SDMA/SS-TDMA

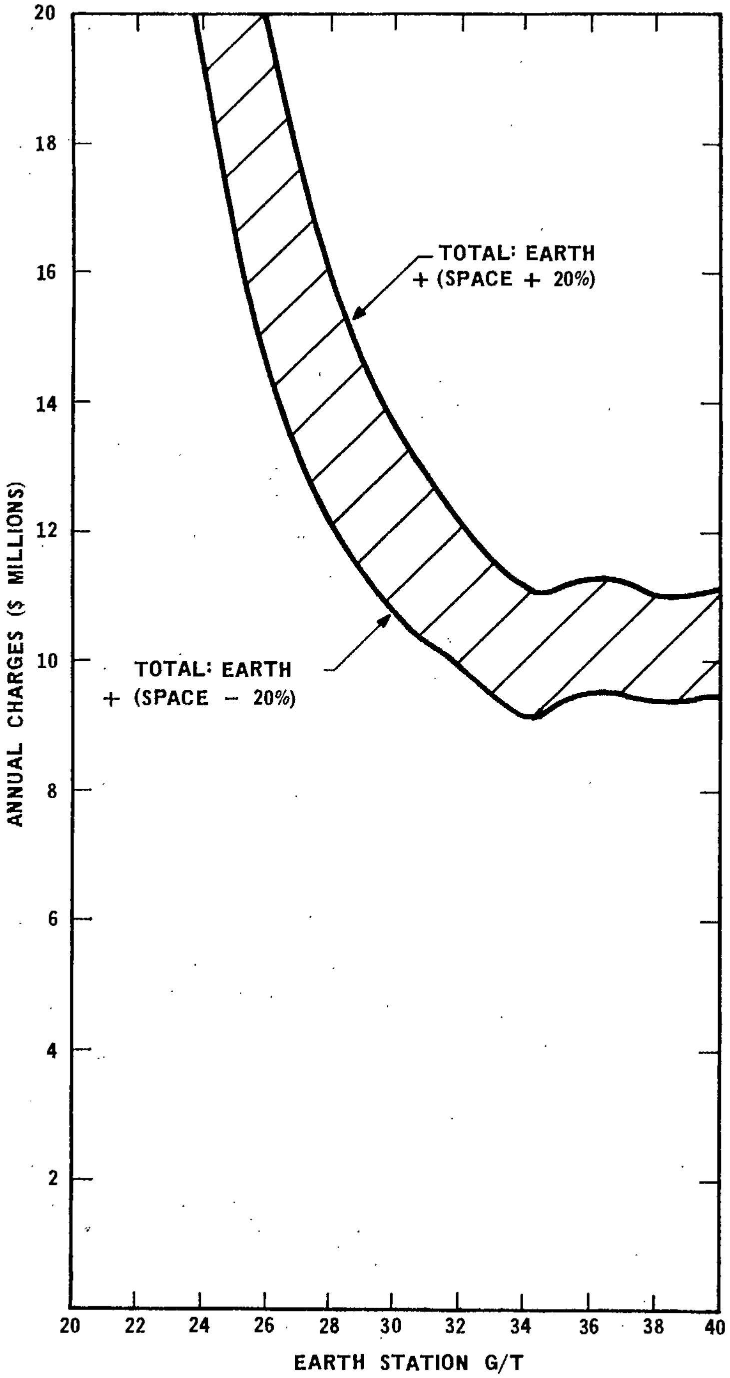

Figure 7-4 Configuration U_3 - Sensitivity to Changes in Space Segment Costs

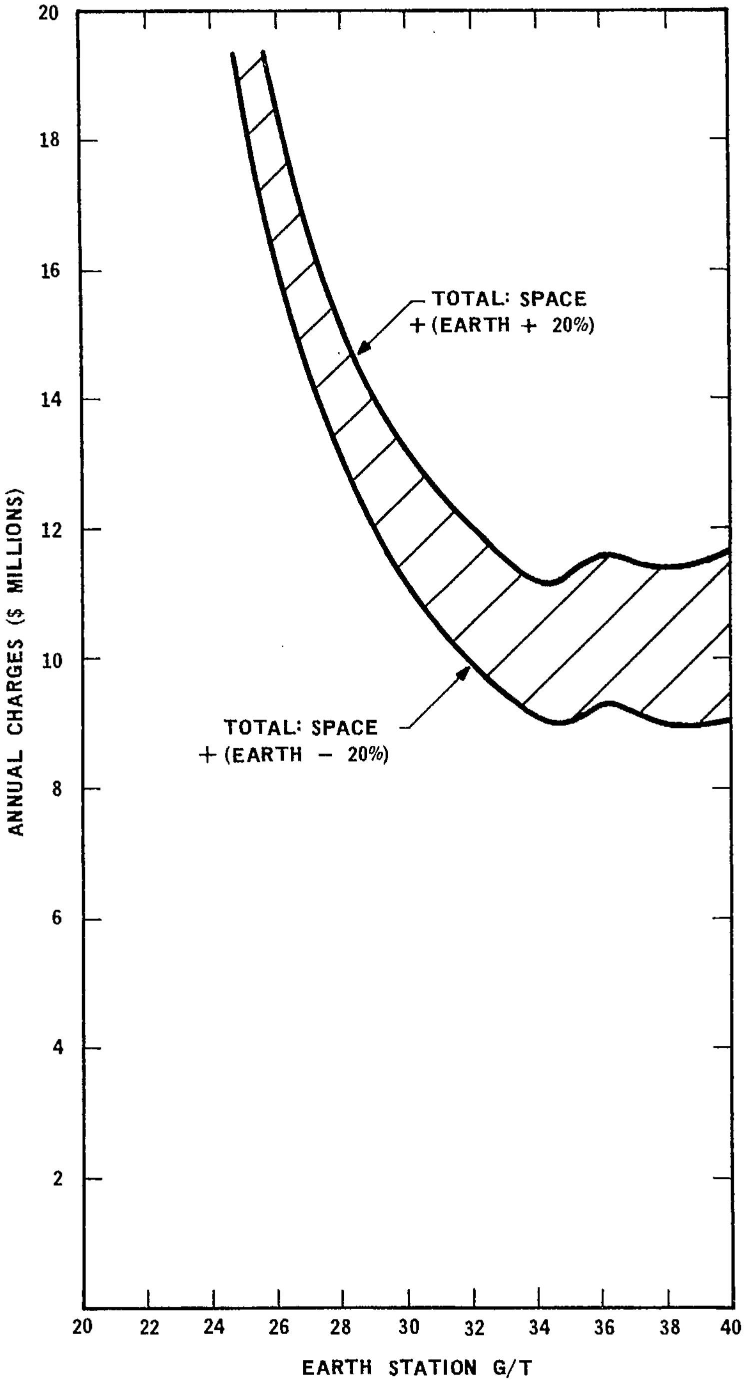

Figure 7-5 Configuration U_3 - Sensitivity to Changes in Earth Segment Costs

7.4.2 MAJOR TRAFFIC: NON-SWITCHED MODELS

The trade-off curves for unswitched system models using digital, FDM-FM and SCPC modulation are given.

All nine system models using fixed interconnections within the spacecraft have been normalized to provide 24 000 voice channel capacity in the spacecraft, with equipment for a total of 10 000 channels being provided on the ground. The optimization curves are given in the following figures:

Figure	
7-6	Digital system, beam plan A, configuration W1
7-7	Digital system, beam plan B, configuration W2
7-8	Digital system, beam plan C, configuration W3
7-9	Configuration W3, sensitivity to changes in space segment costs
7-10	Configuration W3, sensitivity to changes in earth segment costs
7-11	FDM-FM system, beam plan A, configuration X1*
7-12	FDM-FM system, beam plan B, configuration X2*
7-13	FDM-FM system, beam plan C, configuration X3*
7-14	Configuration X3*; sensitivity to changes in space segment costs
7-15	Configuration X3*; sensitivity to changes in earth segment costs
7-16	SCPC system, beam plan A, configuration Y1
7-17	SCPC system, beam plan A, configuration Y2
7-18	SCPC system, beam plan C, configuration Y3
7-19	Configuration Y3; sensitivity to changes in space segment costs
7-20	Configuration Y3; sensitivity to changes in earth segment costs

At the bottom of each trade-off curve an indication of the range of earth station (G/T) values allowable without an increase of Total Annual Charges exceeding 5 percent above the minimum is again given.

The asterisks against the X configurations (FDM-FM) indicate that the system models use speech compression of the Lincompex type. The simple FDM-FM models have been rejected as they cannot fulfil the traffic requirements.

In all cases except configurations W2 and W3 the optimum falls in the range 20 to 40 dB/°K. In the case of W2 an earth station (G/T) above 40 dB/°K is indicated, and this fact renders the W2 configuration hardly feasible on account of the constraint on the largest antenna diameter practicable (as explained in the previous study report[14]). The W3 curve also indicates an optimum at the high end, but in this case the curve of the Total Annual Charges is virtually flat between (G/T) of 33 dB/°K and 40 dB/°K and practical earth station designs are realizable in this range.

The sensitivity curves indicate how the Total Annual Charges change with variations in space and earth segment costs respectively. It should be noted that such changes affect the totals only; the optimum remains at the same point.

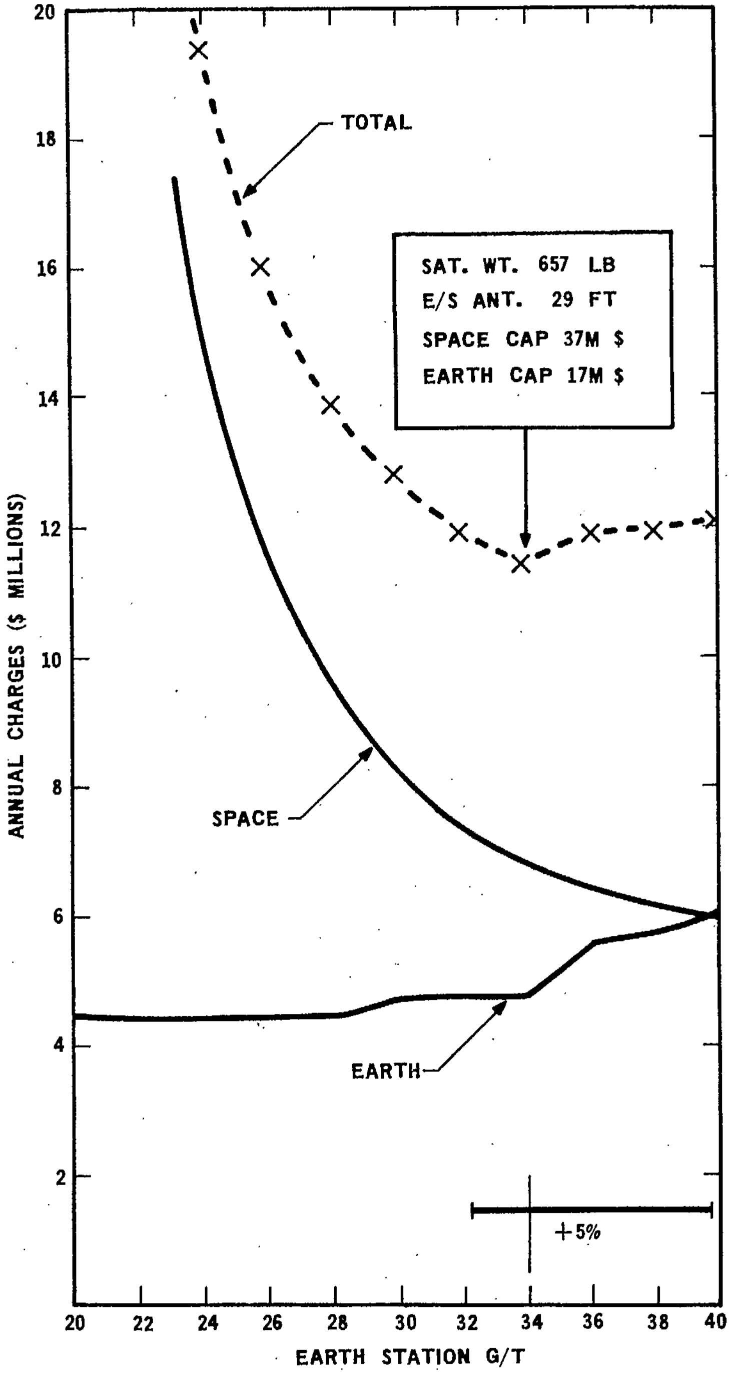

Figure 7-6 Configuration W1, PCM-PSK-TDMA

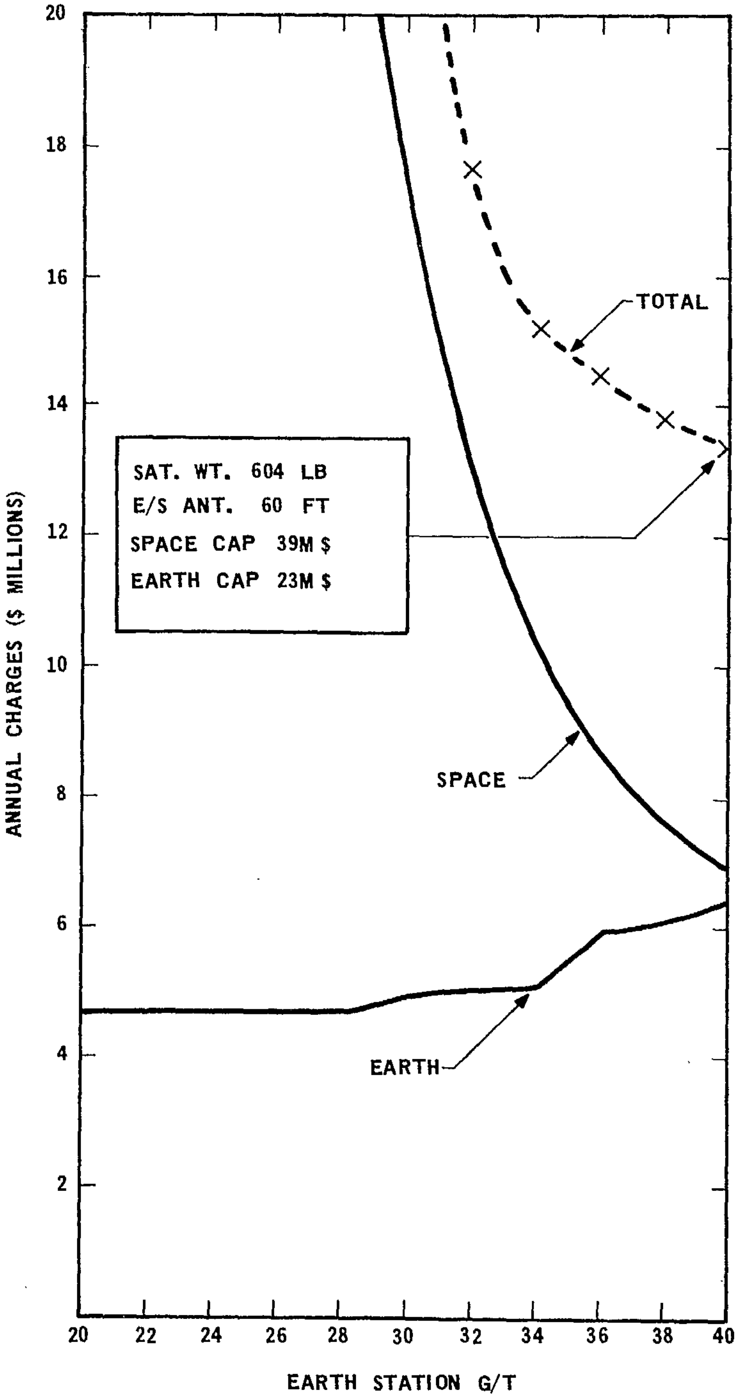

Figure 7-7 Configuration W2, PCM-PSK-TDMA

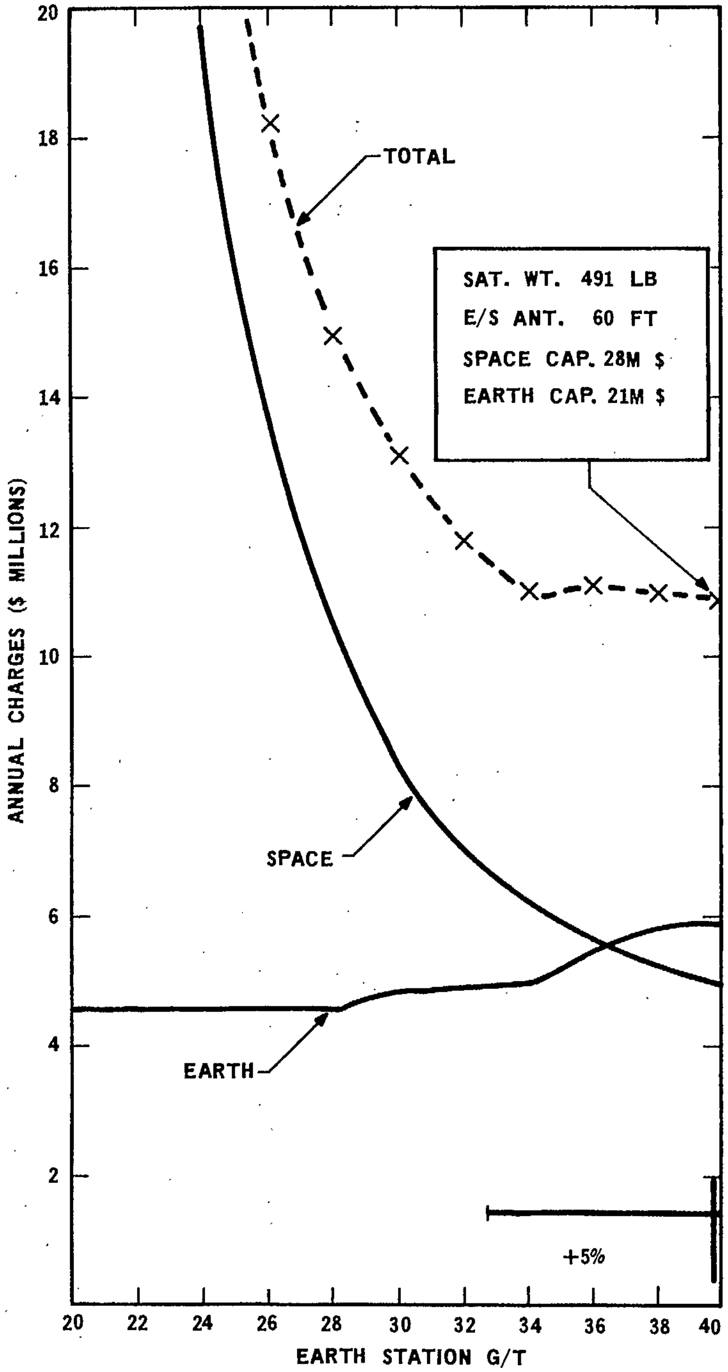

Figure 7-8 Configuration W3, PCM-PSK-TDMA

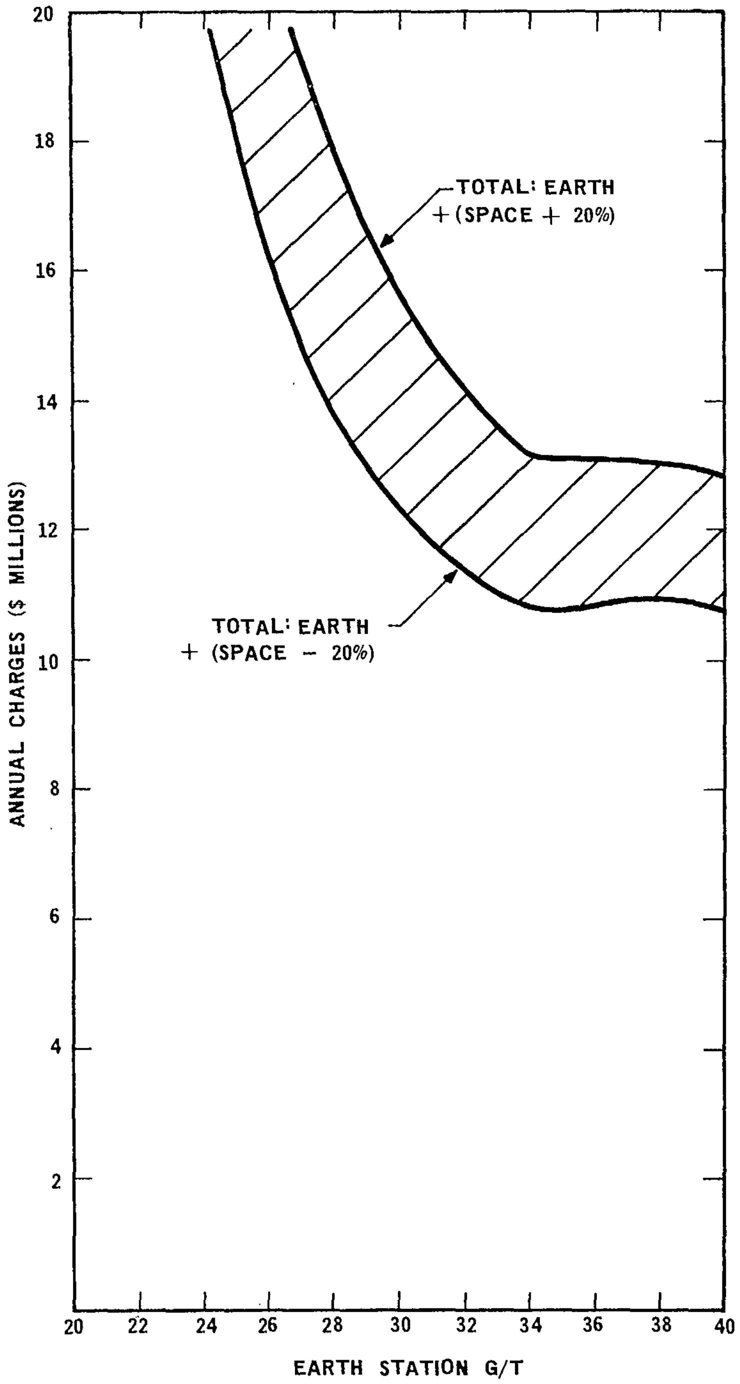

Figure 7-9 Configuration W3 - Sensitivity to Changes in Space Segment Costs

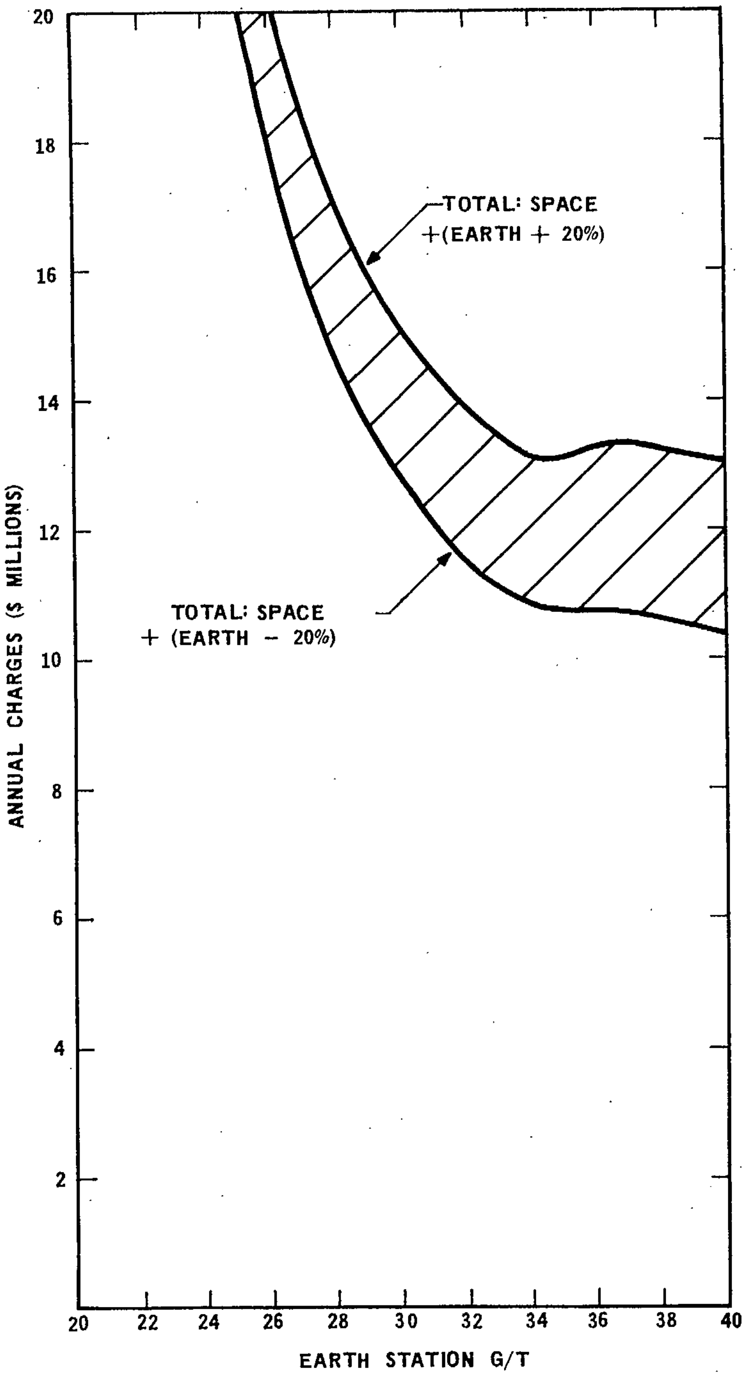

Figure 7-10 Configuration W3 - Sensitivity to Changes in Earth Segment Costs

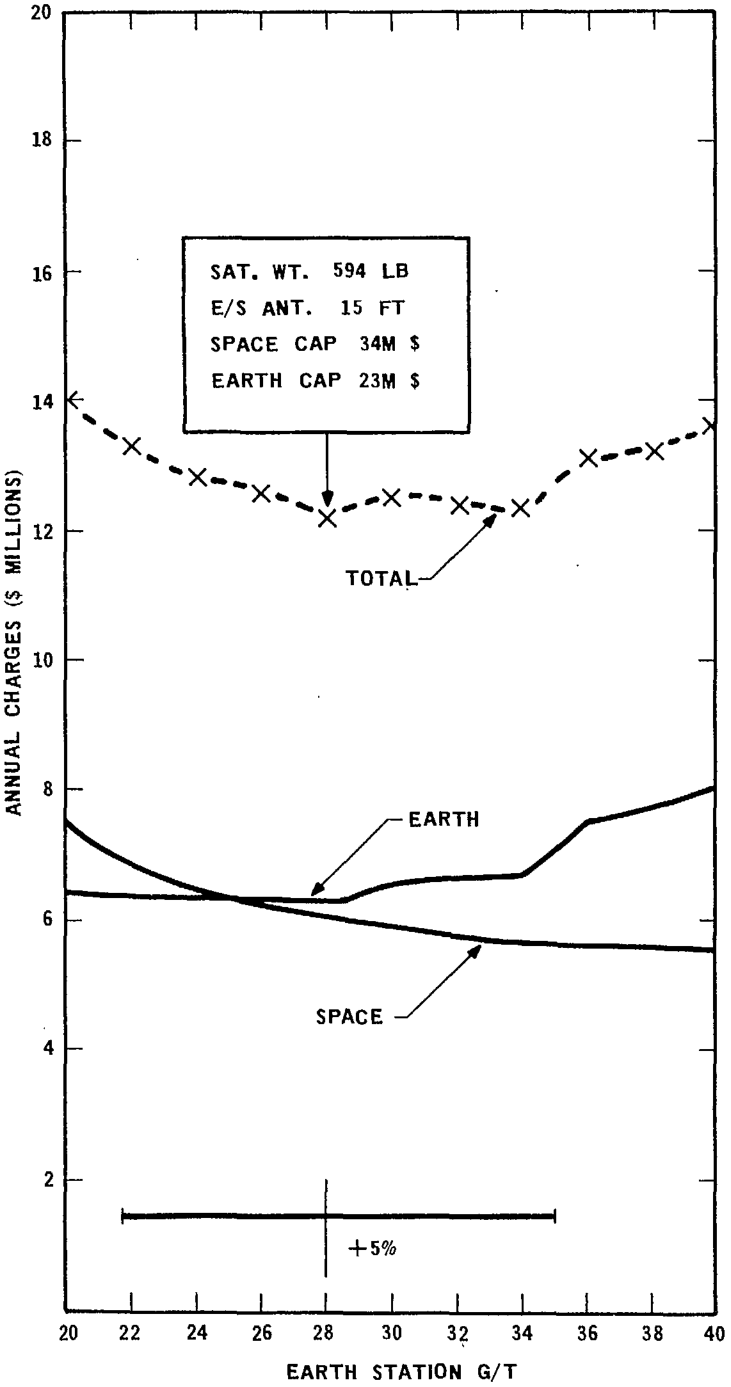

Figure 7-11 Configuration X1*, FDM-FM-FDMA

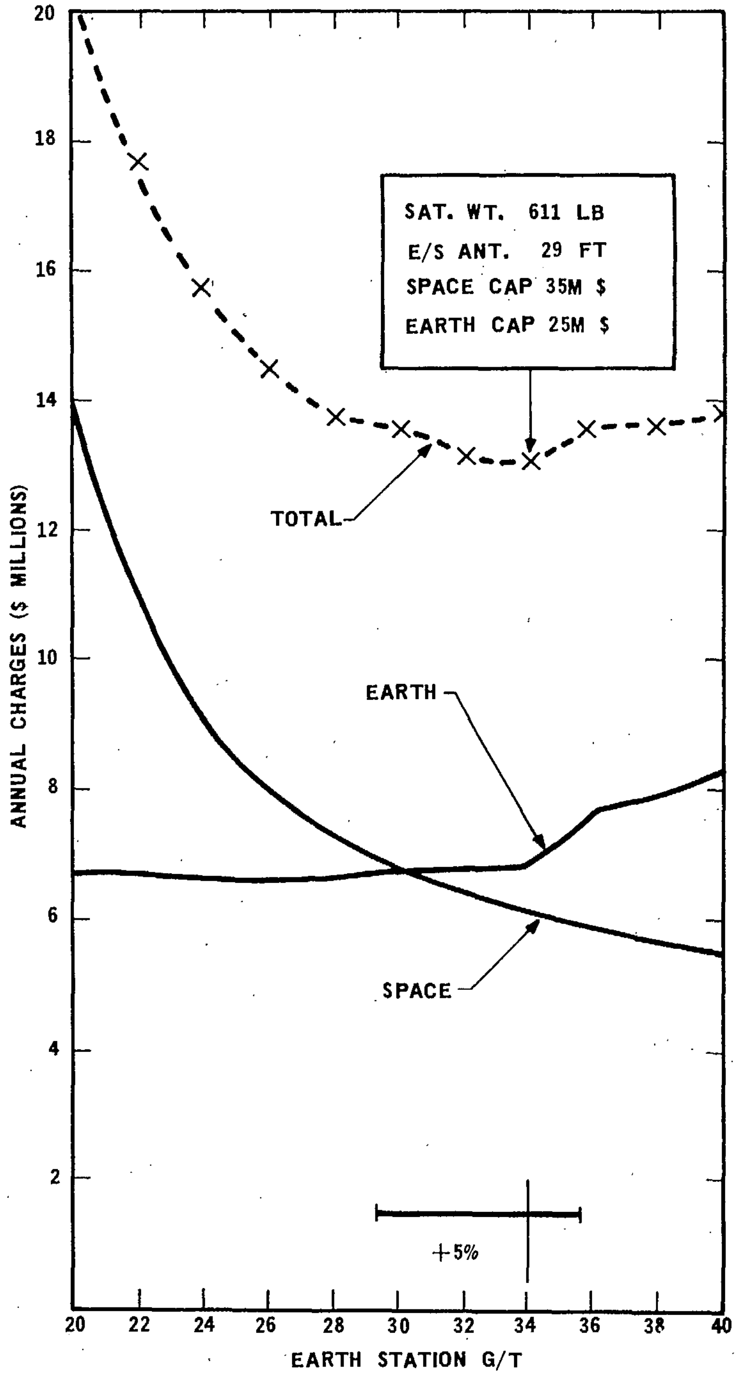

Figure 7-12 Configuration X2*, FDM-FM-FDMA

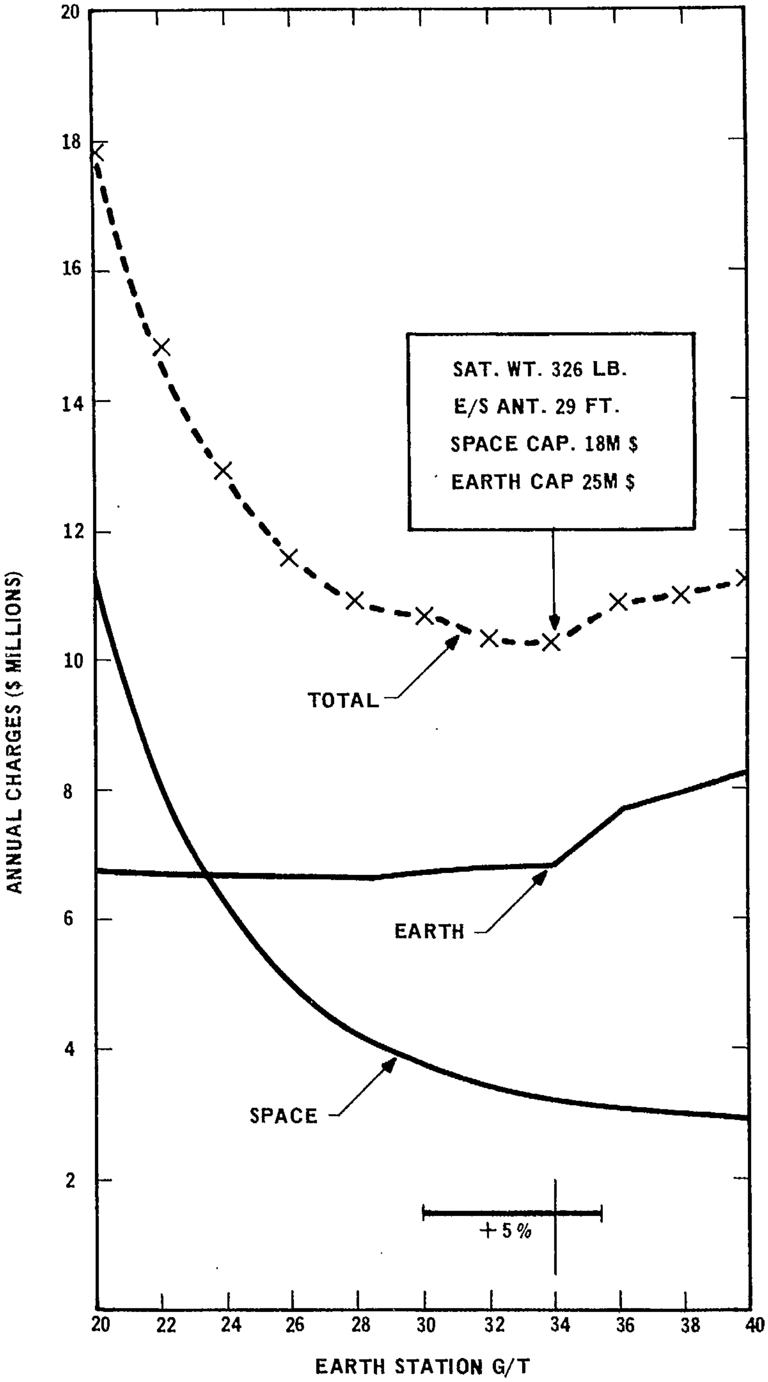

Figure 7-13 Configuration X3*, FDM-FM-FDMA

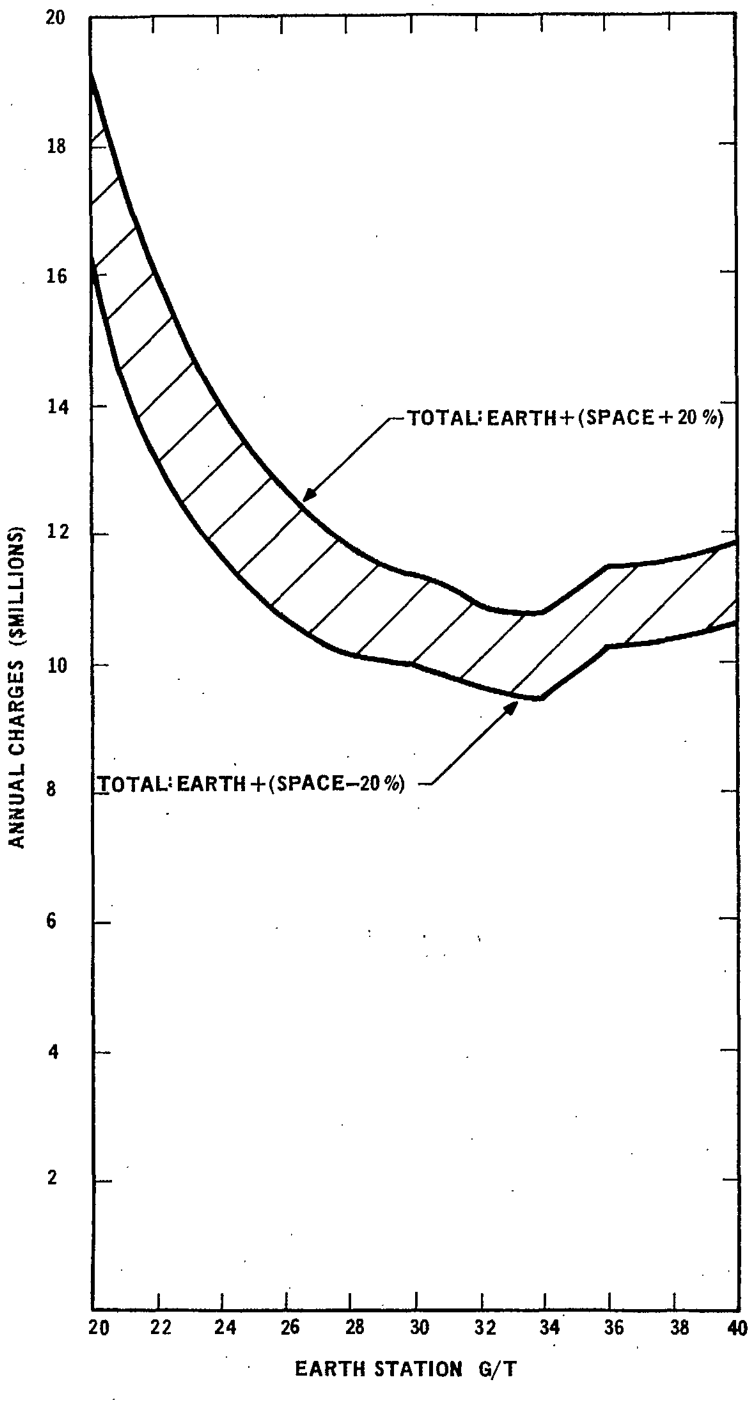

Figure 7-14 Configuration X3* - Sensitivity to Changes in Space Segment Costs

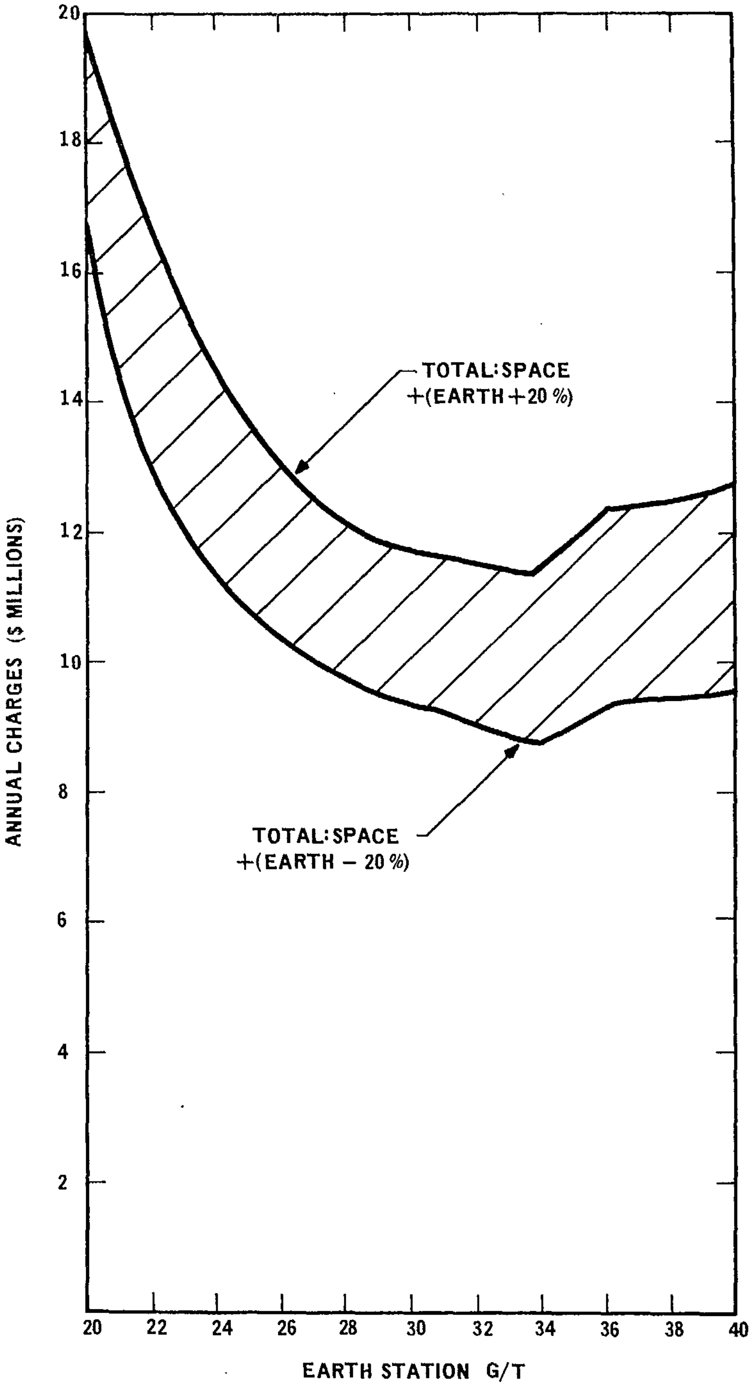

Figure 7-15 Configuration X3* - Sensitivity to Changes in Earth Segment Costs

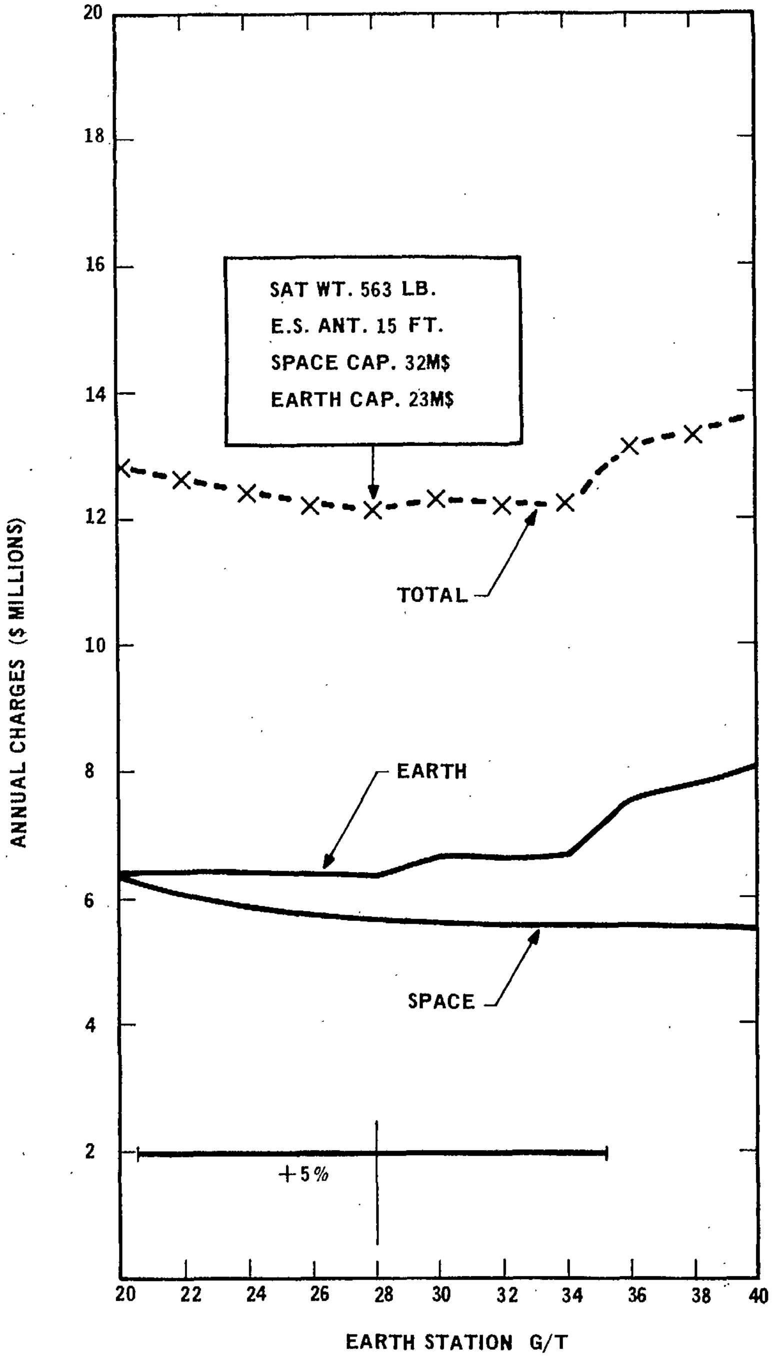

Figure 7-16 Configuration Y1, SCPC-FM

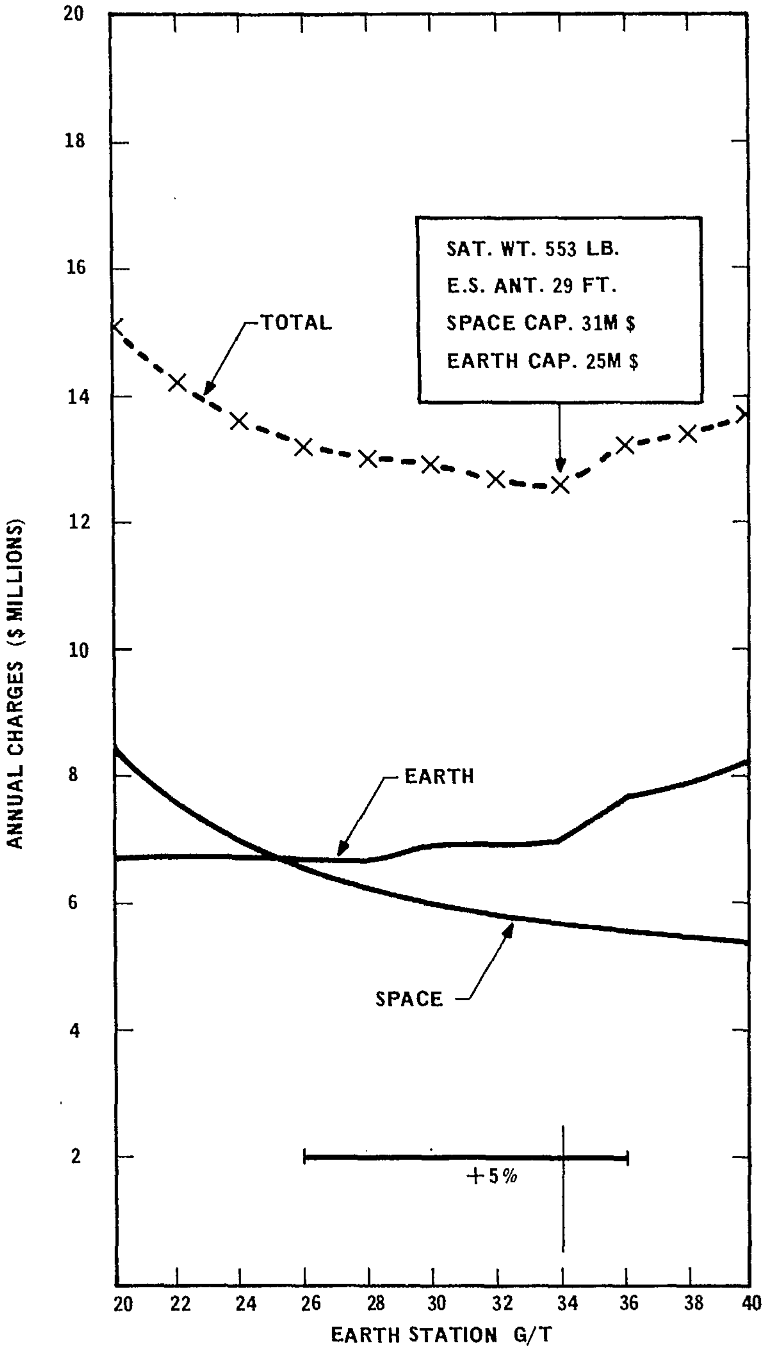

Figure 7-17 Configuration Y2, SCPC-FM

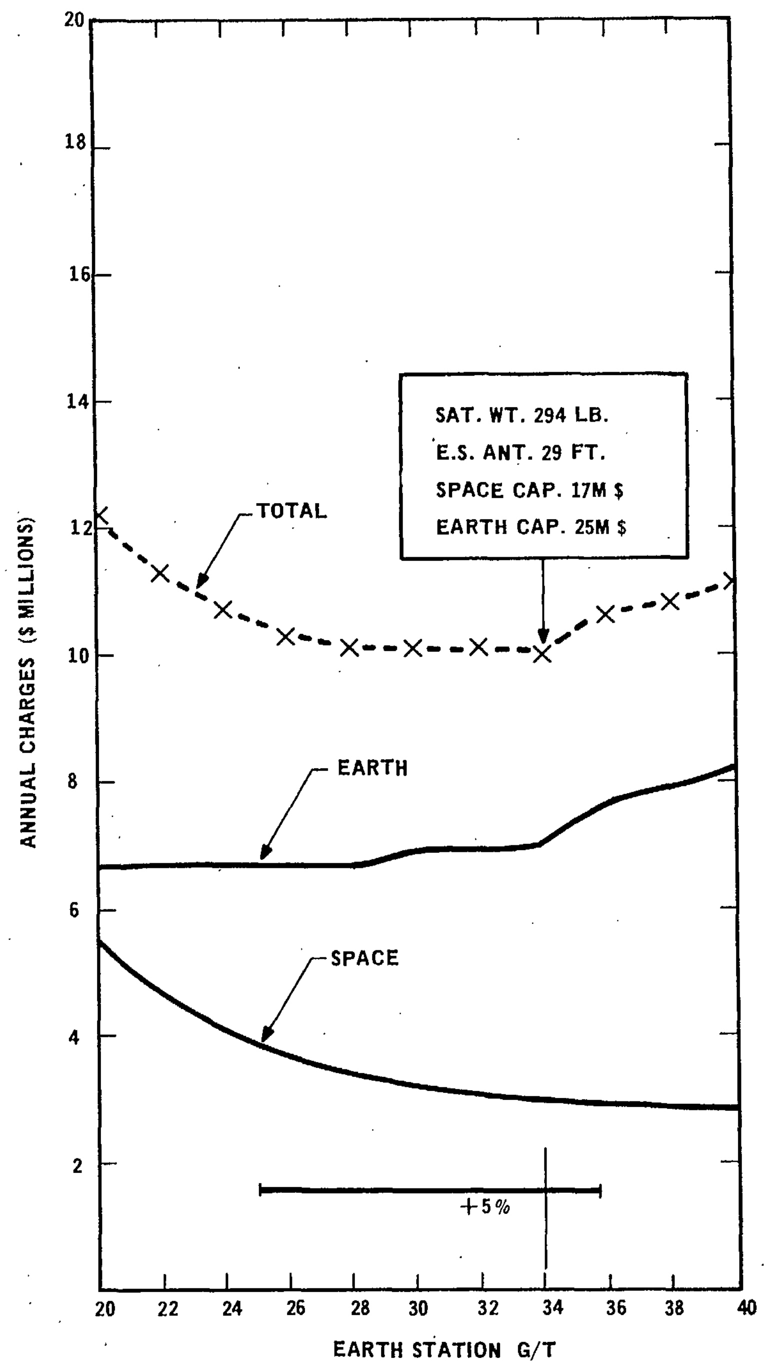

Figure 7-18 Configuration Y3, SCPC-FM

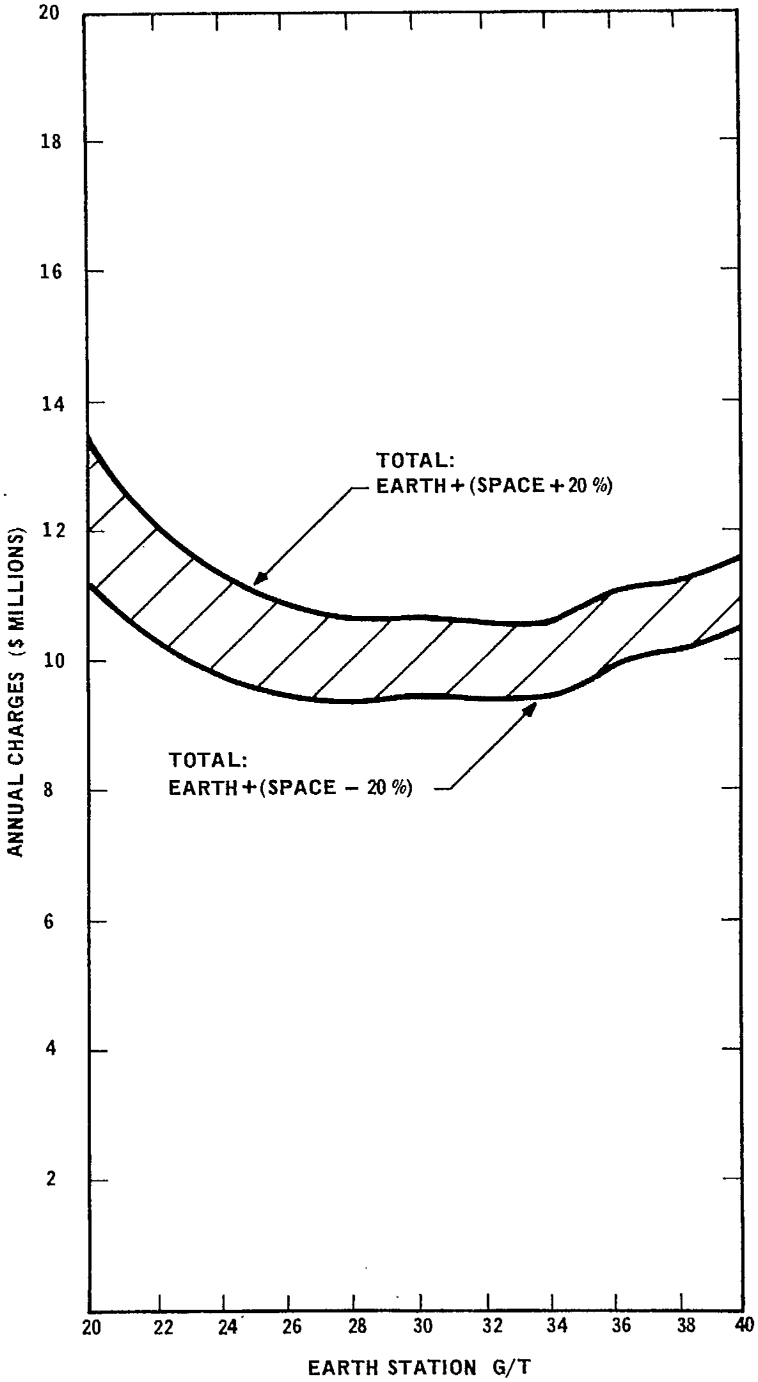

Figure 7-19 Configuration Y3 - Sensitivity to Changes in Space Segment Costs

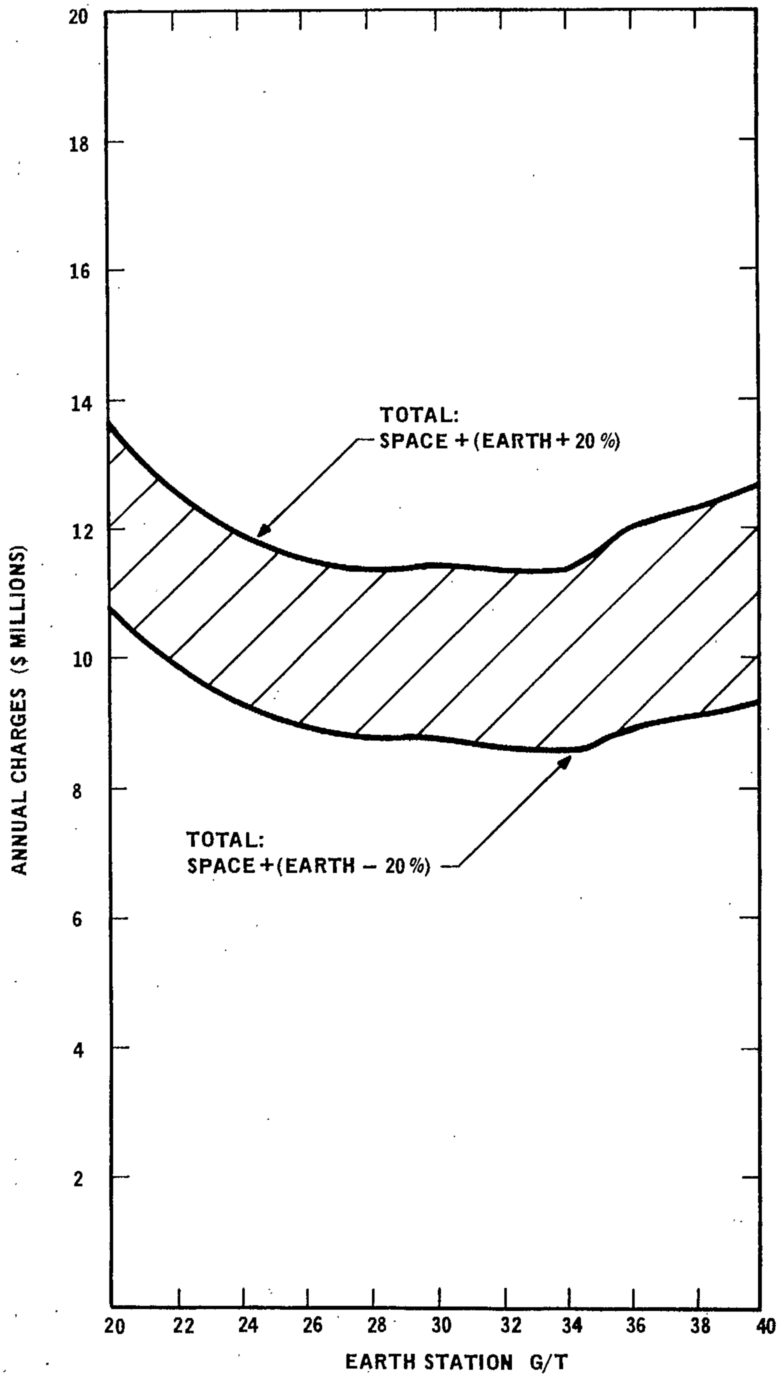

Figure 7-20 Configuration Y3 - Sensitivity to Changes in Earth Segment Costs

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7.4.3 REMOTE COMMUNICATIONS

The trade-off curves for system models providing single-channel-per-carrier service to remote stations (without satellite on-board switching) are given.

The trade-off curves for the remote communications system models Z1/Z3 and Z2 are given in Figures 7-21 and 7-22, and their sensitivity to space and earth segment cost variations are indicated in Figures 7-23 and 7-24 respectively. It should be noted that only two variants exist in this case, since both major route Beam Plans (A) and (C) use the same five-beam configuration for all Canada coverage. In the case of remote communications the total traffic capacity assumed is 1200 one-way voice channels (3 at each of 400 earth stations). Since the large number of earth station antenna cost, the trade-off curves show minima at lower (G/T) values than major route systems, and have therefore been plotted over the range 16 to 36 dB/°K.

The sensitivity curves indicate how the Total Annual Charges change with variations in space and earth segment costs respectively. It should be noted that such changes affect the totals only; the optimum remains at the same point.

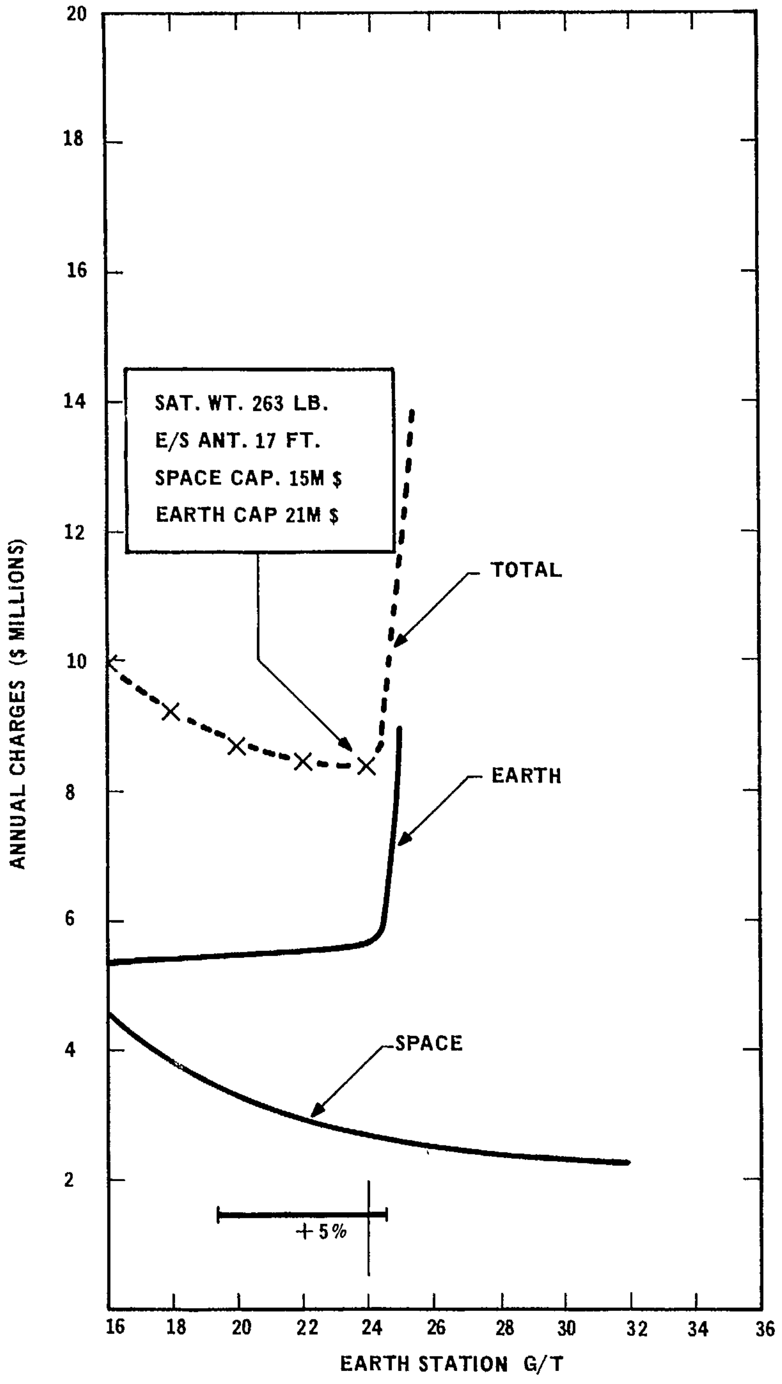

Figure 7-21 Configuration Z1, Z3, SCPC-FM

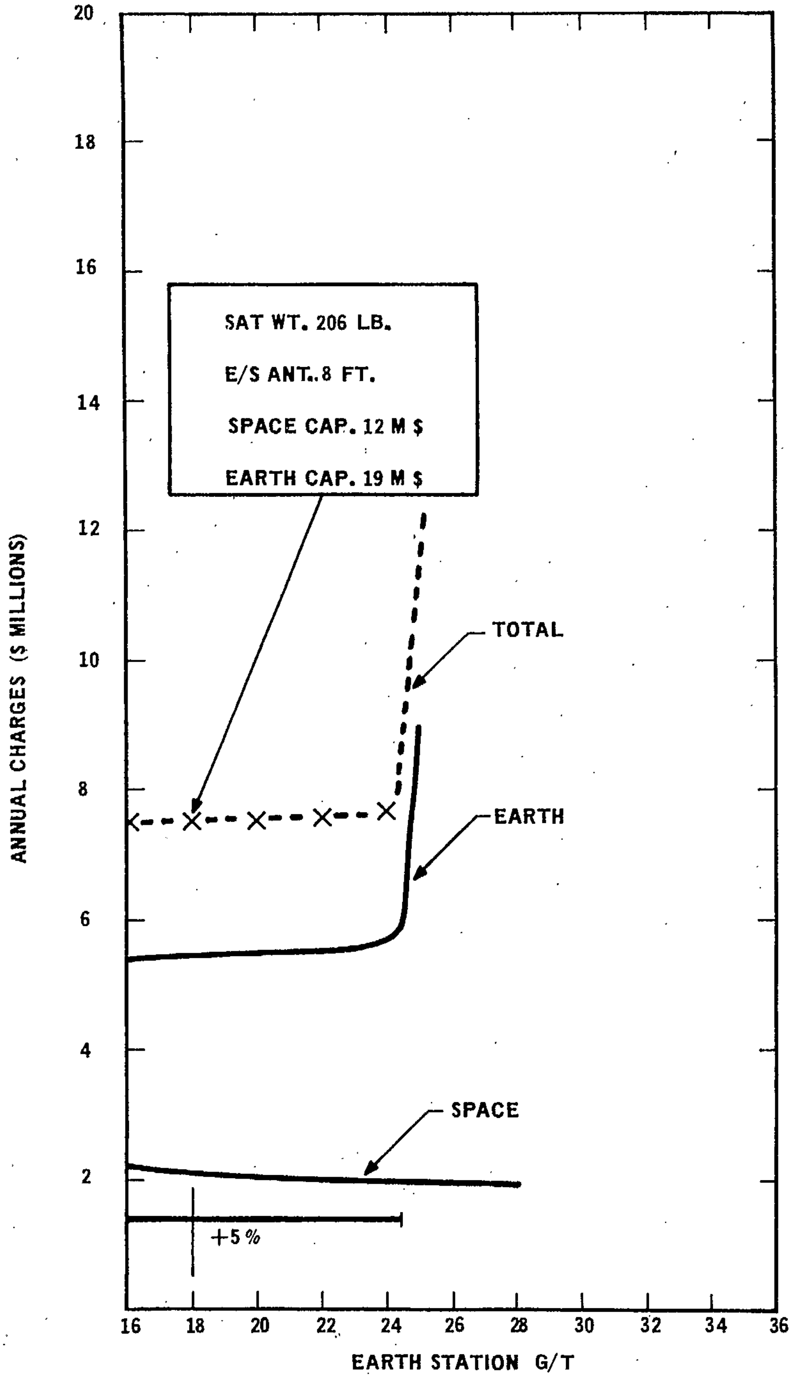

Figure 7-22 Configuration Z2, SCPC-FM

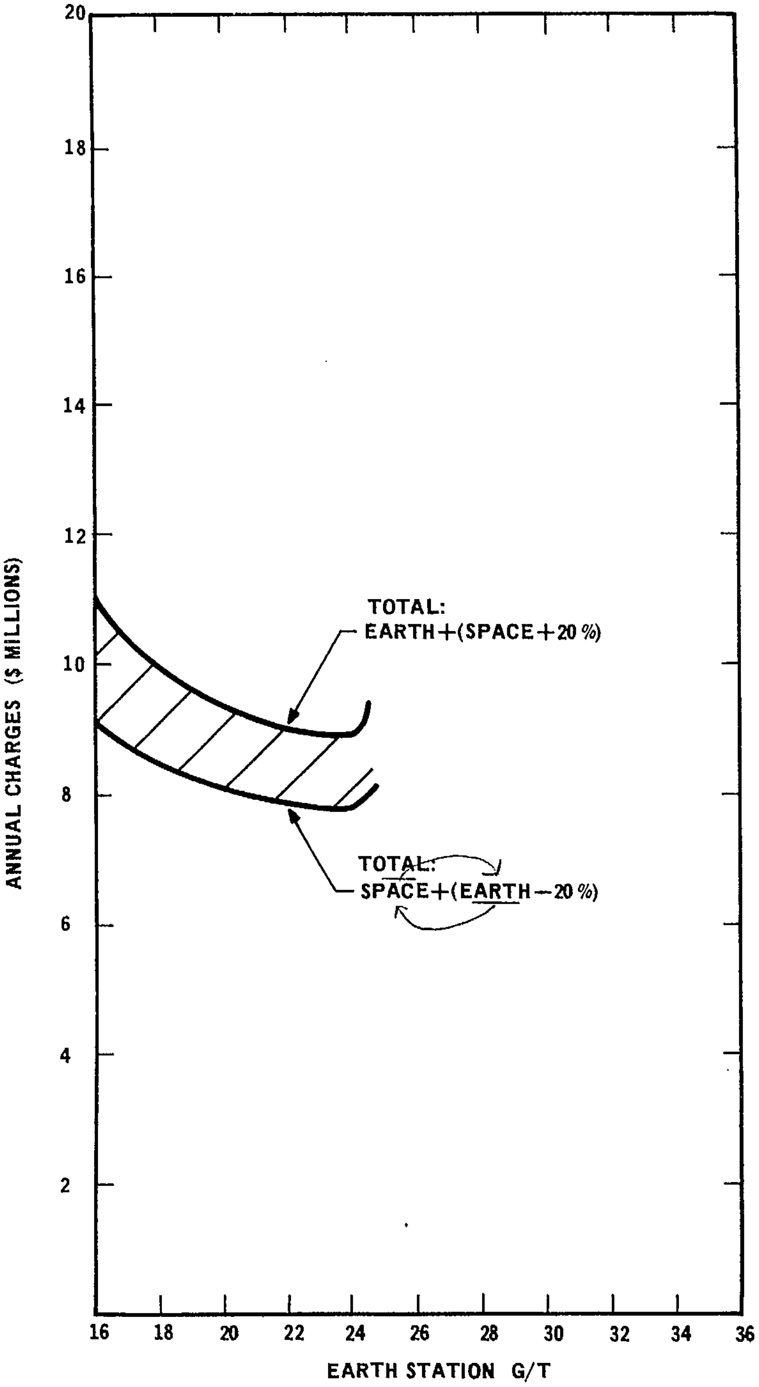

Figure 7-23 Configuration Z1, Z3 - Sensitivity to Changes in Space Segment Costs

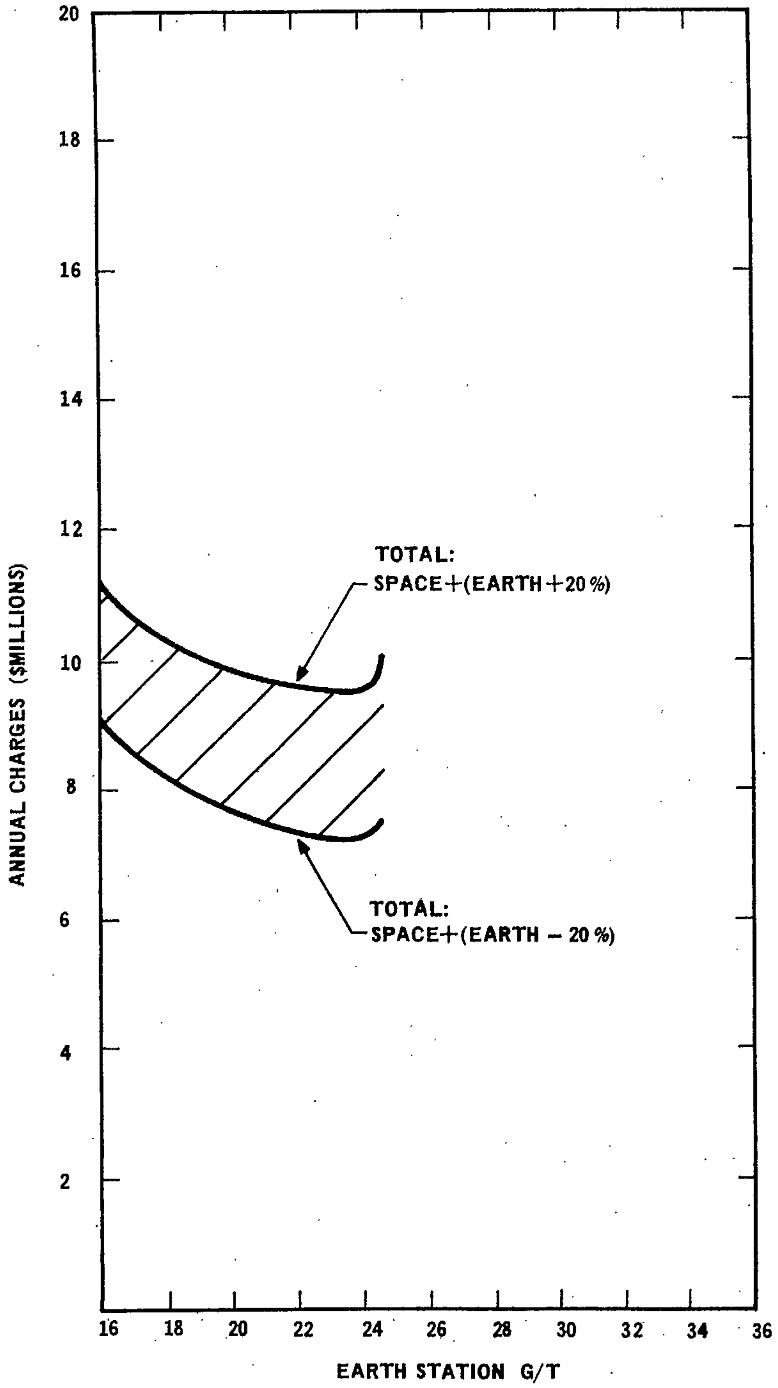

Figure 7-24 Configuration Z1, Z3 - Sensitivity to Changes in Earth Segment Costs

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8. COMPARISON AND CHOICE OF MODELS

Comparisons of relative costs and technical and operational merits of the various models are presented. Major route traffic and remote communications are dealt with. A choice is made of the optimum model.

The main objective of this study is to investigate the costs and broad system concepts of a 12 and 14 GHz multiple spot beam satellite network for Canada. The traffic needs for the period concerned, the technological constraint and the approach adopted in carrying out meaningful system trade-offs have all been dealt with in the previous chapters. In particular, in the matter of technical and cost trade-off, it has been found that in a satellite system with multispot beams and frequency reuse, it is not possible to arbitrarily change one parameter without simultaneously affecting many others. For example, changing the size of the spot beam affects the coverage area. This necessitates increasing or decreasing the number of beams to cater to the traffic needs. This in turn affects the complexity of the equipment on board the satellite, the interconnection between beams, the interference isolation between beams, the bandwidth and frequency allocation of the rf channels, the need or otherwise of cross-polarization frequency reuse, etc. Hence it is not easy to see the effect of any one single change in system parameter. To render the problem more tractable, it has been found necessary to generate certain basic models that satisfy all the traffic, technological and technical constraints and which provide plausible solutions to Canada's needs. This has been the approach taken.

The terms of reference of this study require evaluation of two models for the major traffic network, namely, the on-board switched and the transparent non-switched models. The remote communications service is also to be investigated.

Taking the approach explained above, three antenna beam plans are postulated as being representative of the most likely application of spot beam satellite to Canada. These are Beam Plans (A), (B) and (C) as described in chapter 3. For the switched model, only one modulation technique is considered practicable and that is digital modulation with time division multiple access. For the transparent model, three modulation schemes are considered. These are digital, multichannel fm and single channel-per-carrier fm. The various configurations are designated as shown in Table 8-1.

Table 8-1 Models Investigated

		MODULATION	BEAM PLAN A	BEAM PLAN B	BEAM PLAN C
MAJOR ROUTE SERVICE	SWITCHED SYSTEM	SDMA/SS-TDMA	U1	U2	U3
	NON-SWITCHED SYSTEM	PCM-PSK	W1	W2	W3
		FDM-FM-FDMA (LINCOMPEX)	X1 *	X2 *	X3 *
		SCPC-FM (LINCOMPEX)	Y1	Y2	Y3

REMOTE SERVICE	NON-SWITCHED SYSTEM	SCPC-FM (LINCOMPEX)	Z1/Z3	Z2	Z1/Z3

BEAM PLAN (A)	8 BEAMS	0.5° X 0.5°	MAJOR ROUTE
	5 BEAMS	1.5° X 2.0°	REMOTE
BEAM PLAN (B)	14 BEAMS	0.7° X 1.7°	MAJOR ROUTE REMOTE
BEAM PLAN (C)	4 BEAMS	0.7° X 1.7°	MAJOR ROUTE
	5 BEAMS	1.5° X 2.0°	REMOTE

Table 8-2 Comparison of Optimized Systems

	MAJOR TRAFFIC SYSTEM												REMOTE SYSTEM		
	SWITCHED			NON-SWITCHED									REMOTE		
	SDMA/SS-TDMA			DIGITAL			FDM-FM*			SC/C			SC/C		
BEAM PLAN	A	B	C	A	B	C	A	B	C	A	B	C	A	B	C
	U1	U2	U3	W1	W2	W3	X1*	X2*	X3*	Y1	Y2	Y3	Z1	Z2	Z3
TOTAL ANNUAL CHARGES	10.8	12.8	10.0	11.4	13.4	10.9	12.3	13.1	10.2	12.1	12.6	10.0	8.4	7.5	8.4
Annual charges (space)	6.2	7.8	5.0	6.6	7.0	5.0	6.0	6.2	3.3	5.7	5.7	3.0	2.7	2.1	2.7
Annual charges (earth)	4.6	5.0	5.0	4.8	6.4	5.9	6.3	6.9	6.9	6.4	6.9	7.0	5.7	5.4	5.7
CAPITAL (total)	50	61	46	54	62	49	57	60	43	55	56	42	35	31	35
CAPITAL (space segment)	34	43	28	37	39	28	34	35	18	32	31	17	15	12	15
CAPITAL (earth segment)	16	18	18	17	23	21	23	25	25	23	25	25	20	19	20
Satellite weight lb	609	763	489	657	684	491	594	611	326	563	553	294	263	206	263
Satellite total power W	99	344	326	202	250	89	72	98	82	60	83	67	80	27	80
RF power/transponder W	2.0	5.0	9.0	6.0	3.0	1.1	1.7	1.0	2.0	0.6	0.3	0.6	4.0	3.0	4.0
Earth station antenna dia ft	29	29	29	29	59	59	15	29	29	15	30	30	17	8	17
Earth station rf power W	198	463	927	621	730	292	205	121	121	546	321	321	.07	.12	.07
EXTREMES OF 5% RANGE:-															
Least satellite weight lb	600	658	381	503	684	491	555	560	320	545	541	281	263	196	263
Earth station antenna dia ft	37	59	59	47	59	59	37	47	37	34	37	37	17	17	17
Satellite weight lb	645	800	510	709	844	638	635	659	374	609	628	372	317	213	317
Least earth station antenna dia ft	22	26	26	23	37	26	9	18	19	59	12	9	11	7	11
Max system capacity × 1000 (Ch)	24	24	24	24	24	24	24	24	24	24	24	24	-	-	-
Installed (earth) capacity ×1000 (Ch)	10	10	10	10	10	10	10	10	10	10	10	10	1.2	1.2	1.2

* Based on use of lincompex speech compression @ $500/unit.

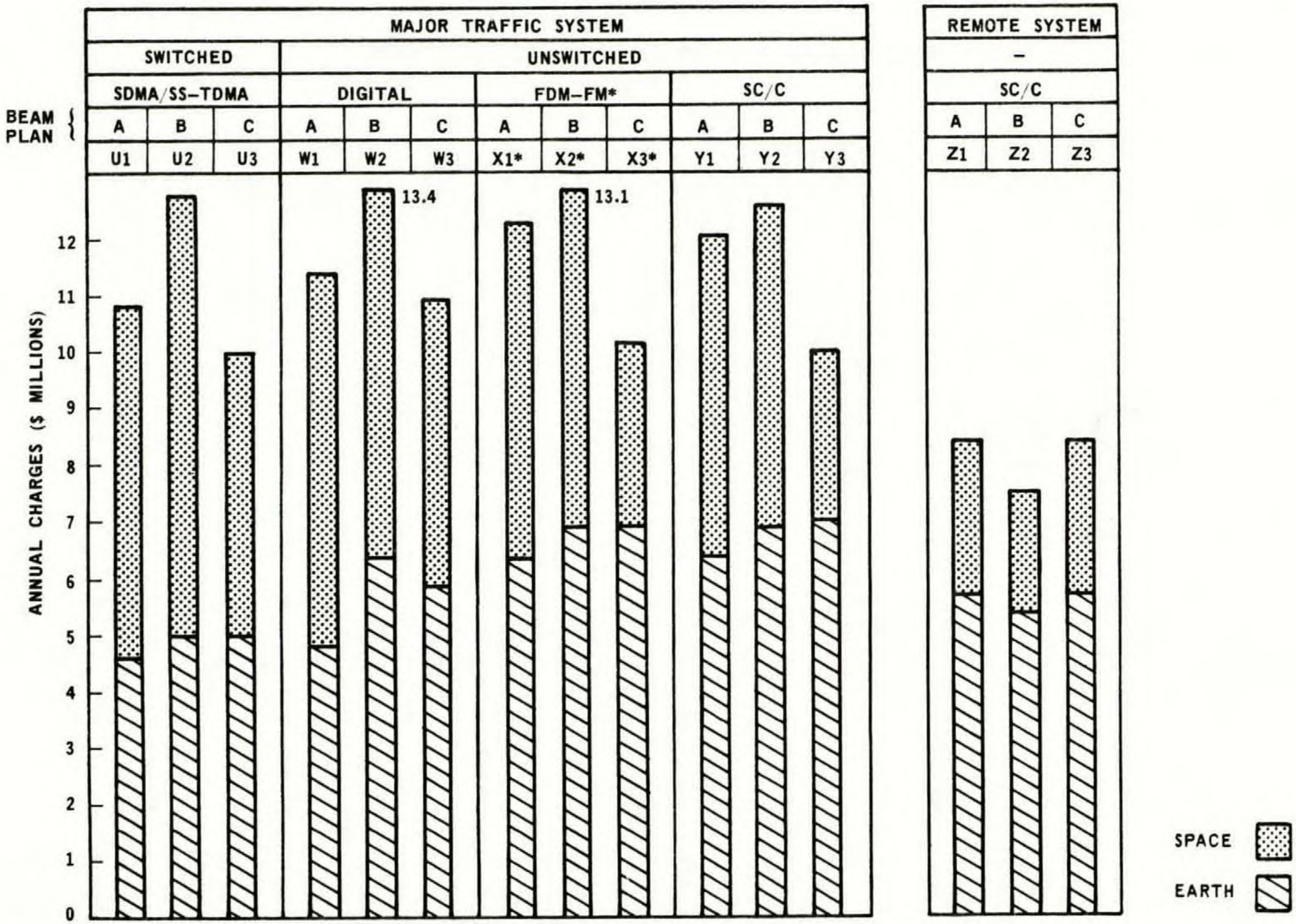

Figure 8-1 Optimized Systems - Annual Charges

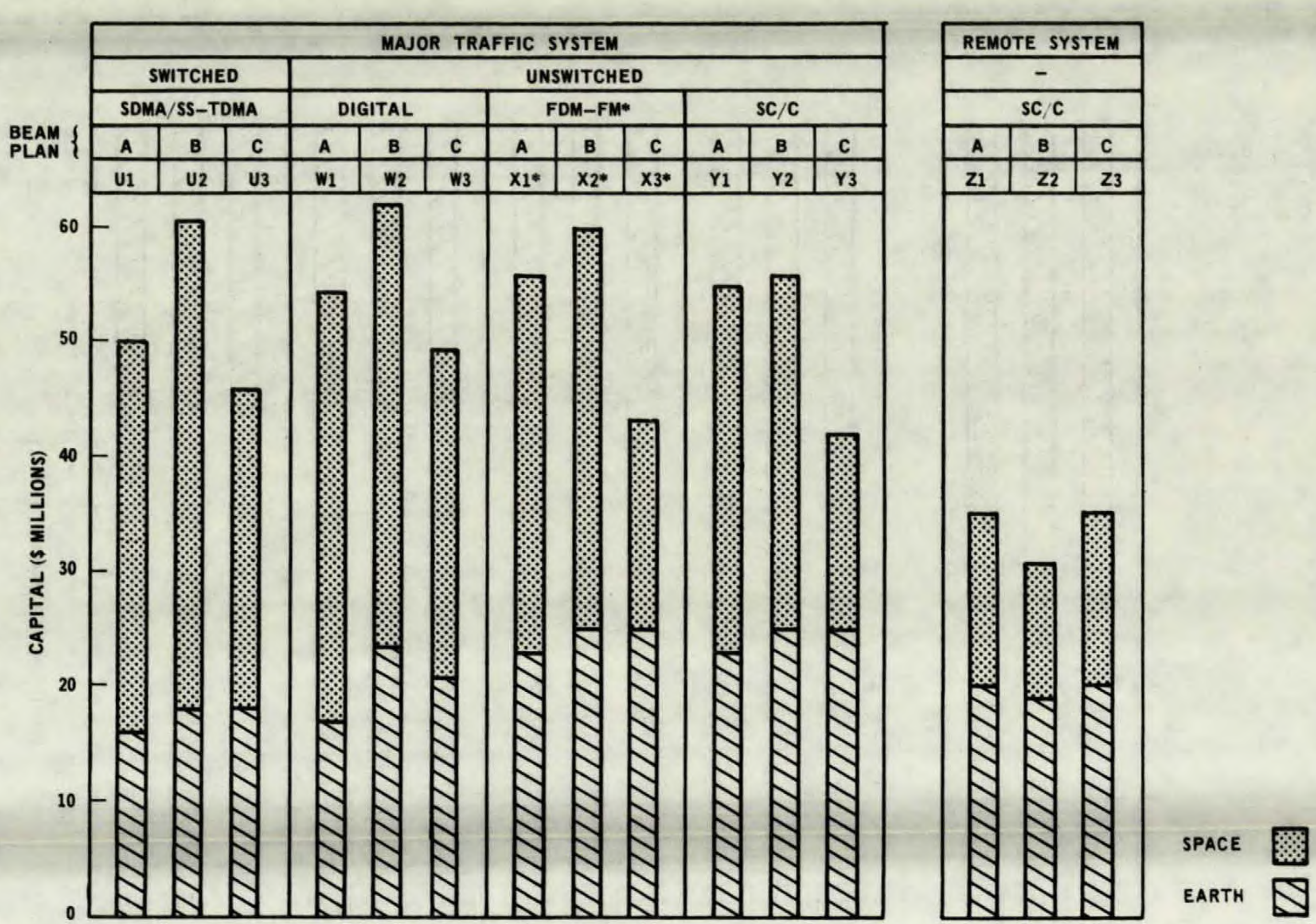

Figure 8-2 Optimized Systems - Capital Costs

In chapter 7, the results of the cost trade-off computations have been presented. The cost and technical comparisons between all these models will now be given below.

Table 8-2 shows the significant cost and technical parameters at the minimum annual charge point for each of the configurations investigated, both major traffic and remote communications. In order to assist the reader to better visualize the relative cost elements, the annual charges and capital costs are presented in Figures 8-1 and 8-2 respectively, in the form of column charts.

From the above and from the trade-off results presented in chapter 7, the following comments may be made:

Major Traffic System

1. In order to make valid comparisons, the system capacities have been normalized to 24 000 channels (one-way) for the space segment and 10 000 equipped channels for the earth segment. The earth segment can of course expand to the full potential of 24 000 channels, but the 1985 traffic forecast is only 10 000 channels and this latter figure is used in the cost computation. The need for 24 000 channels for the space segment is dictated by the nature of the spot beam coverage and the necessity to serve the densest traffic zones such as Toronto and Montreal which require a capacity each of about 3000 channels. Multiplexing equipment cost is included in the system evaluation. It will be noted that the conventional (non-speech-compressed) FDM-FM-FDMA system is excluded in the computed results. This is because of its inefficiency in bandwidth utilization and consequently its inability to meet the capacity requirements within the available spectrum. Therefore only the speech compressed FDM-FM-FDMA system is presented.

As explained above, the 24 000 channel capability of the space segment cannot be fully utilized with the predicted traffic pattern. Any changes that may occur in this pattern will result in more, or less, utilization. However, this will have only a second order effect on the cost ratios and will not invalidate the following comments and recommendations.

2. Irrespective of whether switched or non-switched or whatever the modulation scheme used, the Beam Plan (C) models are the lowest in total capital and annual costs. As can be seen in Table 8-2, this is due largely to the reduction in space segment costs. As will be indicated later, the space segment cost is mainly controlled by equipment deadweight (that is, weight which is a function of equipment configuration and is not dependent on satellite power) and therefore, the simplicity of the four-beam

design of Plan (C) has a clear advantage. A two beam configuration has not been considered on account of its inability to provide the degree of frequency reuse to give the required systems capacity. Hence Beam Plan (C) with four medium size spot beams appears the best choice.

Among the Beam Plan (C) models, there is little variance in annual charges between the switched and non-switched configurations and between the various modulation schemes. The Total Annual Charges are all clustered around the $10 million figure. This is surprising considering the vastly different modulation techniques and systems parameters. From the cost point of view, therefore, there is little to choose between the switched and non-switched approach.

3. In view of the remarks of Item 2, the choice between switched and non-switched models must rest with the technical and operational merits. The following comments apply:

- The switched model is more flexible in catering to changing traffic demands than the non-switched models, on account of its ability to change traffic routings from beam to beam almost instantaneously. This is not to say that the non-switched model has little flexibility. The question is one of degree. A fair amount of system flexibility is inherent in the digital TDMA system and a substantial degree of flexibility is also available from the single-channel-per-carrier system.

- The switched system necessitates a commitment to the modulation technique and mode of operation before the satellite is designed and launched. The non-switched system on the other hand imposes no such constraint. Indeed, with the non-switched system, by virtue of the satellite transparency to up-link and down-link signals, the modulation can be changed from FDM-FM-FDMA to PCM-PSK-TDMA or to single-channel-per-carrier with either fm or digital modulation or to some other new modulation technique at any time during the life of the satellite. This system flexibility is of great importance.

- As of the date of this report, the on-board switching controlled by a mini-computer has not yet been space tested. There is therefore a certain degree of uncertainty regarding system reliability.

From the above considerations, on balance it appears that for the Canadian application in the time frame considered, the non-switched model is to be recommended.

The situation should, however, be reviewed periodically to ensure that the conclusions have not been invalidated by technological and/or network developments.

4. Comparing digital versus analog modulation, it can be seen that in general the space segment annual charge and capital cost of the digital system are higher than that of the analog technique. In regard to earth segment costs, the converse is true. This implies that from the transmission efficiency point of view the compressed speech analog technique is more efficient. However, the currently projected terminal equipment cost for the compressed speech analog scheme is high. This indicates that effort should be directed at further examining techniques of reducing these terminal equipment costs. The digital technique has two significant advantages over any analog scheme, however, and these are:

- Digital modulation is capable of wide band data transmission at very low cost.

- Digital signals are inherently resilient to interference which is a major factor in spot beam and frequency reuse environment.

Since the costs are not significantly different in the models considered, on balance digital modulation is therefore recommended.

Remote Communications

1. Although three models are shown, there are in fact only two since models Z1 and Z3 are the same. It can be seen that model Z2 or the Beam Plan (B) model with 14 uniform size antenna beams is slightly cheaper than Model Z1/Z3 which has five medium-size antenna beams. As indicated in chapter 6, model Z1/Z3 is far more flexible technically and operationally, since every remote station can be connected to every other remote station by single hop, provided demand assignment is used. Therefore model Z1/Z3 is to be preferred and this is in line with the choice of Beam Plan (C) for the major traffic scheme discussed above.

2. The Total Annual Charge for the remote service system is about \$8.4 million. Compared against the \$10.9 million annual charge for the major traffic system, this service is not cheap. The space segment accounts for about one-third of this cost and the earth segment for about two-thirds. Referring to Figure 7-21 the shapes of the trade-off curves indicate that around the optimum point, the space segment cost curve is relatively flat, showing that satellite power is not the major controlling

factor. The total system cost is largely controlled by the earth segment and this is to be expected on account of the large number of earth terminals involved. At around an earth station G/T of 24 dB, which corresponds to an antenna size of about 17 feet, the earth segment cost climbs steeply. As shown in the previous study[14], an antenna size of 17 feet is about the top limit for earth stations without tracking. The steep climb in cost beyond 17 feet is due to the addition of tracking and to the fact that larger antennas are manufactured by techniques other than the spun technique used for the smaller dishes.

3. The earth segment cost curve of Figure 7-21 also indicates that there is very little reduction in cost as the antenna becomes smaller. Indeed, the earth segment cost is mainly controlled by the electronic equipment cost and fixed charges such as buildings, power supplies, installation, etc. Hence it is quite clear that in order to reduce remote communications cost the effort should clearly be directed at cost reduction of electronic equipment manufacture, packaging and installation, and not at increased satellite power or reduced earth station antenna size.

While the above draws the conclusion that Beam Plan (C) with four spot beams is the best and notwithstanding the fact that the beam size is dictated to a large extent by the coverage zone, there remains the nagging question as to how close the above beam size is to a purely theoretical optimum beam. A purely theoretical optimum beam is defined as the optimum trade-off between the antenna weight as the beam size gets smaller and smaller and the TWT and solar cell weight as the rf power increases. In this hypothetical approach, the effect of beam size on the geographical coverage area and all the other side effects that go along with it are ignored. A computer run for such a trade-off was made for the non-switched model and this could be considered as representative. The results are shown in Figure 8-3. It can be seen that although the beam size chosen is not exactly at the theoretical optimum, it is not, from the annual charge point of view, significantly far off.

It is of interest to analyze the curve of Figure 8-3 in three segments. On the left where the cost rises steeply, the antenna weight is obviously the main cost element as the beam size gets very small and the antenna gets correspondingly large. On the right side of the curve, for a constant earth station G/T, the cost curve again rises steeply. This rising curve reflects the increase in cost due to TWT power and associated solar cell weight. In the middle portion of the curve, the cost is relatively flat. The model considered here lies in this portion. This part of the curve indicates that the cost is mainly controlled by equipment deadweight. That satellite deadweight is in control of space segment cost is also evident from the fact that from the trade-off curves of chapter 7, a large increase or decrease of earth station

G/T is required in order to increase the annual charge by 5 percent from the optimum. It is due to this deadweight control that simplicity of system design on board the satellite has significant advantage. This accounts for the superiority of the four-beam plan. A lower number of beams also brings along another important advantage and that is network flexibility. Hence, while there are good reasons for a multiple beam design there are also counter arguments and the final solution, as in most engineering ventures, would inevitably be a compromise.

A similar hypothetical exercise was also carried out for the remote communications antenna beam size. The results are indicated in Figure 8-4. It can be seen that by fortuitous coincidence, the antenna beam size chosen lies about on the optimum point. The beam size was chosen without this knowledge in mind but rather was chosen to serve the regional interests of the educational service in the previous report[14].

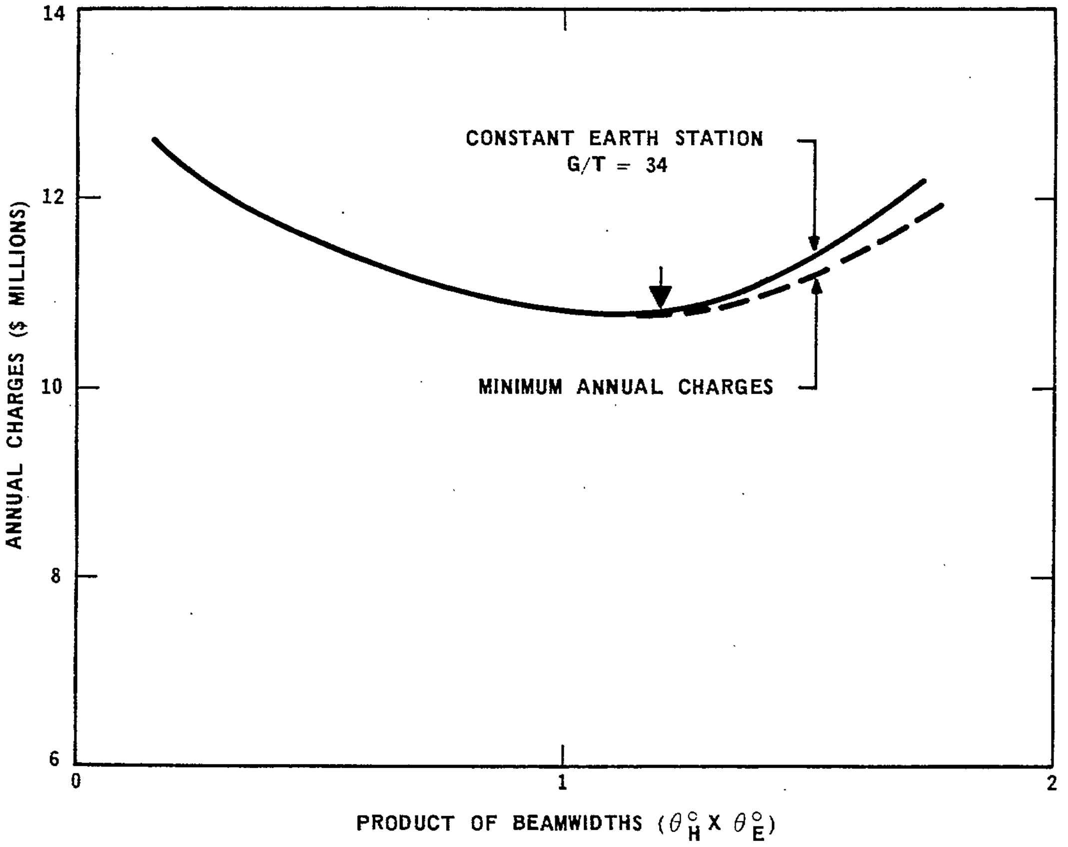

Figure 8-3 Configuration W3

Effect of Beam Size Change

(Arrow shows position of Selected Model)

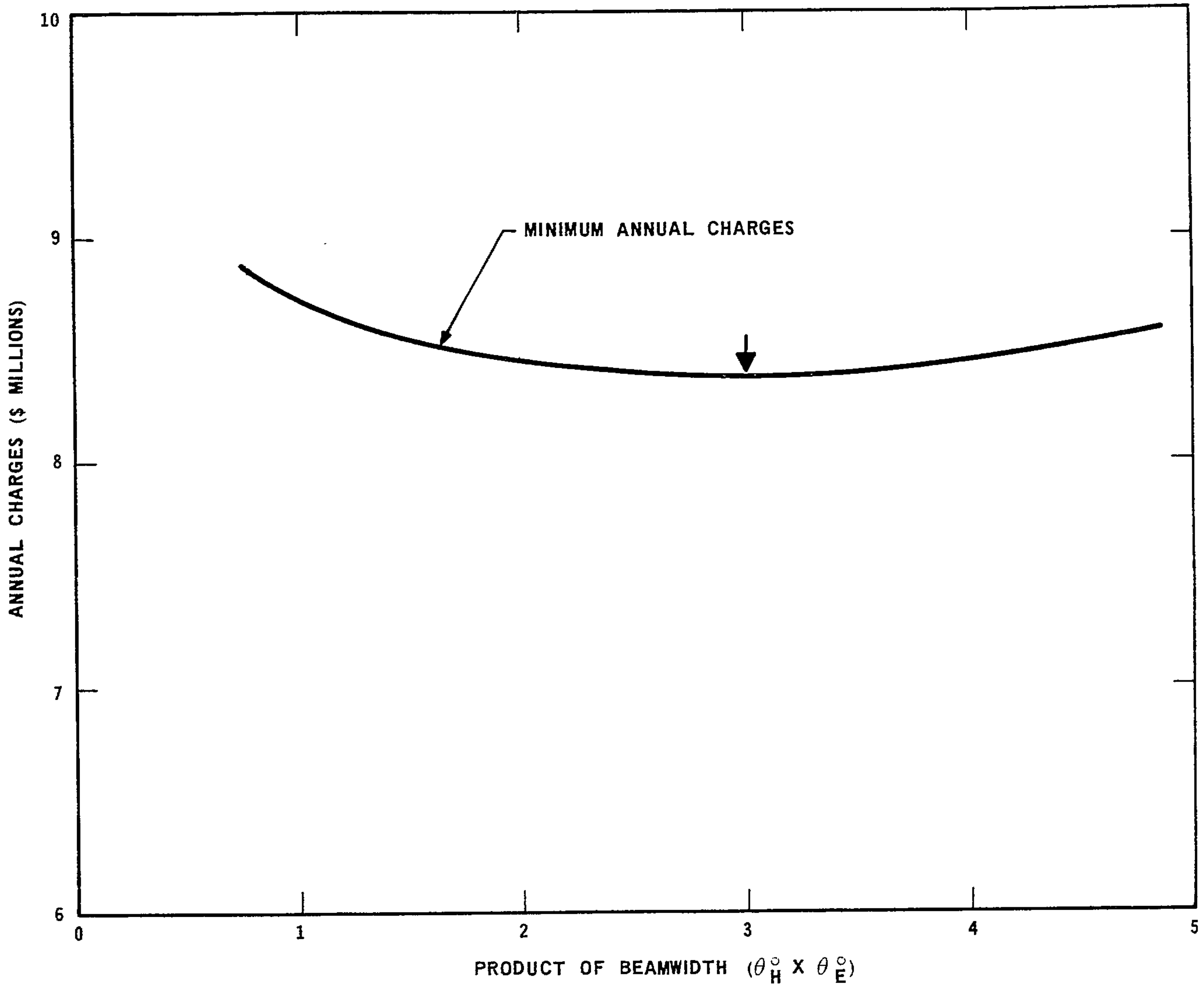

Figure 8-4 Configuration Z1/Z3

Effect of Beam Size Change

(Arrow shows position of Selected Model)

Table 8-3 summarizes the characteristics of the preferred major route and remote service models. The parameters of the best of the on-board switched models are also given for comparison.

Table 8-3

	MAJOR TRAFFIC SERVICE		REMOTE SERVICE
	SWITCHED MODEL (FOR REFERENCE)	RECOMMENDED MODELS	
		UNSWITCHED	UNSWITCHED
Beam plan	C	C	C
Number of beams	4	4	5
Size of beams	1.7°× 0.7°	1.7°× 0.7°	2.0°× 1.5°
Modulation	SDMA/SS-TDMA	PCM-PSK-TDMA (4ϕ)	SCPC-FM**
Dual polarization frequency reuse	yes	yes	no
RF channel bandwidth	125 MHz	62.5 MHz	31.25 MHz
Transponder TWT saturated power	9 W	1.1 W	4 W
Number of TWT amplifiers	2 per beam	4 per beam	1 shared by 5 beams
Satellite weight contribution	489 lb	491 lb	263 lb
Earth station G/T	34 dB	40 dB	24 dB
Earth station antenna diameter	29 ft	59 ft	17 ft
Earth station Tx power	927 W	292 W	70 mW
Space segment capacity	24 000 Ch	24 000 Ch	1200 Ch
Earth segment capacity equipped	10 000 Ch	10 000 Ch	1200 Ch
Number of earth stations	12	12	400
Space segment capital cost	$\$28 \times 10^6$	$\$28 \times 10^6$	$\$15 \times 10^6$
Earth segment capital cost	$\$18 \times 10^6$	$\$21 \times 10^6$	$\$20 \times 10^6$
Space segment annual charge	$\$5.0 \times 10^6$	$\$5.0 \times 10^6$	$\$2.6 \times 10^6$
Earth segment annual charge	$\$5.0 \times 10^6$	$\$5.9 \times 10^6$	$\$5.7 \times 10^6$
Space segment annual charge /Ch *	$208	$208	$2170
Earth segment annual charge /Ch *	$500	$590	$4750
Total per channel annual charges ***	$1000	$1090	$6920

* Based on 100% utilization.

** With Lincompex.

*** Based on utilization of 10 000 channels of space and earth segments for major traffic and 1200 channels for remote service.

It will be seen from this table that the total capital cost and annual charges for the major traffic system are $49 million and $10.9 million respectively. The system provides a network of effectively 10 000 channels. Compared against the results of the previous study in which half Canada coverage beams were postulated and no frequency reuse was employed, the current multibeam model provides about three times the network capacity at about twice the cost. It should be pointed out that there are two significant differences in the approach in the two studies. In the current study, a total of three satellites are used - one in operation, one in orbit standby and one spare - and launch costs of the three are included. In the previous study, only two satellites, together with launch, were postulated. In the present study, in order to enable direct cost comparison with the single-channel-per-carrier system, all systems have included the terminal multiplex equipment cost. In the previous study, the multiplex equipment cost was considered outside the satellite link. Hence taking the above differences into account, it can be seen that the multibeam satellite system is indeed cost effective for the major route service.

In the case of the remote communications, the total capital cost and annual charges are $35 million and $8.4 million respectively. The annual charges are about two and a half times that of the previous study in which a Canada-wide coverage beam was used. The network capacity of the current model, both in terms of the number of channels and in the number of earth terminals, is twice that of the previous study. Hence it would seem that the current multibeam approach is much more expensive. However, after allowing for the difference in the number of launched satellites, the difference in the annual charges between the models in the two studies would amount to only about 10 percent in favor of the single Canada-wide beam model of the previous study. The major contribution to the higher cost of the multibeam approach is the vastly increased amount of electronic and interbeam connecting equipment on board the satellite.

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9. CONCLUSIONS AND RECOMMENDATIONS

The conclusions and recommendations of the study are presented.

Conclusions

The objectives of the study, as defined in chapter 1, have all been met. The following conclusions may be drawn:

1. A satellite system with multiple spot beams and frequency reuse will be needed, if the stated forecast traffic of 10 000 one-way channels for the period 1977 through 1985 materializes, and if the exclusive satellite bands of 14.0 to 14.5 GHz and 11.7 to 12.2 GHz are to be shared among the communication and television services in Canada.

2. The optimum configuration is a satellite system with four medium size beams (0.7° x 1.7°) for the major traffic network, covering the traffic zones as follows:

   - Beam 1: Vancouver, Calgary, Edmonton
   - Beam 2: Regina, Winnipeg
   - Beam 3: Toronto, Ottawa, Montreal, Quebec City
   - Beam 4: Halifax, Moncton, St. John's, Newfoundland

3. The non-switched model is recommended for the major route service. Based on current forecasts of future development, the annual charges for switched and non-switched systems are not significantly different.

4. Within the non-switched model, the modulation recommended is PCM-4ØPSK-TDMA.

5. Speech compressed (lincompex) analog modulation has a higher transmission efficiency and hence a lower space segment cost than the digital technique. The currently projected earth segment terminal equipment cost is, however, high.

6. For the remote service, a five-beam configuration, with single-channel-per-carrier speech compressed frequency modulation is the optimum choice. This configuration gives complete network flexibility since any earth terminal can be linked to any other earth terminal via one hop, assuming demand assignment is used. The five beams cover all of Canada on a regional basis. The beam sizes are approximately 1.5° × 2.0°.

7. Total satellite weight for the two services amounts to about 750 pounds. This is well within the Thor-Delta launch capability (1000 pounds) although this would not leave much for any television services that may be put on. Much of the weight is basic equipment deadweight due to the increased number of electronic subsystems on-board and this is an inherent drawback of a multi-beam satellite system. A trade-off exercise of satellite antenna size and weight versus TWT power and weight indicates that the beam sizes chosen for the two services are close to optimum.

8. While a multibeam satellite system has been shown to be cost effective for the major route service, its most significant advantage appears to be its ability to provide vastly increased system capacity through frequency reuse. As frequency spectrum and orbital space are valuable natural resources, the use of a multibeam satellite appears fully justified on this count alone.

Recommendations

The study has examined all the major aspects of a satellite communications system using multiple spot beams. In carrying out this task, naturally, the best currently available knowledge has been applied. Nevertheless, in certain critical areas current knowledge may not be all that is desired. It is therefore recommended that future effort should be directed at these areas. The recommendations are as follows:

1. The satellite propagation characteristics in the 12 to 14 GHz band should be investigated experimentally to determine more accurately:

- The fade margins
- Cross-polarization coupling resulting from rain scatter and other propagation phenomena.
- Faraday rotation
- Space diversity improvement

2. Satellite up-link power control is postulated in this study. This technique and also the technique of an overall satellite link power control with feedback channel should be investigated experimentally to prove out the technical feasibility.

3. Speech compression and the general question of speech signal processing (digital, analog and time interpolation) are worthy of further investigation, as they hold potentials of decreasing transmission cost and increased efficiencies in bandwidth and power utilization. Further, they

have applications in almost all transmission media, both satellite and terrestrial. Terminal equipment cost is high at present but with the possibility of using digital techniques and hence large scale integrated circuitry, this may be vastly reduced.

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APPENDIX A

Multibeam Satellite Antenna Technical Appraisal

A.1 GENERAL DESIGN APPROACH

There are five types of antennas which may be used for a communication satellite system; reflector lens, horn, traveling wave and array. The single horn antenna is too heavy at the specified beamwidth and therefore no further discussion will be given. The traveling wave antenna is generally frequency sensitive and is not practical as a high gain antenna in the specified frequency band. The reflector antenna and array antenna are the ones most suited for communication satellite use.

In theory, if a perfectly linear or circular polarized element antenna is used for the array, the resultant pattern will be purely polarized. However, the problems of designing a practical linear or circular polarized element and of polarization degradation due to mutual coupling effects cannot be solved at this time. Moreover, the bandwidth requirement will not be met in general, the insertion loss is high and the antenna will likely be restricted to a single beam. Considering all the factors, the array antenna is not the best choice for communication satellite use at present.

Both lens antenna (metal lens or dielectric lens) and reflector antenna are designed based on geometrical optics. There are no fundamental differences between these two types of antenna from the viewpoint of electrical performances. Since there are no blockage effects for the lens antenna it is conceivable that the side lobe level near the main beam region may be lower for the lens antenna than for the reflector antenna. However, the feedhorn forward spillover in the lens antenna will increase the near side lobe level. Moreover, the lens antenna in general is very heavy, the insertion loss is relatively high compared to the reflector antenna and its deployability is poor. Based on the progress of present technology, the performance of lens antennas will not likely lead to the use of this class of antennas in the near future for communication satellites.

It is a fundamental characteristic of the electromagnetic wave that if a linearly polarized wave is incident on a curved surface, the reflected wave is generally elliptically polarized. For a focus-fed reflector antenna with a linearly polarized horn, the far-field pattern of the cross-polarized component is mainly due to the curved reflecting surface. The maximum value of the cross-polarized pattern is approximately -30 dB with respect to the maximum of the main polarization. If a circularly polarized feed horn is used, the maximum value of the cross-polarized component can be reduced considerably, and its contribution to the beam isolation degradation will be negligible compared to the side lobe envelope. However, a circularly polarized feed horn requires a polarizer which changes a linearly polarized wave to a circularly polarized wave, or vice versa. It has been shown (Dudzinsky 1969) that 0.25 dB axial ratio polarizers are required in order to reduce the interference signal level to -30 dB. Such a requirement is difficult to achieve at present. The advantage of circular polarization reduces because of the requirements for other components in the system. For a properly designed reflector antenna with a linearly polarized feed, it is felt that the isolation requirements between antenna beams can be met, as shown in the following sections. Linear polarization is recommended and used in the following analysis. However, if circular polarization is required and the problem of design of near perfect polarizers is solved, the analysis and design in this study can be adopted to determine antenna performance in the circularly polarized case.

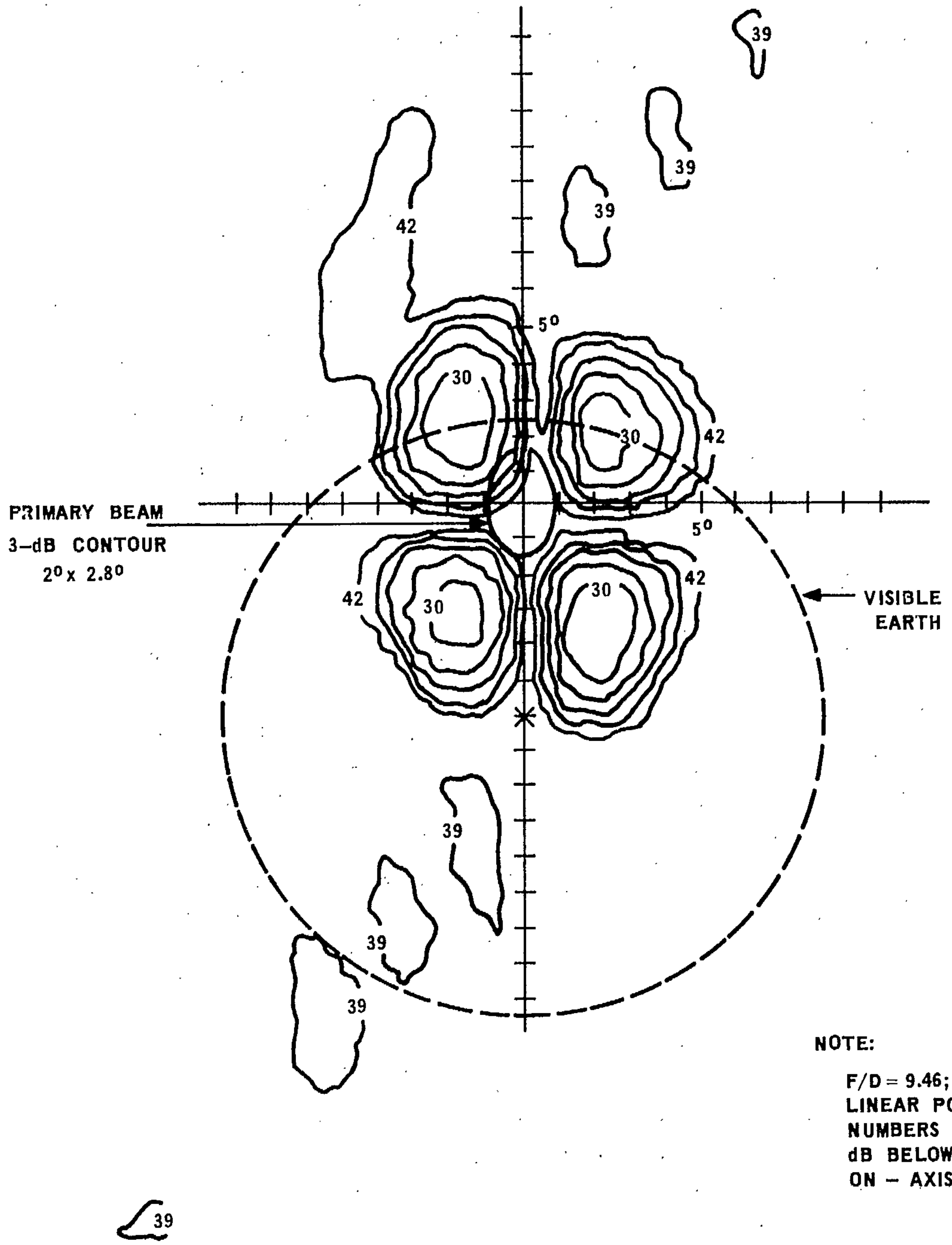
A.2 SINGLE BEAM ANTENNA

Maximum beam isolation is achieved through the use of separate reflectors, each with its own feed. This configuration avoids the defocussing problems associated with multiple feed structures and reduces the mutual coupling between feed horns.

Polarization interference is dependent upon the antenna's cross-polarized side lobe behaviour and mutual coupling between feed horns. The different sources of interference become important for various beam separation angles.

For small separations feed horn coupling, in the case of a multiple feed antennas, becomes dominant. Separate feed-reflector systems are not affected by mutual coupling and for this type of antenna the polarization isolation is limited by the cross-polarized lobes surrounding the main beam. Figure A-1 is a contour map of the measured cross-polarized response of a satellite antenna using a prime focus feed and an f/D of 0.46. Maximum cross-polarized levels of this antenna are 27 dB down, referred to the on-axis co-polarized gain. Along the E and H planes (vertical and horizontal in Figure A-1), the cross-polarized components are at least 35 dB down.

For wide angle separation the isolation is limited by the side lobe levels of the main and cross-polarized fields. The side lobes are dependent upon illumination taper, f/D, and blockage. Generally, when aperture efficiency is not critical the illumination taper can be chosen to suppress side lobes to approximately -30 dB.



NOTE:

F/D = 9.46; PRIME FOCUS.
LINEAR POLARIZATION.
NUMBERS GIVEN REFER TO
dB BELOW CO-POLARIZED.
ON – AXIS RESPONSE.

Figure A-1 Measured Cross-Polarized Response, Typical Satellite Antenna. (From CCIR Study Group, 3 June 1969, DOC IV/268E)

Figure A-2 compares the theoretical response curves for aperture antennas with various amplitude illuminations with a $[1 + (\theta/\theta_0)^4]^{-1}$ envelope. It is apparent that a cos $(\frac{\pi}{2}x)$ distribution satisfies the discrimination predicted by the $(\theta/\theta_0)^4$ curve. For a $\cos^2 (\frac{\pi}{2} x)$ distribution side lobes are suppressed below the -30 dB level and the $(\theta/\theta_0)^4$ curve is exceeded. In the antenna system being considered the aperture efficiency can be reduced to obtain proper side lobe behaviour.

Blockage effects, depending upon the antenna configuration then become limiting. For a prime focus, on-axis feed blockage from the feed and support structure will create side lobes at the -25 dB level. The blockage conditions also favor a lower f/D ratio to minimize feed horn dimensions and support structure. The deeper dish tends to increase the cross-polarized lobes due to the greater dish curvature. Figure A-3 is a comparison of the azimuth and elevation patterns of an on-axis prime-focus feed antenna. The feed is supported by a tripod in the vertical plane thus creating azimuthal plane side lobes. The elevation pattern is not affected by the tripod blockage and in this plane the side lobes are generally at the -30 dB level.

An investigation by Cowan and Zierman on the effects of f/D and aperture blockage shows a 5 dB degradation in side lobe level due to blockage from the feed and the waveguide run. Figure A-4 shows the side lobe level can be minimized for an f/D of 0.34 assuming blockage from the feed and waveguide but when blockage is not present the optimum f/D is 0.38. The predicted optimum side lobe levels for the two cases are -27 and -34 dB respectively.

Considering the overall interference problem of two optimized antennas, the greatest interference will occur when one of the cross-polarized maxima falls on the communication area of the other beam. Present antenna systems described to date have cross-polarized signal levels of -27 to -30 dB in this region. This interference zone extends from near zero (at zero separation the cross-polarized signal is at a null for symmetrical antennas) to approximately one-half power beam width. For greater beam separation angles, the isolation will be between the cross-polarized lobes of one beam and the co-polarized main lobe of the other beam. Isolation under these conditions will be much greater since the side lobes can be reduced to -20 to -30 dB below the main beam and the cross-polarized side-lobe level is generally 10 dB lower than the co-polarized side-lobe level. Thus the overall coupling between a wanted main lobe and an unwanted cross-polarized lobe will be of the order of -30 to -40 dB. As the beam separation increases, side lobe levels will drop and the isolation will increase. The limiting value will be determined by the wide angle random cross-polarized signal. This level is affected by blockage and surface error.

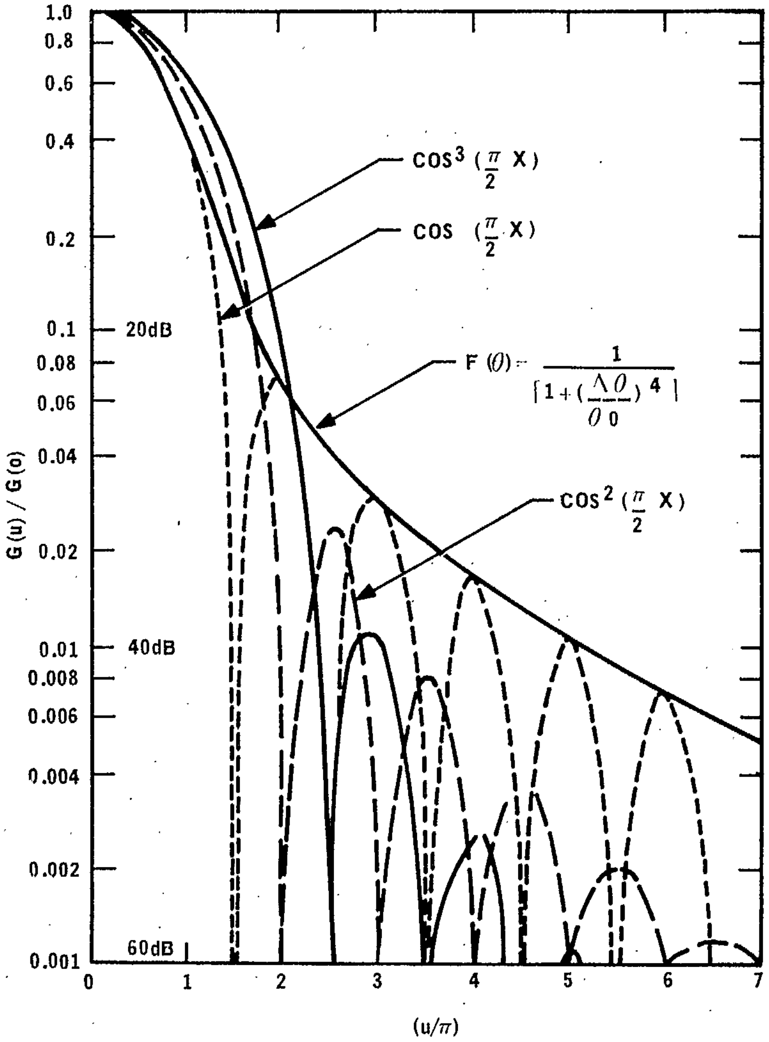


Figure A-2 Calculated Antenna Pattern for $\cos^n[(\pi/2)x]$ Illumination of Square Aperture. a is Aperture Opening. θ is Angle With Normal to Aperture. $u \equiv (\pi/\lambda)$ a $\sin\theta$. (From Tillotson, 1968).

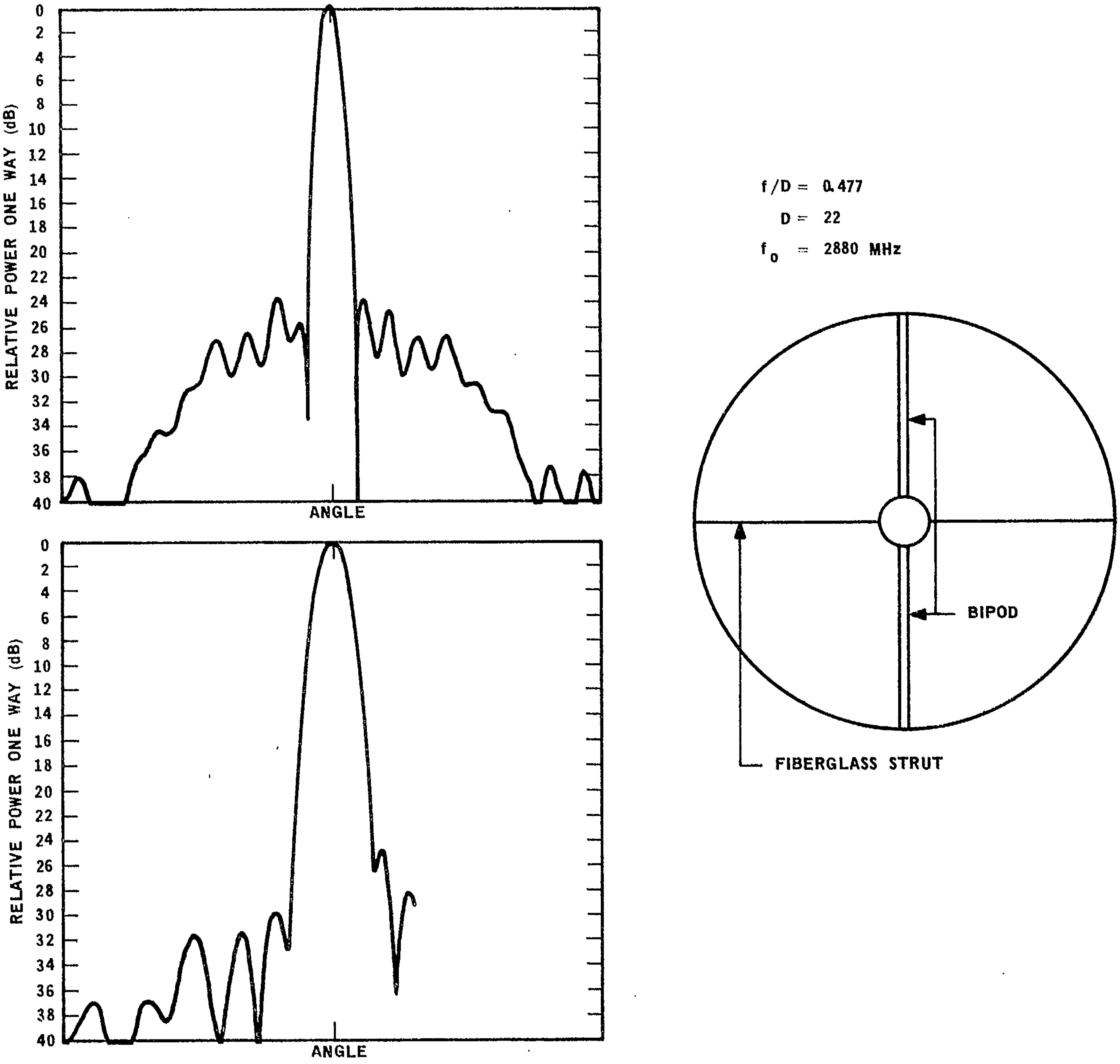


Figure A-3 Azimuth and Elevation Patterns for Typical Polarization Setting.
(From L.E. Allan, R.C. Markell and G.C. McCormick, 1967).

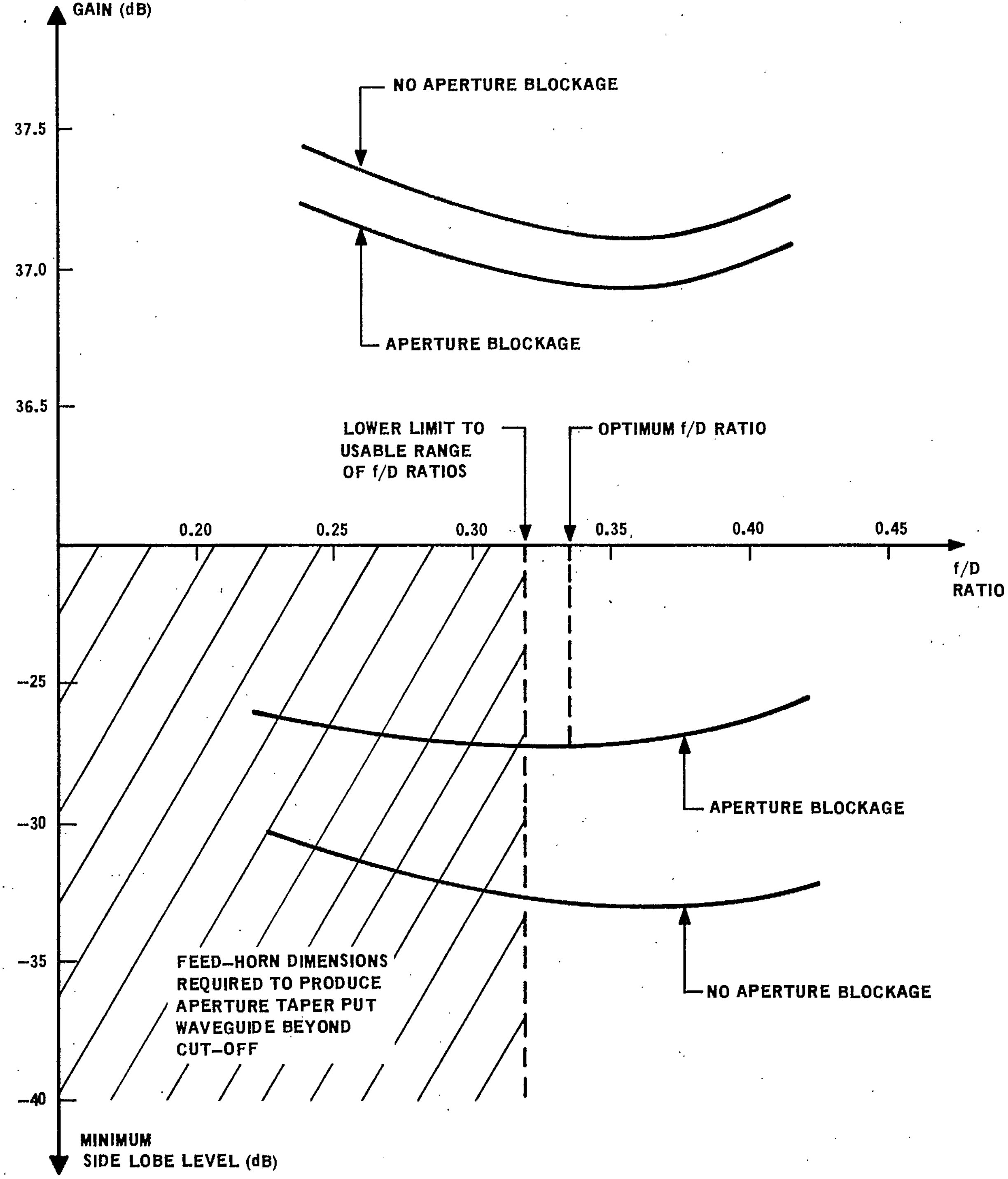


Figure A-4 Prime Focus Reflector, On-Axis Feed: Optimum Sidelobe Performance as a Function of F/D Ratio (From J.H. Cowan and C.A. Zierman, 1970).

A.2.1 Antenna Configurations Considered

A.2.1.1 The Prime Focus On-Axis Feed

This antenna has low cross-polarized components for large *f*/*D* ratios. However, it is subject to blockage from the feed and the waveguide run.

A.2.1.2 On-Axis Cassegrain Feed

Due to the magnification of the focal length, the cross-polarized fields are reduced. However, increased blockage will increase side lobes to an unacceptable level. In addition, for the small dishes required for this application, the hyperbolic mirror will not function well and this will lead to increased diffraction losses.

A.2.1.3 Prime Focus - Offset Feed

The offset feed configuration eliminates all blockage and thus minimum level side lobes are attainable. However, the asymmetrical feed arrangement results in increased cross-polarized levels in the far field pattern. The offset antenna has a smaller effective *f*/*D* ratio than a center-feed antenna of the same focal length and aperture. Thus the dish curvature is greater giving rise to larger cross-polarized components in the radiating aperture. In addition, for a totally offset feed there are only two cross-polarized regions on the dish as opposed to four regions for the center-feed antenna. The result is that the far field pattern has only two cross-polarized lobes instead of the four lobes of conventional antennas as shown in Figure A-1. The cross-polarized pattern has a higher aperture efficiency giving narrower beamwidths and a peak level approximately 6 dB greater than the four-lobe distribution.

A.2.1.4 Double Parabolic Cylinder Antenna

Typical microwave pencil-beam antennas of the reflector type are based on using a paraboloid of revolution to transform a spherical wave, arising from a point source, into a plane wavefront. A similar effect can be achieved by a system of two crossed parabolic cylinder reflectors (Spencer, Holt, etc., 1955) where each cylinder collimates the beam in one plane. A significant difference between the two antenna configurations for the current application lies in their polarization characteristics. In the case of the paraboloidal reflector, a cross-polarized component is introduced into the radiated field purely by the geometry of the system. For the double parabolic cylinder antenna, on the other hand, there exists a feed orientation for which no cross-polarized component is produced by the reflectors.

A typical configuration of reflectors is illustrated in Figure A-5. Radiation from a point source placed on the focal line of the first parabolic cylinder is transformed into a cylindrical wave. The second parabolic reflector, crossed so that its focal line coincides with the directrix of the first reflector, transforms the cylindrical wave into a plane wave. Using geometrical optics, it can be shown that if the electric field vector of the primary source is oriented parallel to the focal line of the first cylinder then the polarization of the reflected rays will be independent of the directions of the rays leaving the source. Thus, the reflected rays will all be polarized in the same direction.

Although the double reflector configuration appears to be desirable from the viewpoint of polarization purity, a detailed analysis of this system will be required to evaluate various effects which can influence its performance. For example, the geometry of a physically practical two-reflector configuration tends to introduce some asymmetry into the aperture amplitude illumination. (This can be partly compensated for by appropriately orienting the feed horn.) Since the two reflectors can be relatively close to each other, the possibility of interference between them will have to be considered. The analysis will also be required to include the effects of various parameters such as feed horn location, separation between reflectors, and focal length of the reflectors. It may be noted that the double cylindrical reflector configuration lends itself to beam shaping, since the two reflectors afford essentially independent control of the beam shapes in orthogonal planes.

A.3 MULTIPLE BEAM ANTENNA

In order to reduce the weight and volume of the antenna, it would be desirable to have the reflector shared by more than one beam by using multiple feed horns. However, it seems inevitable that the electrical performance of the antenna system will be degradated and some additional mechanical requirements may be imposed. The purpose of this section is to discuss various forms of multiple beam antennas in terms of far field pattern and cross-polarization component.

A.3.1 Design Criteria For Multiple Beam Antennas

There are four types of multiple-beam-reflector antennas; paraboloidal reflector with multiple feed horns, torus antenna, bifocal two-reflector antenna and dual cylindrical parabolic reflector

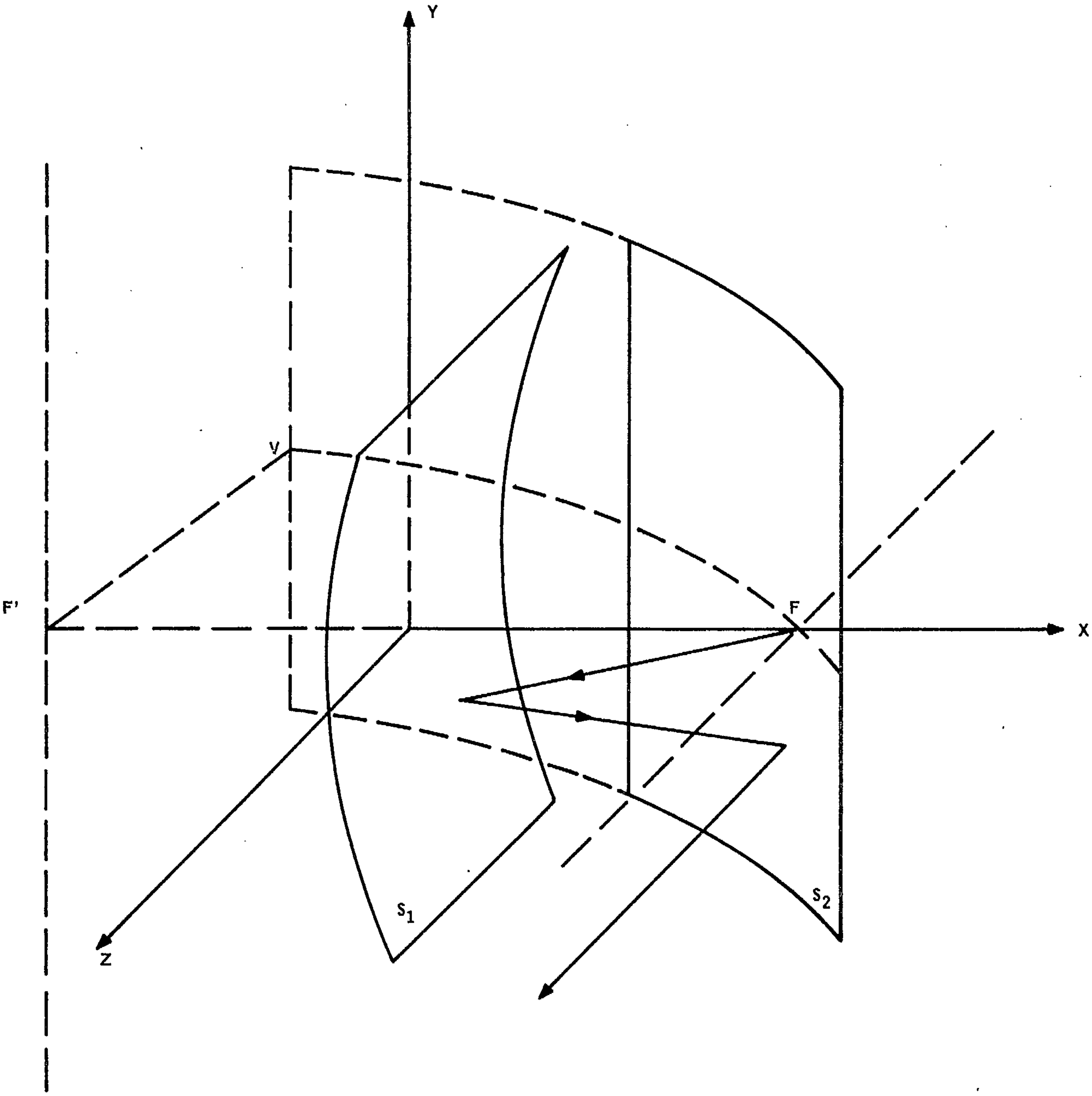


Figure A-5 Double Parabolic Cylinder Antenna

antenna. The basic design principles used for the single-beam antenna such as large effective focal length to diameter ratio, offset reflector, symmetric low side-lobe feed horns, etc., are still applicable for the multiple beam antenna. Some additional design parameters and limitations for various types of antennas are discussed in the sections that follow.

A.3.2 Paraboloidal Reflector

It is well known (Kelleher and Coleman 1952, Lo 1960, Sandler 1960, Ruze 1965) that the antenna beam pointing direction of a paraboloidal reflector can be deviated from the boresight direction if the feed horn is moved laterally off the focus. There is no appreciable degradation in gain or in far field pattern if the antenna beam is pointing up to approximately two beamwidths off the boresight direction. Assuming the cross-polarized far field pattern does not change and the coupling between feed horns is negligible, if multiple feed horns are placed in the neighborhood of focus this maximum beam separation is dictated by comatic aberration.

The minimum beam separation depends on the reflector parameters and the feed horn aperture size. Let D and f be the diameter and focal length of the reflector and d be the lateral distance between the horn and focus. Then,

$$B_W = K_1/D$$

$$\theta_s = K_2\ D/f$$

$$A_H = K_3/F(\frac{f}{D})$$

where K_1, K_2 and K_3 are proportional constants. B_W, θs and A_H are the beamwidth, beam deviation angle and feed horn aperture diameter, and $F(f/D) = 2\ \arctan(\frac{1}{2f/D - D/8f})$. If $(1 - r^2)$ type aperture distribution with 10 dB edge taper is used and the feed horn pattern behaves as an ordinary E-plane sector horn, then

$$K_1 = 1.27\lambda, \qquad K_3 = 2.11\lambda$$

where λ is the wavelength. The minimum beam separation therefore is

$$\theta_{min} = 1.055\ \lambda\ K_2/\{f\ arctan\ (\frac{1}{\frac{2\ fB_W}{1.27\lambda} - \frac{1.27\lambda}{8f\ B_W}})\}$$

It can be shown that the minimum beam separation is approximately 1.5 beamwidths.

A.3.3 Torus Antenna (Peeler and Archer 1954, Hyde 1970)

One of the problems for the multiple-beam paraboloidal reflector is that the far field pattern is deteriorated as the beam deviates from the boresight direction. If the beams share only part of the reflector instead of the whole reflector as in the case of paraboloidal reflector antenna, the far field pattern may be improved. The rectangular torus antenna allows this type of operation.

The rectangular torus antenna surface is generated by a section of parabola which is rotated about the axis perpendicular to the axis of the parabola. The focus of the parabola is approximately situated half way between the vertex and the axis of rotation. One of the advantages of the torus antenna is that all the beams suffer the same kind of phase error and there is no upper limitation on the beam separation. However, the advantage of using a common reflector diminishes as the beam separation becomes so large that the beams do not share any common area of the reflector. The minimum beam separation is determined by the feed horn physical size and is approximately the same as with the paraboloidal reflector.

A.3.4 Double Parabolic Cylinder Reflector Antenna

In order to utilize the available aperture area efficiently, one can also use the double parabolic cylinder reflector antenna as a multiple beam antenna. Although only one beam would be free of cross-polarized components due to geometry, the problems of reducing cross-polarization in the orthogonal beam could be handled in essentially the same way as for the conventional paraboloidal reflector. Beam separation could be obtained by two different methods; by tilting the feed horn in the plane containing the focal line and the axis of the first reflector, or by moving the feed horn off the focal line of the first reflector. Tilting the horn results in perfect collimation by the first reflector and it essentially places the virtual line source for the second reflector off the focal line. Moving the feed horn off the first focal line will tilt the direction of the virtual line source and a different part of the secondary reflector will be used.

There is no low limit of beam separation for this type of antenna and the upper limit of the beam separation should be approximately the same as for the paraboloidal reflector antenna.

A.3.5 Two Reflector Antenna

The most serious aberrations for the off-axis paraboloidal reflector antenna are coma and astigmatism. Comatic aberration is introduced due to violation of the Abbé sine condition and astigmatism is introduced by having different focal points in the principal planes. Aside from a reduction in gain and an increase in the cross-polarization component due to lateral feed displacement, the maximum beam separation of the focus fed paraboloidal reflector antenna is limited by comatic aberration. An over-compensated Schwarzschild system (White and DeSize 1962) can be designed free of comatic aberration and therefore capable of producing multiple beams at large angle separations. The limiting factor then becomes astigmatism which is very small for a beam separation of 12 degrees. The minimum beam separation due to the feed horn size is similar to the paraboloidal reflector.

If only two beams are required, it may be possible to design a system using two reflecting surfaces and having two perfect focal points (focal points at infinity are not included) in the sense of geometrical optics (Holt and Mayer 1957, Brown 1956). Perfect collimation in two specified directions will result. The system may be extended to more than two beams by placing additional feed horns at the proper positions. The minimum beam separation depends on the feed horn aperture dimensions.

A.3.6 Far-Field Pattern

The far-field pattern of a reflector antenna may be obtained by the current distribution method and is given by

$$\vec{E}(\theta,\phi) = -j\,\frac{\omega\mu}{2\pi}\,\exp\left(\frac{-jkR}{R}\right)\,\left[\left(\frac{\varepsilon}{\mu}\right)^{\frac{1}{2}}\frac{P}{2\pi}\right]^{\frac{1}{2}}\int_s \frac{\vec{n}'\times(\vec{r}'\times\vec{E}')}{|r'|^2}\,\exp\left[-jk\left(r' - \frac{\vec{r}'\times\vec{i}_R}{r'}\right)ds\right. \qquad (1)$$

where $\vec{i}_R$ is the unit direction vector of the observing point

$\vec{r}'$ is the position vector of element area ds from feed horn phase center

$\vec{n}'$ is the unit normal of ds

$\vec{E}'$ is the electric field distribution of feed horn

μ is the permeability

k is wave number

ω is the frequency in rad/s

ε is the permittivity

The integration is over the reflecting surface. In the case of a two reflector antenna, equation (1) will be used twice; first, for calculating the reflected field at the second reflecting surface due to the first reflector and, second, for calculating the far field pattern from the induced currents on the second reflecting surface. Most of the theoretical analysis of the far-field pattern can be obtained by using equation (1).

It has been shown (Sandler 1960) that for a central open waveguide fed paraboloidal reflector antenna the first two side-lobe levels in the H-plane are -15.2 dB and -23 dB respectively if the antenna beam is pointed at 1.85 beamwidths from the boresight direction, and the maximum cross-polarization component becomes -25 dB if the beam is deviated for 1.12 beamwidths. With properly designed feed horns it is possible that the side-lobe levels of all the beams can be reduced, but the reduction of cross-polarized components requires further investigation.

Since the two reflector system offers a slightly lower cross-polarized component due to the long effective focal length, and shaped reflecting surfaces will have less phase error for each antenna beam, it is felt that the electrical performance of the two-reflector multiple-beam antenna would be better than the multi-horn focus fed paraboloidal reflector antenna.

A.3.7 Mutual Coupling

There are two types of mutual coupling for multiple beam antennas; the mutual coupling between feed horns and the mutual coupling between antennas. The feed horn far field pattern may be changed if a second horn is placed in the vicinity of the near field (Kuecken 1957, Lyon et al. 1964). For low side-lobe level and high gain horns, the coupling occurs mainly via the side lobes and the phase center of individual horns may be varied very slightly. A certain amount of experimental adjustment will be required in order to achieve the correct primary feed patterns and cross-polarization levels. One of the major difficulties which may be encountered in designing multiple beam antennas is that the beam directions may vary with frequency owing to phase center variations of the individual feed horns.

The antenna far field pattern may also be degraded because of other antennas being placed in the vicinity of the near field. The mutual coupling can be reduced by properly choosing the separation between antennas.

Another type of antenna mutual coupling is due to the feed horn side lobes which illuminate the adjacent reflecting surfaces. Since the side lobes are usually elliptically polarized and the source position is well off the focus, a high-level cross-polarized far field pattern may result. This type of mutual coupling can be eliminated by using feed horns with low side lobes and by orienting the feed horns in such a way that the reflector will not intercept the side lobe energy.

A.3.8 Possible Antenna Configurations

There are four multiple beam designs which seem promising at this time. They are the parabolic reflector with on-axis focal feeds, the paraboloidal reflector with off-axis feeds, the two-reflector system and the double parabolic cylinder reflector system. The advantages and disadvantages of each are summarized as follows:

a) Paraboloidal reflector with on-axis focal feeds..

- Advantages: Simple to design, light in weight, easy to deploy in space.
- Disadvantages: Minimum beam separation is approximately 1.5 beamwidth; possible inferior performance for large beam separation.
- Possible Difficulties: Mutual coupling between horns.

b) Paraboloidal reflector with off-axis feeds.

- Advantages: Simple to design, light in weight.
- Disadvantages: Minimum beam separation is approximately 1.5 beamwidth; possible inferior performance for large beam separation.
- Possible Difficulties: Mutual coupling between horns.

c) Two-reflector system.

- Advantages: Possible good electrical performance for all the beams.
- Disadvantages: Difficult to design; minimum beam separation may be similar to paraboloidal reflector.

- Possible Difficulties: Mutual coupling between horns. Deployment in space may be difficult.

d) Double parabolic cylinder antenna.

- Advantages: Simple to design; no minimum beam separation restriction; cross-polarized component may be controlled; mutual coupling between horns may be reduced by increasing the separation between horns.

- Disadvantages: Heavy; large reflector size; possible inferior performance for large beam separation; almost rectangular (or square) aperture.

- Possible Difficulties: Asymmetry of the far field pattern. Deployment in space may be difficult.

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APPENDIX B

Considerations of Beam Pointing Stability and Interference Objectives Affecting Optimum Beam Size Design for Canada

The antenna peak side lobe envelope, as indicated in chapter 3, can be expressed as a function of the angular deviation (θ_I) from the boresight axis, i.e.

$$F(\theta_I) = \frac{1}{1 + (\frac{\theta_I}{\theta_0})^4} \qquad \text{(B1)}$$

where θ_0 = half the 3 dB beamwidth.

It is convenient to convert this relationship into discrimination(D):

$$D = 1 + (\frac{\theta_I}{\theta_0})^4 \qquad \text{(B2)}$$

For large values of D(>10) this can be approximated by

$$D = (\frac{\theta_I}{\theta_0})^4$$

or, in decibels, $D = 40 \log (\frac{\theta_I}{\theta_0})$ dB. (B3)

Assuming that equal powers are fed to two beams, one wanted and the other interfering, and as a worst case situation, the earth station is located at the 3 dB beam edge of the wanted beam, then the down-link carrier-to-interference ratio $(C/I)_d$ is given by:

$$(C/I)_d = 40 \log (\frac{\theta_I}{\theta_0}) - 3 \text{ dB} \qquad \text{(B5)}$$

For a homogeneous system, the up-link $(C/I)_u$ would also be the same and hence the total $(C/I)_t$ would be given by:

$$(C/I)_t = 40 \log (\frac{\theta_I}{\theta_0}) - 6 \text{ dB} \qquad \text{(B6)}$$

If there is another interfering beam but located on the opposite side to the interfering beam above, then the interference angle is $(\theta_I + 2\theta_0)$. In general, for the beam sizes and beam separations considered in this report, this second interfering beam angle is wide enough not to cause a significant contribution. Similarly, other beams further away on either side would not contribute any significant interference. Hence equation (B6) may be taken, as a first approximation, to be representative of the total carrier-to-interference ratio.

If a total $(C/I)_t$ of 15 dB is acceptable then using equation (B6):

$$15 = 40 \log \left(\frac{\theta_I}{\theta_0}\right) - 6$$

$$\therefore \quad \frac{\theta_I}{\theta_0} = 3.35 \qquad \text{(B7)}$$

If the satellite antenna beam pointing stability is ±P, then the effective coverage beamwidth is given by:

$$\theta_\varepsilon = 2\theta_0 - 2P \qquad \text{(B8)}$$

For a satellite system in which one-third frequency reuse is employed (see Beam Plan B, chapter 3), it is easy to show that:

$$\theta_I = \frac{5}{2}\,\theta_\varepsilon - P \qquad \text{(B9)}$$

Combining equations (B7), (B8) and (B9), and putting the beam pointing stability $P = 0.1°$ (see chapter 3), then:

$$\theta_\varepsilon = 0.55°$$

$$\theta_0 = 0.7°$$

For a total Canada coverage from East to West, a span of 7.5° is needed. The number of beams required is therefore:

$$\frac{7.5°}{0.55°} = 14.$$

APPENDIX C

Part 1. Use of Speech Compression in FDM-FM-FDMA and Single-Channel-per-Carrier Systems

RMS Noise Effects

Compandors

The orthodox compandor suppresses idle noise by 30 dB, provided the noise level at the input to the expandor is -22 dBm or less[2]. The maximum signal is taken to be 0 dBm0. For a compandor with pivot point at +5 dBm0 and a compression ratio of 2:1, 0 dBm0 becomes + 5/2 or + 2.5 dBm0 after the compressor. Hence the worst signal-to-noise ratio at the input to the expandor for full noise expansion consistent with undistorted peak signal is:

$$(S/N)_c = 2.5 -(-22) = 24.5 \text{ dB.}$$

The resulting peak signal to idle noise ratio is 24.5 + 30 - 5 = 49.5 dB, allowing 5 dB degradation for subjective effects[2].

If peak signal is set at test tone level (0 dBm0) the resulting test tone/noise ratio is 49.5 dB. Hence 88 - 49.5 = 38.5 dBrnc0 is the highest noise which can be permitted if the full 25 dB advantage possible with companding is desired.

Pre-emphasis

In order to compare systems using pre-emphasis with those that do not, it is necessary to make the total power at the modulator input the same. Table C-1 shows how the pre-emphasis improvement is calculated. A speech spectrum s(f) as described in reference 2 (page 2676) was assumed, and a pre-emphasis network of the form

$$|E(f)|^2 = k_e\left(1 + \left(\frac{f}{400}\right)^2\right) \times \left(1 + \left(\frac{f}{1730}\right)^2\right) \text{ was used, where}$$

f is in Hz

k_e is a constant (9.0 dB)

If the total power is held constant,

$$\int_{300}^{3400} s(f)df = \int_{300}^{3400} s(f)\ |E(f)|^2\ df$$

and the noise reduction is equal to

$$10 \log \frac{\int_{300}^{3400} f^2 \, df}{\int_{300}^{3400} f^2/|E(f)|^2 \, df} = 8.6 \text{ dB}$$

which is the pre-emphasis advantage. The signal/noise ratio at the output of the demodulator $(\frac{S}{N})_{FM} = (S/N) - 8.6$

$$= 24.5 - 8.6 \text{ or } \approx 16 \text{ dB}$$

This figure is used in the FM equation

$$(S/N)_{FM} = 3/2 \, M^2 (\frac{C}{N_o}) \cdot \frac{1}{fm}$$

where M is the modulation index

$(\frac{C}{N_o})$ is the carrier/noise ratio

fm is a top modulation frequency

Lincompex

As can be seen from Figures C-1 and C-2 a lincompex system derived from an adaptive delta modulator equipped with spectral matching in the feedback network will have the effect of amplifying the higher frequencies which in turn will lead to higher levels in the control channel.

The output level in the signal channel in the lincompex at the transmitter end is flat with respect to frequency and the level is given in Figure C-2 as a function of pre-emphasis input level. This is also constant over the dynamic range of speech levels (0-50 dB). This signal is transmitted over FDM multiplex FM or single carrier PM, which both introduce flat noise into the signal channel. This line noise will undergo a de-emphasis and level adjustment in order that the original voice signal can be reproduced.

Conventional design objectives expressed in form of dBrnc0 (or pWp) deal with the noise under no-signal conditions (idle noise). The C-message weighting is used to find the subjective effects upon the listener. The procedure followed and reflected in the tables is to take the flat noise generated by the front end of the satellite and/or earth station (Table C-2) and send it through the de-emphasis and the C-message weighting networks. As shown, the de-emphasis improvement effect is 39/1.311 = 29.7 or 14.7 dB. C-message weighting gives 2.5 dB further reduction, consequently the total improvement is 14.7 - 2.5 = 12.2 dB. Use of pre-emphasis networks for analog and/or syllabic compandored systems will increase the rms level by 9 dB, which therefore has to be compensated for by reducing the level 9 dB at the transmitter end in order to avoid overload. However, for the lincompex case the level is transmitted in the control channel and therefore the signal channel level remains nearly unchanged due to the level compression.

The idle noise level generated by the line in a lincompex system is calculated as follows:

Peak voice signal with and without compression is set equal to test tone level (TT)[3].

The test tone to noise level[4] is set equal to	TT/N =	15 dB
Hence line noise at output of demodulator is (flat)	0-TT/N =	-15 dBm0
Expansion equal to (maximum, no signal)	=	40 dB
		-55 dBm0
Effects of de-emphasis		12.2
Output level of lincompex		-67.2 dBm0

or in dBrnc0 = 88 - 67.2 = 20.8 dBrnc0

The performance must therefore be rated as excellent.

Now the idle noise objectives for lincompex are rather artificial, because the output at the receiver end can readily be terminated during idle speech conditions, since the necessary information is available in the control channel. Whenever voice activation is employed (as in the single-channel-per-carrier case) this termination is essential. Also in order to make the introduction of the high performance lincompex-equipped channels into the present network acceptable it would be necessary to increase the idle noise above the level of the line derived noise, hence termination during idle speech periods will be necessary. The present method of matching performance cannot be used, because the voice signal will drown in noise during active talker conditions. However, the well known "cliff effects" so prevalent for digital transmission are not observed. Useful performance for a TT/N as low as 6 dB has been reported[4].

Peak Noise Level Effects

Compression

Since the expandor in the lincompex system is under the control of the control channel and always inserts a loss, "clicks" generated in the demodulator will reach the listener at a level nearly equal to the signal level present at the time of the "click". Consequently the annoyance will be far less than in PCM systems, where bit errors in the most significant digits translate into high level signals. In addition, during 60 percent of the time (idle periods) the channel will be terminated.

De-emphasis

For single-channel-per-carrier units, where PM modulation is used, the "click" noise is highest at the low end of the spectrum and inversely proportional to the frequency squared. The noise is further attenuated by the de-emphasis built into the lincompex receiver unit, as shown in Table C-3. The C-message weighting also discriminates against low frequency noise. The combined effect is a 14.5 dB subjective reduction in noise, which is effectively the same reduction as the system provides against rms noise.

According to reference 5, the power spectral density is equal to $N_t(f) = 2\mu\ [1/f^2]$ where μ is the number of clicks per second and f is frequency in Hz. Setting that noise contribution equal on a subjective basis to the rms noise, we can find μ from:

$$TT/N = 15 \text{ dB};\ TT/N_o = 15 + 10 \log 3.100$$

$$= 50.3 \text{ dB}$$

and $10 \log N_t(f)$ at $f = 1000$

$$= 10 \log 2\mu - 60 = -50.3$$

$$\therefore\ 2\mu = 9.3 : \mu = 4.65$$

The C/N of 3 dB will give such conditions[6] hence the minimum $(C/N_o) \simeq 3 + 10 \log 15 \times 10^3 = 45.8$ dB. Designs were for 49.3 dB, and so some 3 dB margin will be in hand.

Table C-1 Pre-emphasis Improvement

FREQ. (Hz)	RELATIVE SPECTRUM	FILTERS	VOLUME RELATIVE	LEVEL REAL	SUM	PRE-EMP. (dB)	VOLUME RELATIVE (dB)	LEVEL REAL	SUM	TRIANG. NOISE (dB)	PRE-EMP. (dB)	NOISE (dB)	NOISE	
													REAL	SUM
100	- 4.0	-20	-24.0	0.004	0.0036	0	-24.0	0.004	0.0045	-39.5		-39.5		0.034
200	- 1.7	-10	-11.7	0.068	0.0267	1.0	-10.7	0.085	0.0405	-23.5	- 1.0	-34.5	0.035	0.075
300	- 0.3	- 3	- 3.3	0.467	0.0734	1.9	- 1.4	0.725	0.1407	-20.0	- 1.9	-21.9	0.650	0.195
400	0		0	1.000	0.1710	3.2	+ 3.2	2.090	0.4660	-17.5	- 3.2	-20.7	0.850	0.227
600	- 1.5		- 1.5	0.710	0.1026	5.6	+ 4.1	2.570	0.4470	-14.0	- 5.6	-19.6	1.100	0.232
800	- 5.0		- 5.0	0.316	0.0474	7.8	+ 2.8	1.900	0.3450	-11.5	- 7.8	-19.3	1.170	0.935
1000	- 8.0		- 8.0	0.158	0.0870	9.9	+ 1.9	1.550	1.2850	- 9.5	- 9.9	-19.4	1.150	0.580
2000	-17.8		-17.8	0.016	0.0100	17.9	+ 0.1	1.020	1.0200	- 3.5	-17.9	-21.4	0.720	0.235
3000	-23.5		-23.5	0.004	0.0020	23.6	+ 0.1	1.020	0.5250	0	-23.6	-23.6	0.440	
4000	-27.7	-15	-42.0			28.0	-15.0	0.032		2.5	-28.0	-35.5	0.030	
					0.5437				4.2737					2.513

Sum = 1.45

$$\text{Noise reduction} = 20 - 10 \log \frac{2.513}{1.45} = 17.6 \text{ dB}$$

$$\text{Signal change} \quad 10 \log \frac{4.274}{0.544} = 9.0 \text{ dB}$$

$$\text{Improvement} = \underline{\underline{8.6 \text{ dB}}}$$

Table C-2 Flat Noise

FREQUENCY (Hz)	RELATIVE NOISE POWER/Hz dB	FACTOR	Σ	DE-EMPHASIS (dB)	C-MESSAGE WEIGHTING (dB)	RESULTANT (dB)	RESULTANT (FACTOR)	TOTAL
100	0	1		0	-42.0	-42.0	0.00002	
200	0	1	1	-1.0	-25.0	-26.0	0.00300	0.0015
300	0	1	1	-1.9	-17.0	-19.9	0.01000	0.0065
400	0	1	1	-3.2	-11.0	-14.2	0.03800	0.0240
600	0	1	2	-5.6	- 4.0	- 9.6	0.11000	0.1480
800	0	1	2	-7.8	- 1.0	- 8.8	0.13200	0.2720
1000	0	1	2	-9.9	0	- 9.9	0.10600	0.2340
2000	0	1	10	-17.9	- 1.0	-18.9	0.01300	0.5750
3000	0	1	10	-23.6	- 2.5	-26.1	0.00300	0.0800
4000	0	1	10	-28.0	-15.0	-45.0		
			39					1.3110

Σ = Factor x (No. of 100 Hz bands)

Noise reduction = 10 log $\left(\frac{39}{1.3110}\right)$ = 14.7 dB

Table C-3 Click Noise

FREQUENCY (Hz)	RELATIVE NOISE POWER/Hz			DE-EMPHASIS (dB)	C-MESSAGE WEIGHTING (dB)	RESULTANT (dB)	RESULTANT (FACTOR)	TOTAL
	dB	FACTOR	Σ					
100	20.0	100			-42.0	-22.0	0.006	0.003
200	14.0	25	50	- 1.0	-25.0	-12.0	0.063	0.035
300	10.4	11.2	16.6	- 1.9	-17.0	- 8.5	0.141	0.102
400	8.0	6.3	8.3	- 3.2	-11.0	- 6.2	0.240	0.190
600	4.4	2.8	8.3	- 5.6	- 4.0	- 5.2	0.300	0.540
800	2.0	1.6	4.15	- 7.8	- 1.0	- 6.8	0.210	0.510
1000	0	1.0	2.5	- 9.9	0	- 9.9	0.103	1.565
2000	-6.0	0.25	5.0	-17.9	- 1.0	-24.9	0.003	0.530
3000	-9.6	0.11	1.7	-23.6	- 2.5	-35.7		0.010
4000	-12.0	0.06	0.85	-28.0	-15.0			
			97.1					3.475

Σ = factor x (No. of 100 Hz bands)

Noise Reduction = $10 \log \left(\frac{97.1}{3.475}\right) = 14.5$ dB

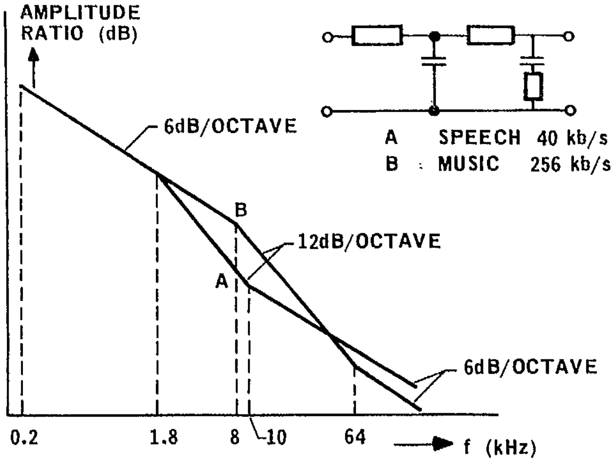

Figure C-1

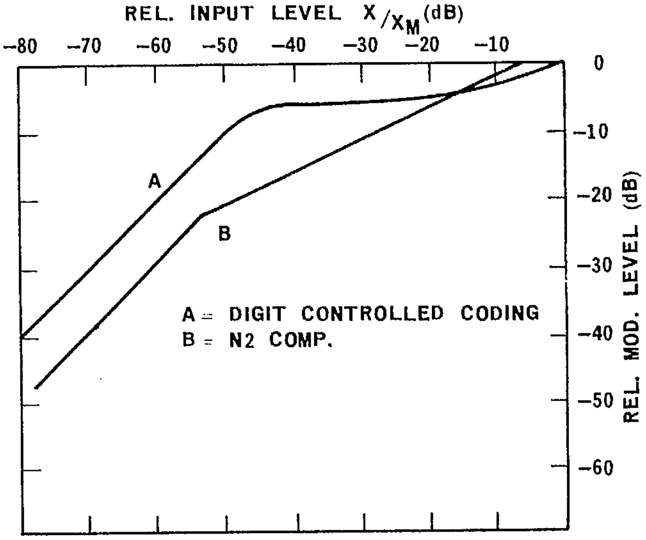

Figure C-2

APPENDIX C

Part 2. Interference Effects in FDM-FM-FDMA and Single-Channel-per-Carrier FM Systems

The FDM-FM co-channel interference from a very wide deviation signal in a homogenous system is given[1] by:

$$\frac{S_c}{I_c} = D_s^3 \times \sqrt{4\pi} \times \frac{C}{I} \times PE \qquad (1)$$

where S_c is speech level in a telephone channel relative to test tone level

I_c is tolerable interference noise level in a telephone channel relative to test tone level

D_s is RMS frequency deviation ratio

$\frac{C}{I}$ is carrier to interference ratio (unmodulated)

PE is pre-emphasis advantage

For conventional FDM, the noise allocated to interference is 35 dBrnc0 (as stated in chapter 5). It should be noted that in this case the interference is coming from the same system and is therefore under the control of the system designer.

Expressing S_c and I_c in dBm0,

$S_c = -15$ dBm0

$I_c = -88 + 35 = -53$ dBm0.

Now $D_s = M/\sqrt{10}$

where M is the peak deviation. This is shown in chapter 5 to be 6.3, hence $D_s = 2$.

Substituting in equation (1) and rearranging,

$$\frac{C}{I} = -5 \log 4\pi - 30 \log D_s + S_c - I_c - PE \text{ (in dB)}$$

$$= -5.5 - 9 - 15 + 53 - 4 = 19.5 \text{ dB}$$

For lincompex-equipped FDM-FM-FDMA systems using small deviations,

$$\frac{S_c}{I_c} = \frac{C}{I} \qquad (2)$$

The average speech and interference noise levels (as discussed in chapter 5) are:

$S_c = -6$ dBm0

$I_c = -18$ dBm0

hence

$\frac{C}{I} = 12$ dB

Since the system is designed for $\frac{C}{I}$ of at least 18 dB, there is therefore ample interference margin in a lincompex-equipped FDM-FM-FDMA system.

In the case of single-channel-per-carrier systems low deviation is used ($M \simeq 0.5$) with 15 kHz carrier spacing. The peak carrier deviation will be 2 × 3.4 × 0.5 = 3.4 kHz. By off-setting the carriers in adjacent beams by half the carrier spacing, the interference effects can be reduced to negligible proportions. It should be noted that this device cannot be used in the FDM-FM systems because of the relatively large bandwidths used.

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